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**of the 6th International Conference
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(SEIA' 2020)**

&

**the 2nd IFSA Frequency & Time Conference
(IFTC' 2020)**

Edited by Sergey Y. Yurish



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on Sensors and Electronic Instrumentation Advances**

and

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Conference**

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Edited by Sergey Y. Yurish

Sergey Y. Yurish, *Editor*
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Foreword

On behalf of the SEIA' 2020 and IFTC' 2020 Organizing Committees, we introduce with pleasure these proceedings devoted to contributions from the 6th International Conference on Sensors and Electronic Instrumentation Advances 2020 and 2nd IFSA Frequency & Time Conference held in Porto, Portugal. The conference is organized by the International Frequency Sensor Association (IFSA) - one of the major professional, non-profit association serving for sensor industry and academy more than 20 years in technical cooperation with IFSA Publishing, S.L., web portal 'My Dear Drone' and our media partners – journals '*Soft Measurements and Computing*'; MDPI *Sensors*, and *Sensors and Materials*. The conference program provides an opportunity for researchers interested in various applications of sensing and measurement to discuss their latest results and exchange ideas on the new trends. The main objective of the conference umbrella is to encourage discussion on a broad range of related topics and to stimulate new collaborations among the participants.

Extending the tradition that began in 2015 in Dubai, UAE, this conference umbrella attracts researchers and practitioners from around the world including 3 keynote speakers from a distinguished researchers of industry and academia from Italy, Portugal and USA, who were invited to overview the progress in selected research trends. This year, we had more than 70 submissions from 21 countries (12 European and 9 non-European countries), from which 50 abstracts (43 oral and 7 posters) were selected for presentation at the Conference covering theory, design, device technology, and applications of sensors and sensing systems.

The proceedings contains all papers of both: oral and poster presentations (in-person and virtual) and keynote presentations. We hope that these proceedings will give readers an excellent overview of important and diversity topics discussed at the conference.

We thank all authors for submitting their latest work, thus contributing to the excellent technical contents of the Conference. Especially, we would like to thank the individuals and organizations that worked together diligently to make this Conference a success, and to the members of the International Program Committee for the thorough and careful review of the papers. It is important to point out that the great majority of the efforts in organizing the technical program of the Conference came from volunteers.

Prof., Dr. Sergey Y. Yurish

Dr. Amin Daneshmand Malayeri

Backscattering by Closely Spaced Scatterers Using the K Matrix Data from an Active Array of N Transceivers

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Summary: A coincident array of N transceivers is used to estimate the location of a multitude of closely spaced scatterers under arbitrary wave propagation conditions. Multiple Signal Classification (MUSIC) algorithm analyses a multitude of scatterers placed in specific geometries. The formulation of the inverse scattering problem and the multistatic data matrix K are defined in two approximations: the Foldy-Lax (FL) formulation of the full multiple scattering model and the distorted-wave Born approximation (DWBA) model. The Fréchet distances (FD) between the amplitude and phase curves derived from K matrix data and of the amplitude of the scattered signals estimates the effectiveness of the approximation methods. The numerical results showed a slight effectiveness of the Foldy-Lax approximation for scatterers location. The problem of considering the phase estimation from the K matrix is not a solution for signal reconstruction and representation.

Keywords: Inverse scattering, Foldy-Lax formulation, Distorted-wave Born approximation, MUSIC, Fréchet distance, Average visibility index.

1. Introduction

The Multiple Signal Classification (MUSIC) is a method applied to locate a number of small targets with arbitrary locations. MUSIC operates based on the assumption that the far field patterns of point sources are almost orthogonal to the noise subspace of the far field operator or so called multistatic matrix data K. K is obtained for a collection of M targets and an active array of N transceivers [1-6]. Gruber et al. [1] established that the MUSIC is able to locate point targets using both the Born approximation and more general scattering models that consider multiple scattering processes between the targets. The MUSIC algorithm takes in as input a K matrix which is approximated using two approximations: the Foldy-Lax (FL) formulation of the multiple scattering model [7-10] and the distorted-wave Born approximation model (DWBA) [11, 12]. The FL considers the multiple scattering among the scatterers but the self-interactions at the target locations are excluded. DWBA ignores this multiple scattering among the scatterers effect.

Therefore, the aims of this paper are to evaluate the accuracy of target location using the values of phase and amplitude provided by the K matrix. The issue of interest is to have either closely spaced targets and/or sparse transceiver arrays. The targets are placed in a background medium with known properties. The formulation of the inverse scattering problem is defined under the FL and DWBA approximation methods. More specific, we formulate the K matrix in

these two approximations and the MUSIC algorithm takes in as input these K matrices. Neither approximation intends to recover the type or properties of these scatterers. In our case, we are focusing on the data collected at 5, 10, 50 and 100 Hz (or normalized wavelengths of 1, 0.5, 0.1, and 0.05) in order to develop a new scanning technology applicable in a nondivest and noncontact condition. Additive white Gaussian noise (AWGN) is used to investigate the noise immunity and the noise-free case is presented for comparison purpose. A new aspect is addressed here: the investigation of the amplitude and phase distribution of the singular vectors obtained from the SVD of the matrix K in both approximations through the Fréchet distance measurement in order to quantify the distance between the amplitude and phase curves vs. the number of transceivers in both approximations [13-15]. The proper scatterer detection is provided by those target geometries and set of parameters, which minimize the Fréchet distance.

This problem has a practical implementation in sonar systems, medicine and ultrasound imaging, non-destructive testing, and smart fighting methods against border smuggling through hidden object detection, as this paper intended to investigate.

The rest of this paper is organized as follows: Section 2 is devoted to the problem statement (a theoretical framework for MUSIC and approximation models and the selection algorithm). Section 3 presents the results, which are discussed in Section 4. Finally, conclusions are presented in Section 5.

2. Materials and Methods

2.1. Inverse Scattering Problem Formulation

2.1.1. Foldy-Lax Formulation

The sum of the scattered field and the incident field for the k^{th} transceiver build the total field:

$$u_k(\mathbf{r}) = u_k^{\text{inc}}(\mathbf{r}) + \sum_{m=1}^M \tau_m G_0(\mathbf{r}, \mathbf{x}_m) u_k(\mathbf{x}_m), \quad (1)$$

where $\tau_m > 0$ denotes the target scattering strength for an m^{th} point scatterer, and G_0 is the background Green function that is defined as, for a scatterer locations $\mathbf{r} = \mathbf{x}_m$:

$$G_0(\mathbf{r}, \mathbf{r}') = \frac{e^{ik|\mathbf{r}-\mathbf{r}'|}}{4\pi|\mathbf{r}-\mathbf{r}'|} \quad (2)$$

In the case of a collection of M point scatterers located at $\mathbf{x}_m$, $m = 1, 2, \dots, M$, the FL model for the scattered wave field when the self-interactions at the target locations are excluded, is [7, 16-18]

$$u_k(\mathbf{x}_m) = u_k^{\text{inc}}(\mathbf{x}_m) + \sum_{m' \neq m} \tau_{m'} G_0(\mathbf{x}_m - \mathbf{x}_{m'}) u_k(\mathbf{x}_{m'}) \quad (3)$$

The FL model hypothesis that the data on the incident field u_k^{inc} , scatterer locations $\mathbf{x}_m$, and target scattering strength τ_m are known. A coincident array meets the condition $j = k = N$, and (1) and (3) became

$$G(\mathbf{x}_m, \alpha_k) = G_0(\mathbf{x}_m, \alpha_k) + \sum_{m' \neq m} \tau_{m'} G_0(\mathbf{x}_m, \mathbf{x}_{m'}) G(\mathbf{x}_{m'}, \alpha_k), \quad (4)$$

where $G(\mathbf{x}_m, \alpha_k)$ is a complete Green function of the combined background with the addition of the target medium for a source point $\mathbf{r}' = \alpha_k$ at the k^{th} emitter and, similarly, $G(\alpha_j, \mathbf{x}_m)$ for the j^{th} receiver location. The K matrix in the FL approximation is:

$$\begin{aligned} K^{\text{FL}} &= \\ &= \sum_{m=1}^M \tau_m G_0(\alpha_j, \mathbf{x}_m) G(\mathbf{x}_m, \alpha_k) = \\ &= \sum_{m=1}^M \tau_m G_0(\mathbf{x}_m, \alpha_j) G(\mathbf{x}_m, \alpha_k), \end{aligned} \quad (5)$$

$j = 1, \dots, M; k = 1, \dots, N$

The matrix formulation (5) is a correction of Green's function accounting for the scattering of considered targets.

2.1.2. Distorted-wave Born Approximation (DWBA) Model

The DWBA model is a tool for weak scatterers analysis [11]. When the scattering targets have

acoustic properties different from the background medium, then the DWBA model is as follows:

$$G(\mathbf{x}_m, \alpha_j) \approx G_0(\mathbf{x}_m, \alpha_j) \quad (6)$$

The multistatic scattering matrix within the DWBA approximation is

$$K^{\text{DWBA}} = \sum_{m=1}^M \tau_m G_0(\mathbf{x}_m, \alpha_j) G_0^T(\mathbf{x}_m, \alpha_k) \quad (7)$$

DWBA estimates the target locations from the multistatic scattering matrix when the condition $M < N$ is met. In addition, this is a requirement related to well-separated scatterers, so they are 'viewed' by the transceiver array. However, DWBA is only relevant at small incident angles and at small exit/reflection angles.

2.2. MUSIC Algorithm

The MUSIC algorithm uses the SVD of K matrix and both discrete version of the multistatic scattering matrix (5) and (7) to define the wave scattering amplitude [19]. For M point scatterers located at $\mathbf{x}_m = \{x_1, \dots, x_M\} \in \mathbb{R}^3$ and when $N \geq M$, two matrices S_{jm} , $j = 1, \dots, N$ and $m = 1, \dots, M$ consisting of the background Green function vectors computed at the scatterer locations and $\mathbf{T} = \text{diag}(\tau_m)$ are defined so that the multistatic matrix is $K = \mathbf{S} \mathbf{T} \mathbf{S}^*$. $\mathbf{S}^*$ denotes the adjoint of $\mathbf{S}$. MUSIC characterizes the range of a self-adjoint operator. When nonstationary signals are analyzed, a better approach is singular value decomposition. SVD reduces a matrix to its constituent parts and ensures an orthogonal basis for the signal space of the matrix. The eigenvectors of K provide information on the scatterers, and SVD determines the number of inclusions in the propagation space. K can be transformed into a Hermitian matrix using the frequency-domain version of a time-reversed multistatic response matrix. Further, the eigenvectors and corresponding eigenvalues of the Hermitian multistatic scattering matrix are computed using SVD. SVD is more stable than the standard eigenvalue method [19]. These statements are essential observations of the MUSIC approach. This means that one can estimate the locations of the scatterers associated with the received signals by finding those steering vectors $\mathbf{f}(\mathbf{x})$, which are most nearly orthogonal in the null (noise) space.

2.3. Fréchet Distance

The Fréchet distance (FD) is a measure of the similarity between two parametrized curves X and Y defined in $\mathbb{R}^n$ [13-15]:

$$\begin{aligned} \text{FD}(X, Y) &= \\ &= \min_{\substack{a: [0,1] \rightarrow [0,N] \\ b: [0,1] \rightarrow [0,M]}} \left\{ \max_{t \in [0,1]} d(X(a(t)), Y(b(t))) \right\}, \end{aligned} \quad (8)$$

where $a(t)$ and $b(t)$ are parametrized functions with $a(0) = 0$, $a(1) = N$, $b(0) = 0$, and $b(1) = M$. $d(\cdot, \cdot)$ is a common distance metric such as the Hausdorff or Euclidian distance. We must mention that these common distance metrics are not suitable for estimating the similarity between two curves because they simply measure the distance between points in a set of points but fail to consider the structure of the curves and the point correlations in the same set. FD indicates matches between the analyzed curves by computing the distance between a point traveling along the X curve and a point traveling along the Y curve. Usually, each point can travel at a different speed on each curve.

The FD between the amplitude and phase curves plotted using the data provided by the singular vectors obtained from the SVD of the matrix K in the FL and DWBA approximations is determined. The best target locations are provided by the smaller values of FD.

2.4. Proposed Algorithm

The target location problem is investigated for four spatial distribution geometries. The MUSIC algorithm for both approximations provides reconstruction methods that depend on the choice of the signal and noise subspaces of the multistatic scattering matrix. For small measurement noise n_1 or for higher SNR it is easy to choose these subspaces and to detect and locate the scatterers in a finer way. A small amount of noise depicts almost ideal conditions whereas a large amount of noise n_2 will strongly disturb the detection of the scatterers. The algorithm is as follows:

Step 1: Define the set of parameters as $p:\{\lambda, n_1, n_2\}$, where n_1 and n_2 denote AWGNs having a variance of 0.0009 and 0.09, respectively; Define the results as $r:\{FD_{amplitude}^{FL}, FD_{amplitude}^{DWBA}, FD_{phase}^{FL}, FD_{phase}^{DWBA}\}$, where $\lambda = (1, 0.5, 0.1, 0.01)$ and $P = (\text{noise-free}, n_1, n_2)$ (see Table 1);

Step 2: Generate all possible combinations, the experimental parameters and results are distributed into six lists: $L1 = \{\lambda, P, FD_{amplitude}^{FL}\}$, $L2 = \{\lambda, P, FD_{amplitude}^{DWBA}\}$, $L3 = \{\lambda, P, FD_{phase}^{FL}\}$, $L4 = \{\lambda, P, FD_{phase}^{DWBA}\}$;

Step 3: Extract the smallest FD values from the first four lists L_i ($i = 1:4$) and generate the FD list $\{FD_{amplitude}^{FL}, FD_{phase}^{FL}, FD_{amplitude}^{DWBA}, FD_{phase}^{DWBA}\}$ (see data in Tables 2 and 3);

Step 4: Run steps 2 and 3 for 16 times for each target distribution; 64 measurements are made.

3. Results

3.1. Simulation Framework

The simulation approach was implemented in the MATLAB 2017b environment [20]. The following parameters were used:

$N = 10$ transceivers in a coincident linear array located at $(-20,0)$, $(-15,0)$, $(-10,0)$, $(-5,0)$, and $(-2,0)$. The targets are located as follows: triangle (T): $(0,-8)$, $(-1,-9)$, $(1,-9)$, $(-2,-10)$, $(0,-10)$, and $(2,-10)$; parallelogram (P): $(-2,-8)$, $(0,-8)$, $(2,-8)$, $(0,-9)$, $(2,-9)$, and $(4,-9)$; diamond (D): $(0,-6)$, $(-2,-8)$, $(0,-10)$, $(2,-8)$, $(0,-7)$, and $(0,-9)$; and ellipse (E): $(0,-8)$, $(-3,-9)$, $(3,-9)$, $(-3,-10)$, $(3,-10)$, and $(0,-11)$;

$M = 6$ closely spaced scatterers having a spatial distribution into triangle (T), parallelogram (P), diamond (D), and ellipse (E) geometries. The target scattering strength is $\tau = (1, 1.3, 1.6, 1.6, 1.3, 1)$;

AWGN with variances of 0.0009 and 0.09 is added; noise disturbs the multistatic scattering matrices K^{FL} and K^{DWBA} and results in various signal-to-noise (SNR) values. In addition, the noise-free case is considered.

These parameters maintain the computational time within reasonable limits.

A block diagram of the proposed method is displayed in Fig. 1.

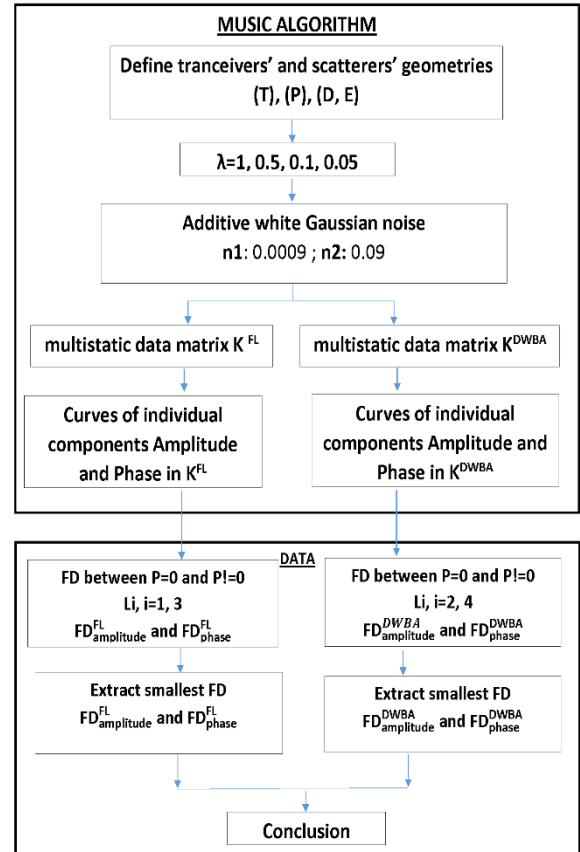


Fig. 1. Block diagram.

3.2. Results

Fig. 2 displays plots transceiver and target locations. Both the phase and amplitude distribution of the singular vectors obtained from SVD of matrix K in FL and DWBA methods are presented in Fig. 3.

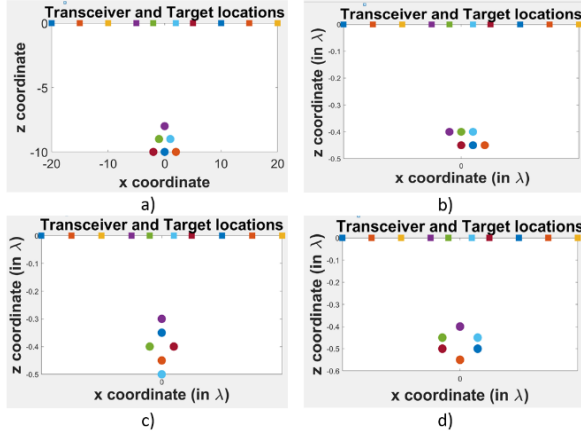


Fig. 2. Problem setup: 10 transceivers (depicted as squares) and 6 targets (depicted as circles) are embedded in uniform background. Targets' spatial distributions are represented as a) triangle (T), b) parallelogram (P), c) diamond (D), and d) ellipse (E).

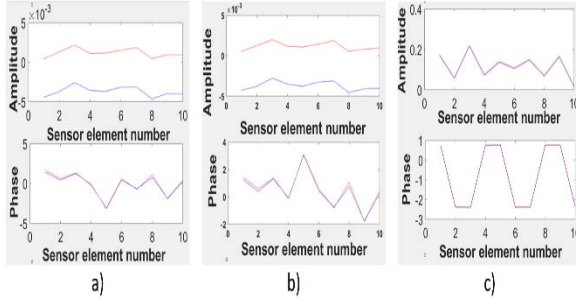


Fig. 3. Curves of 'amplitude' (top) and 'phase' (bottom) provided by the left singular vectors obtained from SVD of the matrix K in FL (blue) and DWBA (red) vs. number $N = 10$ of array elements in triangle (T) geometry and for $\lambda = 1$. a) Noise-free; b) n_1 ; and c) n_2 .

FD estimates the errors of the approximate solutions of the data provided by the singular vectors obtained from SVD of the matrix K, and allows the establishment of an efficient geometry of the target localization. According to the data presented in Table 1, the smallest $FD_{\text{amplitude}}$ values indicate an improvement in the FL approximation. As expected, the case of noise-free scattering returns the same results. By contrast, the noisy experiment indicates that DWBA provided somewhat worse results in terms of the FDs and amplitude values of the reconstructed signals. The exception is the ellipse spatial distribution for noise n_2 , where almost similar results were obtained.

The $FD_{\text{amplitude}}$ for both approximations (Table 1) shows a variability with the wavelength and spatial distribution of the targets. Diamond (D) and ellipse (E) distributions have smaller $FD_{\text{amplitude}}$ values when a comparison between approximations is performed (Table 2). The FL approximation shows a fair robustness against noise, while DWBA drives an unstable reconstruction process. On the other hand, FD_{phase} remains almost unchanged with no contribution to the optimization problem regardless of the targets' spatial distributions (Tables 1 and 2).

Table 1. Average $FD_{\text{amplitude}}$ and FD_{phase} values of scattered signals attained in simulation experiments for 2D obstacles, for FL and DWBA approximations.

Geometry	AWGN noise	SNR (dB)	$FD_{\text{amplitude}}^{\text{FL}}$	$FD_{\text{amplitude}}^{\text{DWBA}}$	$FD_{\text{phase}}^{\text{FL}}$	$FD_{\text{phase}}^{\text{DWBA}}$
(T)	n_1	22.38	$0.18 \cdot 10^{-3}$	0.0002	3.035	3.1249
	n_2	6.3	0.2168	0.2183	2.946	3.0001
(P)	n_1	22.66	$0.21 \cdot 10^{-3}$	0.0103	0.278	0.2875
	n_2	8.9	0.2222	0.2225	2.216	2.2747
(D)	n_1	22.12	$0.18 \cdot 10^{-3}$	0.0003	0.157	0.2192
	n_2	3.3	0.2210	0.2212	3.701	3.7075
(E)	n_1	21.62	$0.13 \cdot 10^{-3}$	0.0001	1.896	0.2143
	n_2	9.32	0.22068	0.2205	2.001	2.0671

Table 2. Ratio of AWGN noise-added (n_1 and n_2) to noise-free values for smallest $FD_{\text{amplitude}}^{\text{FL}}$ and $FD_{\text{amplitude}}^{\text{DWBA}}$ values across various spatial distributions and wavelengths.

Noise-free vs. n_1					
λ	Geometry	$FD_{\text{amplitude}}^{\text{FL}}$	λ	Geometry	$FD_{\text{amplitude}}^{\text{DWBA}}$
0.5	(P)	$0.20 \cdot 10^{-3}$	1	(D)	$0.94 \cdot 10^{-3}$
0.5	(T)	$0.23 \cdot 10^{-3}$	0.5	(E)	$0.33 \cdot 10^{-3}$
Noise-free vs. n_2					
1	(E)	0.020	1	(E)	0.022
1	(D)	0.042	1	(D)	0.062

Table 3. Ratio of AWGN noise-added (n_1 and n_2) to noise-free values for $FD_{\text{phase}}^{\text{FL}}$ and $FD_{\text{phase}}^{\text{DWBA}}$ smallest values across various spatial distributions and wavelengths.

Noise-free vs. n_1					
λ	Geometry	$FD_{\text{amplitude}}^{\text{FL}}$	λ	Geometry	$FD_{\text{amplitude}}^{\text{DWBA}}$
0.5	(T)	0.010	1	(D)	0.011
0.5	(D)	0.021	0.5	(E)	0.024
Noise-free vs. n_2					
1	(D)	1.143	1	(E)	2.958
1	(P)	2.015	1	(D)	2.106

4. Discussion

The formulation of the inverse scattering problem is defined under FL and DWBA approximation methods. The K matrix is generated via these approximations. SVD identifies the right singular and the left singular vectors associated to the singular value decomposition of K. The plots the amplitude and phase distribution of the singular vectors are analyzed using the Fréchet distance. Simulated data shows that the

Fréchet distance values are affected by the target locations and their spatial distributions, normalized wavelengths, and noise (Fig. 3 and Table 1). Noise disturbs the multistatic scattering matrices K^{FL} and K^{DWBA} and acoustic field distribution.

Data displayed in Fig. 3 and Tables 1-3 indicates that the 'phase' provided by the singular vectors obtained from SVD of the matrix K is not dramatically affected by the location of the targets. The 'amplitude' in matrix K is more affected by changes in the target location. In addition, the distribution of the 'amplitude' FD and 'phase' FD components in K (Table 1) shows variability for each wavelength, although there is no FD-wavelength relationship. Smaller FD values indicate a higher similarity between curves and allows for estimating the validity of the approximation model. Moreover, for each target's geometry, the SNR values are almost similar for the same experimental conditions. This indicates that it is quite difficult and expensive to obtain the 'phase' and 'amplitude' data in the SDV singular vectors of the K matrix, under these experimental conditions.

4. Conclusions

This work investigated questions related to a new scanning technology applicable in a nondivest and noncontact condition based on the inverse scattering of point scatterers embedded in a known background environment. We are currently working on a low-frequency interrogation technology for a possible contactless pat down system based on the automatic detection of external concealed commodities viewed as scatterers or targets.

The inverse scattering problem was approached using two well-known approximations: FL formulation of the multiple scattering model and the DWBA model. The K matrix was generated using these approximations and has two different mathematical forms depending on the used approximation. The singular vectors obtained from the SVD of the matrix K in FL and DWBA approximations and the Fréchet distance were used to estimate the similarity between the 'amplitude' and 'phase' curves of the K matrix. The best target location was determined by minimizing the values of FD.

The numerical results showed a slight effectiveness of the Foldy-Lax approximation for scatterers location. The problem of considering the phase estimation from the K matrix is not a solution for signal reconstruction and representation. However, our results indicated that the efficiency of the target location marginally depends on the placement of the interior points and strongly depends on the wavelength and level of noise.

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Hydrogen Peroxide Vapor Sensor Based on Zinc Oxide

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Summary: In this study, we report the design and fabrication of a high-performance hydrogen peroxide vapor sensor based on La doped ZnO nanostructured film using the high-frequency magnetron sputtering method. The response (R_{gas}/R_{air}) of the sensor was measured at various operating temperature to different concentrations of hydrogen peroxide vapors. It was also studied the gas sensing properties of the sensor by impedance spectroscopy. Gas sensing tests indicate that La-doped ZnO sensor exhibit a good sensitivity to low concentration of hydrogen peroxide vapor. We expect that in the future, La doped ZnO nanostructured sensitive films will be able to be utilized in highly sensitive, real-time hydrogen peroxide vapor sensors.

Keywords: Zinc oxide, Conductometric sensor, Hydrogen peroxide vapor, Thin film, Sensitivity.

1. Introduction

Hydrogen peroxide (H_2O_2) is an odorless and colorless liquid and it is more viscous than water. It is an important metabolic product in chemical and food industries. It is used as a fuel and has chemical applications, biological function, domestic uses, as well as therapeutic use, including use as an antimicrobial and oxidizing agent. Different concentrations of hydrogen peroxide with water solution are also used for industrial applications such as pulp- and paper-bleaching. It is available to consumers most commonly with a 3 % solution as a general disinfectant [1-4].

Due to such a wide range of applications the demand for developing hydrogen peroxide vapor detection methods is increasing year by year. Research on the detection and measurement of the concentration of H_2O_2 vapors received considerable attention in the last decade. Various methods such as gas chromatography, surface plasma resonance, fluorescence, Fourier-transform infrared spectroscopy, chemiluminescence detectors, mass spectrometry, colorimetry, electrochemistry and semiconductor gas sensors have been applied in hydrogen peroxide detection. Unlike other techniques, gas sensors based on metal oxide semiconductor (MOS) nanostructured materials have been widely used, having to the advantages of low cost, easy to miniaturize and production, high stability, selectivity, sensitivity, reliability and fast response [5-8].

In the manufacturing of conductometric gas sensors several types of MOS materials have been explored, including SnO_2 , ZnO , In_2O_3 , TiO_2 , WO_3 , CuO , and Fe_2O_3 with different dopants and morphologies. Among the oxides, ZnO has some advantages such as its low cost, wide bandgap energy (3.37 eV), simple synthesis of nanostructures, high mobility of electron carriers, large exciton binding

energy (60 meV) at room temperature and non-toxicity in nature that makes it eco-friendly. Zinc oxide is one of the most studied MOS material due to its unique surface properties. Although pure ZnO has not represented high sensing performance but conductometric sensors based on doped ZnO nanostructured material exhibits high sensitivity, selectivity and stability toward both oxidizing and reducing gases [9-13].

In this study, we prepared La doped ZnO based sensor for the detection of hydrogen peroxide vapor and confirmed its high sensitivity by means of gas sensing studies.

2. Experimental

The sensitive layers were deposited by magnetron sputtering method. First, sputtering target based on ZnO doped 1 at.% La was fabricated using solid-phase reaction method. The process used to prepare the magnetron sputtering target may be found in our previous reports [14, 15]. The synthesized $ZnO<La>$ composition was subjected to mechanical processing in order to eliminate surface defects having smooth, parallel tablet with the diameter of 50 mm and the thickness of 2 mm (see Fig. 1).

2.1. Magnetron Sputtered ZnO layer

Magnetron sputtering is one of the best methods for obtaining thin films from very wide range of materials including metals, dielectrics, ceramics, polymers and so on [16].

Magnetron sputtering is a process whereby atoms of a solid target material are ejected (or vaporized) due to the momentum transfer from an atomic-sized energetic bombarding particle impinging on the target surface. Then these vaporized particles are condensed

on the substrate material. The sputtering is carried out using gaseous ions from plasma that are then accelerated and directed toward the target. The plasma used during the sputtering is created and controlled by magnetron guns [17].

The ZnO<La> sensitive layers were deposited by the VTC-600-2HD DC/RF Dual-Head High Vacuum Magnetron Plasma System (see Fig. 1). As known, plasma is weakly ionized quasi-neutral gas made up of ions and electrons to be a good electrical conductor. Generally, processes that depend on the use of the plasma are referred to as plasma processing. The plasma processing has plasma densities between 10^8 – $10^{13}/\text{cm}^3$ and during the sputtering of the ZnO<La> layer the plasma density was $10^{10}/\text{cm}^3$. The gas atoms may not be fully ionized and there are not only electrons and ions but also neutrals. The electron velocities are much higher than that of ions and neutrals. The plasma properties are heavily dependent on the excitation and ionization energies of the gas. The most suitable gas is argon (Ar) because it is an 'inert' or 'noble' gas, thus not explosive when subjected to the RF (13.56 MHz) magnetic field or spark and it has low ionization energy (15.7 eV). High purity argon (99.99%) was used as a plasma sustaining gas during the sputtering of ZnO<La> target obtaining nanostructured gas-sensitive film with enhanced parameters. The krypton gas is also a good candidate for this application, but it's expensive. Ar is relatively cheap to manufacture since it is extracted directly from the atmosphere [18].



Fig. 1. The photos of the VTC-600-2HD DC/RF Dual-Head High Vacuum Magnetron Plasma System and the ZnO<La> sputtering target.

Generally, plasma can be used for different purposes depending on the type of the PVD process. It can be used as a source of ions or as a source of electrons. For applying power sources for discharge there are two sources: DC (continuous DC, Pulsed DC) and AC (LF (~100 Hz), MF (~100 Hz - ~1 MHz) and RF (> ~1 MHz)). The RF magnetic field with 13.56 MHz was used by us as a power source. For magnetron DC and AC sputtering processes plasma is used as a source of ions for the sputtering from a target material if the plasma ions are directed toward it. The

uniformity of plasma is very important concern for more tunable deposition process environment. Plasma uniformity is mainly dependent on sputtering system geometry and the means by which the plasma was produced. The plasma produced by DC diode is not ideal for sputtering because the electrons ejected from the cathode are accelerated away from the cathode and are not efficiently used for sustaining the discharge [19].

Ceramic substrates with the interdigitated electrodes, the temperature sensor (Pt 1000) and the heater were used as sensor substrate (Multi-Sensor-Platform) (see Fig. 2). Gas sensitive ZnO<La> layers were deposited onto Multi-Sensor-Platform by the VTC-600-2HD DC/RF Dual-Head High Vacuum Magnetron Plasma System under the following conditions: 60 W input power, 150 °C deposition temperature, 70 mm distance between the target and substrate, 2×10^{-4} Pa base pressure, 6×10^{-1} Pa deposition pressure, and 20 minutes duration of sputtering. Then, the sensors were sensitized by deposition of palladium catalytic particles on the surface of the sensing layers by the ion-beam sputtering method (the deposition time was 3 seconds, the cathode current and the anode voltage were equal to 65 A and 25 V, respectively). Finally, the samples were annealed at 350 °C during 3 hours for stabilization of sensing parameters.

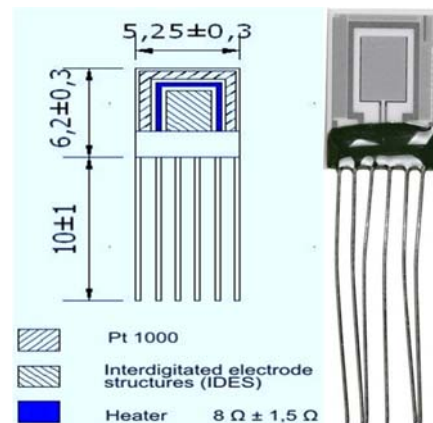


Fig. 2. The schematic diagram of the Multi-Sensor-Platform (with mm unit) and the photo of Multi-Sensor-Platform based sensor.

3. Results and Discussion

The gas sensing properties of the ZnO<La> thin layer were studied in the presence of hydrogen peroxide vapors using a home-made computer-controlled gas testing system [1]. To have the necessary concentration of H_2O_2 in the chamber, the liquid hydrogen peroxide was introduced into the chamber on the special hot plate designed for the quick conversion of the liquid H_2O_2 to the gas phase. The response of the fabricated sensors is calculated as, $R = R_g/R_a$, where R_a is the resistance of the sensor in the presence of air and R_g is the resistance in the presence of the target gas.

Initially, the ZnO<La> sensor's responses were measured at different operating temperatures in the presence of 100 ppm hydrogen peroxide vapors. Fig. 3 shows the dependence of the sensor's response on the operating temperature. The maximum response value was observed at 200 °C operating temperature, but from the temperature of 150 °C the sensor showed quite promising sensing results (at 50 and 100 °C operating temperatures the ZnO<La> sensor showed response to hydrogen peroxide vapors, but did not recover after eliminating the gas effect).

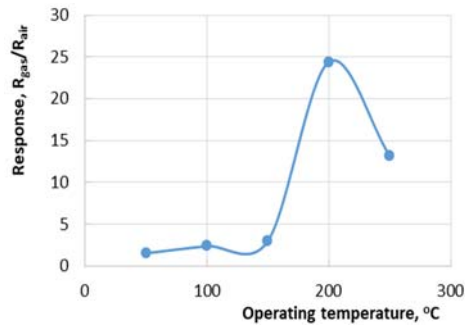


Fig. 3. The dependence of the ZnO<La> sensor's response on the operating temperature in the presence of 100 ppm hydrogen peroxide vapors.

The changes in the sensor's resistance in the presence of 100 ppm hydrogen peroxide vapors at different operating temperatures were also measured (see Fig. 4). At both temperatures (200 and 250 °C), the response times of the sensor are a few seconds, but the recovery times are longer. It is noteworthy that at 250 °C operating temperature the sensor's resistance changed about 25 times in the presence of 100 ppm hydrogen peroxide vapors.

The sensing responses of the La-doped ZnO film at 150 °C operating temperature towards ppm level of H₂O₂ were tested and the results are shown in Fig. 5. The results indicated, that the sensor exhibited clear and reversible gas responses to ppm levels of the gas. The absolute sensor responses to H₂O₂ vapors at the concentrations of 100 and 200 ppm, were about 3 and 14 respectively.

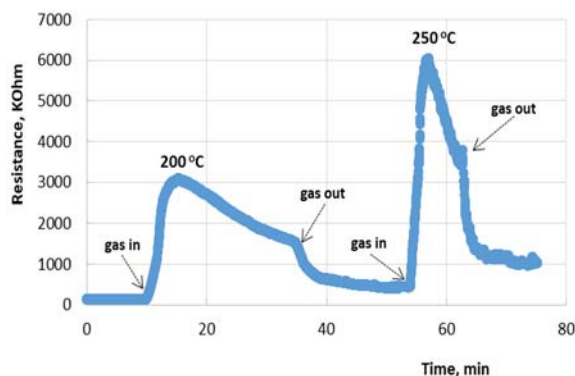


Fig. 4. The resistance variation of the ZnO<La> sensor in the presence of 100 ppm hydrogen peroxide vapors at 200 and 250 °C operating temperatures.

The gas-sensitive properties of the sensor are excellent at 200 u. 250 °C operating temperatures, but at 150 °C the power consumption of the sensor is quite low. From this point of view, it is quite difficult to choose a specific operating temperature. On the other hand, the impedance properties of the sensor are fairly promising at 150 °C.

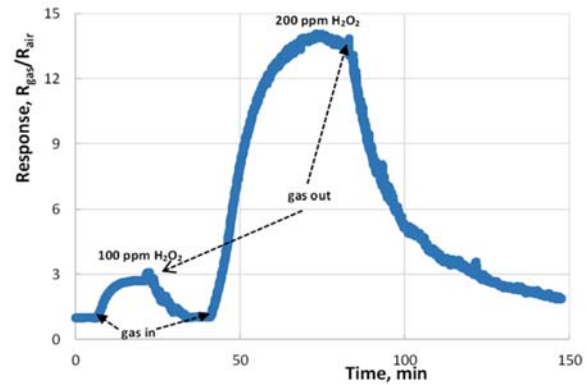


Fig. 5. The response-recovery curves of ZnO<La> sensor observed at different concentration of H₂O₂ vapors at 150 °C operating temperature.

There were also studied the gas sensing properties of the sensor by impedance spectroscopy using the ZIVE-SP1 Potentiostat and the Keithley 4200-SCS (Semiconductor Characterization System).

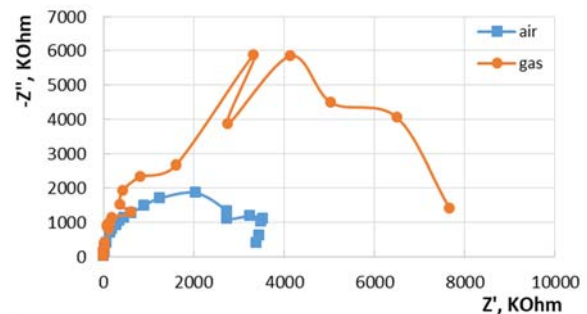


Fig. 6. The Nyquist plots of ZnO<La> sensor observed in the air and in the presence of 200 ppm H₂O₂ vapors at 150 °C operating temperature.

The Nyquist plot is commonly depicted as a semicircle at high frequency. The Nyquist plots of ZnO<La> sensor were observed in the air and in the presence of 200, 300 and 400 ppm hydrogen peroxide vapors at 150 °C operating temperature. Fig. 6 and Fig. 7 show the significant deviations of Nyquist plots at the presence of H₂O₂ vapors and in the air.

Future researches will focus not only on the study of the deviations of Nyquist plots at the presence of H₂O₂, but also on the build of the sensor equivalent circuit, obtaining more comprehensive information about the gas-sensitive element.

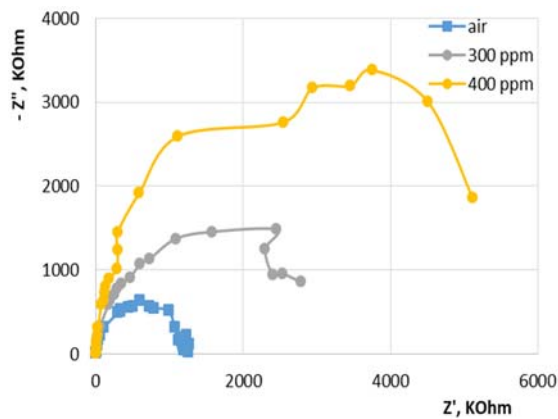


Fig. 7. The Nyquist plots of ZnO<La> sensor observed in the air and in the presence of 300 and 400 ppm H₂O₂ vapors at 150 °C operating temperature.

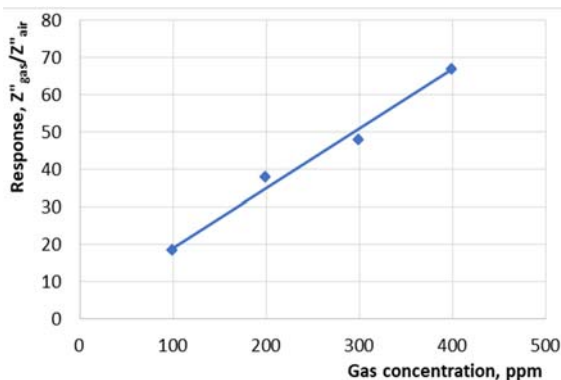


Fig. 8. The dependence of the ZnO<La> sensor's response on the H₂O₂ vapor concentration at 150 °C operating temperature.

It was extracted the response vs. concentration curve for the ZnO doped 1 at.% La film. Fig. 8 shows the dependence of response on the hydrogen peroxide vapor concentration at 150 °C operating temperature. Here, similar to the resistive response, we consider a impedance response with the following definition: $R = Z''_{\text{gas}}/Z''_{\text{air}}$ (at 460 Hz frequency). It's clear, that the dependence has a linear characteristic, which will allow not only to detect of hydrogen peroxide vapors but also to accurately measure the low concentrations of this gas.

4. Conclusions

We fabricated the Ln doped ZnO sensitive films and demonstrated hydrogen peroxide vapor sensor at low operating temperature. The excellent gas sensing performance can be attributed to the following aspects: the sensor provides not only good response to ppm level of H₂O₂ vapors at low operating temperature but also significant deviation of Nyquist plots at the presence of hydrogen peroxide vapor. Future works includes change the dopant percentage in ZnO material to enhance the best sensitivity and selectivity of the sensor.

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Proposal of a Reconfigurable Sensor for Measuring Temperature and Capacitance

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Summary: We propose the use of a modified fringing field capacitive sensor based on PCB technology to measure both capacitance and temperature. The conventional interdigitated structure of the fingers in a capacitive fringing field sensor based on PCB technology is modified to allow its reconfiguration to measure both capacitance and temperature (using the copper tracks of the PCB as an RTD sensor). A proof-of-concept prototype of the sensor was fabricated and measurements of temperature and capacitance were taken in laboratory. Due to the small copper's temperature variation, a simple signal conditioning circuits were used to measure the temperature in the reconfigurable sensor. The capacitance was measured using a commercial capacitance meter. The measured values of temperature and capacitance show that the proposed sensor can be successfully employed in applications where a low-cost temperature and moisture sensor is required.

Keywords: Reconfigurable sensor, Capacitance measurements, Temperature measurements, PCB technology, Fringing field capacitor.

1. Introduction

Capacitance fringing field sensors based on PCB technology are used for measuring moisture in several applications like cement [1], stored grains [2], soils [3], air [4] and paper pulp [5, 6]. The fabrication of a conventional fringing field PCB capacitive sensor consists of using a large number of interdigitated co-planar electrodes, as shown in Fig. 1.

If n copper tracks (called fingers) with thickness h separated by a distance d (between fingers) and with a length l are fabricated on one side of a PCB, the capacitance of the structure is given by:

$$C = \epsilon_r \epsilon_0 \frac{nlh}{d}, \quad (1)$$

where ϵ_0 and ϵ_r are, respectively, the permittivity of free space and the relative permittivity of the dielectric material between two fingers. Although a correction factor may be necessary in (1), with the usual dimensions used in the sensors ($lh \gg d^2$), this is not necessary.

If a material with a different ϵ_r (for example, water) is placed in contact with the PCB surface that contains the fingers, the value of the capacitance changes. Therefore, if only a part of the PCB surface is in contact with the water, the measured variations in the sensor's capacitance can be correlated with the amount of area covered with water.

In this paper we propose a new configuration for the interdigitated copper tracks structure. With this modification, the sensor can be used to measure both capacitance and temperature. The reconfigurable

sensor can be fabricated using standard low-cost standard PCB technology.

2. Proposed Sensor

2.1. Modified PCB Structure

Instead of the traditional fingers configuration presented in Fig. 1, the interdigitated pattern was made using two parallel long copper tracks, with a serpentine shape. To avoid short-circuits between the fingers and to eliminate the possibility of moisture absorption by the FR-4 board, a protection layer with conventional solder mask was applied to the PCB. A simplified diagram of the sensor copper tracks is presented in Fig. 2.

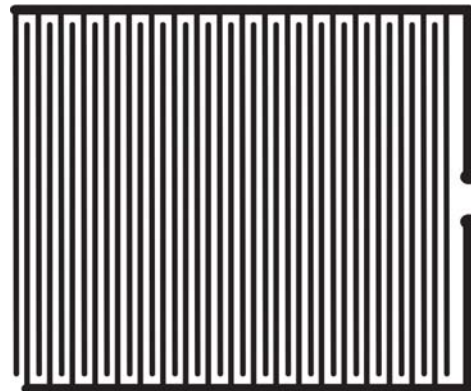


Fig. 1. Layout of a conventional PCB interdigitated capacitor with 40 fingers of copper tracks.

2.2. The Reconfigurable Sensor

The advantage of using the new configuration is that, besides the conventional capacitance measurement (that can be performed making contacts at the beginning - or end - of the two parallel tracks), if we make contacts at the beginning and at the end of any of the single continuous tracks, we have a resistor composed of a long line of copper. These points of measurement are shown in Fig. 2.

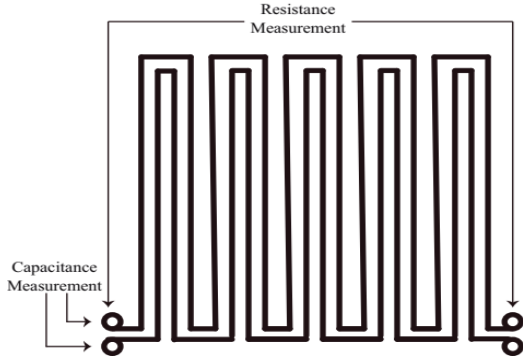


Fig. 2. Simplified layout of the proposed reconfigurable sensor.

Since the copper presents a well-known temperature coefficient (TC), we can use this copper resistor as a resistance temperature detector (RTD) [7-9]. Copper presents a low TC, but with excellent linearity. Thus, by measuring the resistance of the copper track, one can obtain a precise temperature measurement.

3. Experimental Results

A simple prototype (a proof of the concept of the reconfigurable sensor) was built using a double sided standard FR-4 PCB, with a thickness of 1.6 mm and copper thickness of 35.6 μm . The 90 fingers of the interdigitated structure have a length $l = 90$ mm and are separated by a distance $d = 0.5$ mm. The copper tracks were patterned on the top side of the PCB and, to provide shielding from external electric fields, the copper was not removed from the bottom side, forming a ground plane underneath the sensor's electrodes.

3.1. Capacitance Measurement

The novel interdigitated fringing field capacitor was measured using a GenRad 1659 RLC Digibridge, with the sensor inserted inside a recipient that was filled with controlled amounts of water. A photograph of the sensor inserted inside the tank partially filled with water is shown in Fig. 3.

We filled the transparent container with tap water (in several steps) and measured, with a rule attached to the edge of the container, the water level that is covering the sensor. For each water level we measured

the capacitance with the GenRad 1659 RLC Digibridge, connecting wires to the capacitance terminals of the sensor. The measured results are presented in Fig. 4.

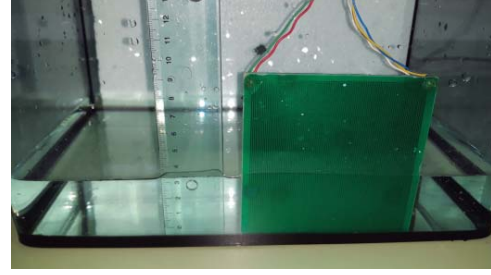


Fig. 3. Measurement set-up used for the capacitance test.

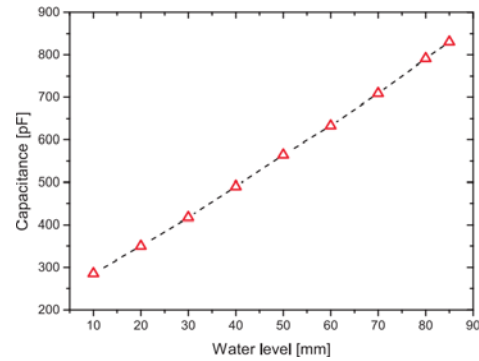


Fig. 4. Measured capacitance as a function of the water level.

3.2. Temperature Measurements

Copper has a very small TC, and the resistance of a copper wire can be written as:

$$R(T) = R_0[1 + \alpha(T - T_0)], \quad (2)$$

where R_0 is the value of the resistance at a temperature $T = T_0$, and $\alpha \approx 0.00420 \Omega / ^\circ\text{C}$.

From (2) we calculate:

$$\frac{dR(T)}{dT} = \alpha R_0 \quad (3)$$

In our sensor we measured, at $T = 25$ $^\circ\text{C}$, $R_0 = 7.913 \Omega$, so we have $dR(T)/dT \approx 33.2 \text{ m}\Omega / ^\circ\text{C}$. Due to this small TC, with a current $I_0 = 10$ mA flowing through the RTD (R_T), a temperature variation of $\Delta T = T - T_0 = 1$ $^\circ\text{C}$ results in a voltage change of only $\Delta V \approx 332 \mu\text{V}$.

Thus, to make it easier to measure the temperature with this RTD, we developed the simple circuit shown in Fig. 5.

A current source $I_0 = 10$ mA was applied to the sensor. The voltage on R_T is sent to an INA125 instrumentation amplifier, adjusted with a gain $G_V = 30.12$, so that the output of the instrumentation amplifier has a slope $dV_{\text{out}}/dT \approx 10 \text{ mV} / ^\circ\text{C}$.

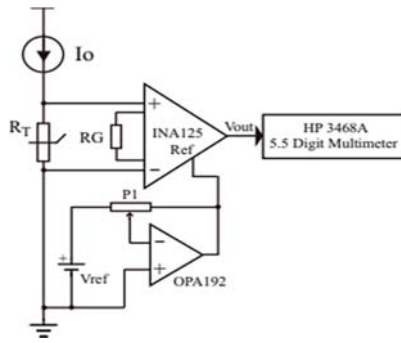


Fig. 5. Simplified schematic diagram of the circuit used to measure the temperature with the developed sensor.

Since at $T = 0^{\circ}\text{C}$ the resistance of the copper track is not zero, there is an off-set at the output of the amplifier at this condition. To cancel for this off-set, we added a negative voltage to the reference input of the instrumentation amplifier.

The sensor was installed inside a thermal environment chamber and a LM135A temperature sensor (from Texas Instruments, Dallas, TX, USA) was placed in contact with the sensor, to read its temperature.

The chamber's temperature was varied in the 9°C to 50°C range, and the values of V_{out} were measured with a HP 3468A digital voltmeter. The measured results of V_{out} as a function of the temperature (measured with the LM135A) are plotted in Fig. 6.

We notice that the sensor is extremely linear with temperature, since in the linear fit (presented in Fig. 6) we obtained an $R^2 = 0.99999$.

It is also possible to observe that the gain of the instrumentation amplifier was not set precisely to obtain a slope of $10.0\text{ mV}/^{\circ}\text{C}$ (we measured $9.956\text{ mV}/^{\circ}\text{C}$) and the off-set of the INA125 was also not precisely adjusted, since the linear fit indicates that at 0°C we will have $V_{out} \approx 2.28\text{ mV}$ (equivalent to 0.228°C).

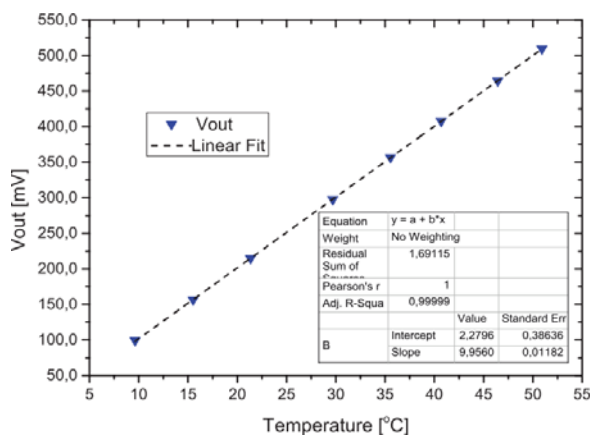


Fig. 6. Measured values of V_{out} as a function of the temperature.

4. Conclusions

A proof-of-concept prototype of the proposed reconfigurable sensor based on standard PCB

technology was designed, fabricated and tested in laboratory. The sensor can be used to measure capacitance in applications like soil moisture, air humidity, paper pulp water content. The sensor's copper tracks are used as an RTD and can measure temperature with high precision.

The proposed structure is suitable to design reconfigurable smart-sensors, since the signal processing circuit can be designed and implemented on the bottom side of the PCB. The measured results for the capacitance and temperature were in good agreement with the theoretical models. The proposed sensor can be used in applications where a rigid control of temperature and humidity is required, as for example in grain stored in silos.

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(007)

Modelling of Photoresponse Components Induced by Laser Pulse Across a p-n Junction

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Summary: The photoresponse signal induced by 1.06 μm laser pulse across a GaAs p-n junction is shown to be composed of three pronounced components resulting from hot carrier, heating lattice, and two-photon absorption phenomena. The first one is fast and has polarity of thermoelectromotive force of hot carriers. The second one has the same polarity, is slower, and is caused by the junction heating. The third one, respectively, is the classical electron-hole pair generation-induced photovoltage with polarity opposite to that of the first two. Good agreement between experimental and calculated results is achieved. The proposed model enables revealing contribution of each component to the net magnitude of the photoresponse, and can open the way in reducing negative impact of hot carriers into photovoltage of a solar cell.

Keywords: Photovoltage, Hot carriers, p-n junction, Laser radiation, Solar cell, GaAs.

1. Introduction

Renewable energy such as the one generated by solar cells is the most promising and environmentally friendly energy. Reduce of the production price of a solar cell and increase of its conversion efficiency will make photovoltaic energy more profit-making. Conversion efficiency of a single p-n junction semiconductor solar cell is fundamentally constrained by the Shockley-Queisser limit since only light photons having energy close to the forbidden energy gap are effectively used [1].

When photon energy is larger than semiconductor forbidden energy gap E_g , the illumination leads to electron-hole pair generation, and an ordinary photovoltage arises across a p-n junction due to separation of electrons and holes in the internal electric field of the junction. Extra photon energy left after the pair generation is used of no avail for the lattice heating. The Shockley-Queisser theory assumes that lower energy photons are not absorbed at all. However, it was shown that when the photon energy is lower than the forbidden energy gap, the intraband free carrier absorption leads to carrier heating. As a result, hot carrier photovoltage is induced across a potential barrier [2].

In this communication, we reveal simultaneous coexistence of the three photoresponse components induced by pulsed laser radiation across a GaAs p-n junction.

2. Experimental

The investigated GaAs p-n junctions (Fig. 1) were fabricated by liquid phase epitaxy-grown 5 μm -thick p-type layer (hole density $5 \times 10^{17} \text{ cm}^{-3}$) on n-type substrate with electron density $3 \times 10^{17} \text{ cm}^{-3}$.

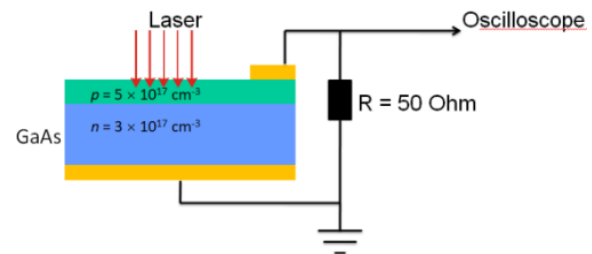


Fig. 1. Schematic of the sample and measurement circuit.

For excitation, Nd:YAG laser with wavelength 1.06 μm , pulse duration 25 ns, repetition rate 50 Hz, and maximum pulse power $1.6 \times 10^4 \text{ W}$ was used. Measurement scheme is presented in Fig. 1. Temporal behavior of the photoresponse and laser pulse shape were recorded by digital storage oscilloscope Agilent Technologies DSO6102A. All the measurements were carried out at room temperature.

3. Results and Discussion

Fig. 2 presents temporal profile of the photoresponse across the GaAs p-n junction induced by the laser pulse at moderate excitation level (20 % of maximum value). Obviously, the response consists of at least two components of opposite polarity.

The laser photon energy is less than GaAs forbidden energy gap, 1.17 eV and 1.42 eV, respectively. Therefore, our model assumes that the photoresponse across the GaAs p-n junction can be expressed as

$$U(t) = U_f(t) + U_t(t) + U_G(t), \quad (1)$$

where $U_f(t)$ is the hot carrier emf, $U_t(t)$ is the lattice heating component resulting from thermalization of

hot carriers, and $U_G(t)$ is electron-hole pair generation based component arising due to two photon absorption [3]. The polarities of the first two components and the third one are opposite.

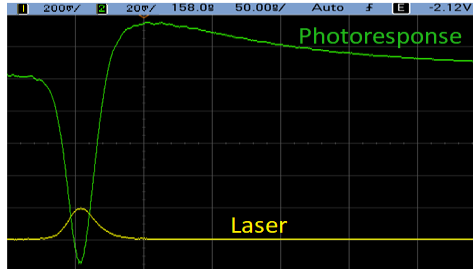


Fig. 2. Oscilloscope traces of photoresponse (green) across GaAs p-n junction and laser pulse (yellow).

Temporal analysis of the photoresponse was carried out to separate the input of each component. Time-dependence of the laser pulse intensity can be properly approximated as

$$I(t) = I_m \left(\frac{t}{\tau_p} \right)^{10} \exp \left[10 \left(1 - \frac{t}{\tau_p} \right) \right], \quad (2)$$

where I_m is the peak intensity and τ_p is the laser rise-time. The hot carrier component $U_f(t)$ can be expressed as

$$U_f(t) = I(t) \cdot K_f, \quad (3)$$

where K_f is the amplification coefficient of hot carrier emf. The lattice heating component $U_t(t)$ attributed to the thermoelectric emf can be found by solving the differential equation

$$\frac{dU_t(t)}{dt} = \frac{\bar{U}_t(t) - U_t(t)}{\tau_t}, \quad (4)$$

where τ_t is the thermalization time constant, and the forcing function is

$$\bar{U}_t(t) = I(t) \cdot K_t, \quad (5)$$

where K_t is the amplification coefficient of the thermoelectric emf. The two-photon excitation caused component of the photoresponse, $U_G(t)$, is found from

$$\frac{dU_G(t)}{dt} = \frac{\bar{U}_G(t) - U_G(t)}{\tau_G}, \quad (6)$$

where τ_G is the characteristic of decay time of $U_G(t)$ and

$$\bar{U}_G(t) = I^2(t) \cdot K_G, \quad (7)$$

where K_G is the amplification coefficient.

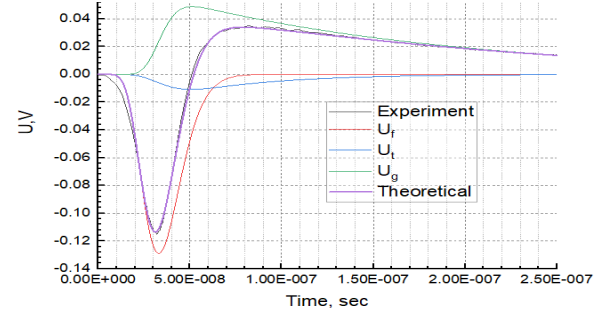


Fig. 3. Experimental photoresponse across GaAs p-n junction (black) and calculated its components.

The values of all time constants were taken from the experiment. Fig. 3 shows influence of all three components on photoresponse and good matching between experimental and theoretical photoresponses.

The first component, U_f , has polarity of thermoelectromotive force of hot carriers. The second one, U_t , also has the same polarity. The third component, U_G , is a classical carrier pair generation caused photovoltage with polarity opposite to the two others.

4. Conclusions

The proposed model of detection separate photoresponse components agrees well with the experimental results. Thus, it enables purifying individual input of each component and opens possibility to change their contribution into the net.

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(008)

Detection and Discrimination of Formaldehyde with CuO/SnO₂ Dual Layers MOS Gas Sensors Operated with a Pulsed Temperature Modulation

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Summary: In this work, the authors present a new sensor able to detect and discriminate sub-ppm level of formaldehyde. This metal oxide semi-conductors (MOS) gas sensor is based on nanostructured layers of SnO₂, CuO, and a combination of the two. These sensitive layers are deposited on silicon micro-hotplates and operated with a pulsed temperature modulation between 100 and 500 °C. This method yields to a very specific sensor response to formaldehyde compared to other gases such as CO or NO₂ under 50 % of relative humidity (%RH). This sensor set up (combination of sensitive layers and pulsed temperature mode) opens the way toward an efficient formaldehyde detection method for indoor air quality applications.

Keywords: Formaldehyde, Gas sensor, Metal oxides semiconductors, CuO-SnO₂ Bilayer material, Pulsed temperature modulation, Selectivity, Indoor air quality.

1. Introduction

Since few years, indoor air pollution has become a real health concern due to the awareness of presence of numerous gas pollutants issued from our direct environment (building, decoration or wooden materials, paints, etc.) [1]. These materials may release various volatile organic compounds (VOCs) such as formaldehyde, acetaldehyde, benzene, toluene etc.

According to the World Health Organization (WHO), formaldehyde (CH₂O) is considered as the main pollutants found in indoor air and it is known to be carcinogenic even for low concentration and short-term exposure conditions.

It is easily absorbed through the respiratory tract. Once entered into the respiratory system, it oxidizes rapidly to form carbon dioxide and formic acid which damages the epithelial cells of the respiratory system and therefore exerts a strong cytotoxic effect [2]. The threshold detection of this gas is set at 80 ppb according to the same organization [3]. Its detection has consequently become urgent.

We report in this paper a new MOS sensor based on a dual sensitive layer made of nanostructured SnO₂ and CuO, which can detect and discriminate sub-ppm concentration of formaldehyde at low energy consumption. This sensor performance is compared to sensors with simple layers of SnO₂ and CuO in order to evidence the role of the dual layer setup.

All these sensors are characterized under gas with an automatized experimental set-up and operated with a pulsed temperature modulation, which is a key feature for selective formaldehyde detection.

In Section 2, we describe the materials and methods used to build this paper and in the Section 3 we present

the famous results and our observations before short conclusion and perspectives.

2. Materials and Methods

The different sensitive layers of this study are prepared in the form of serigraphy pastes by a liquid chemical method with molecular precursors. This approach allows excellent control of shape, size and composition which are essential parameters for obtaining efficient gas sensors. In the next, we describe these sensitive layers.

SnO₂ nanoparticles have been obtained by the reaction of controlled quantities of water (8 molar equivalents) on the precursor of bis(dimethylamido) Tin(II) [(Sn ((NMe)₂)₂)₂] in the presence of dodecylamine (C₁₂H₂₇N) (DDA) (10 molar equivalent). The SnO₂ nanoparticles were observed with SEM at 1 μm and 100 μm, the pictures obtained are presented in Fig. 1.

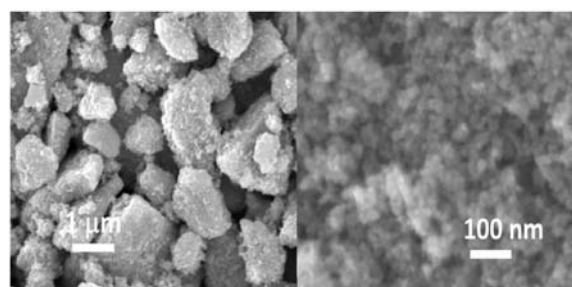


Fig. 1. SEM images of SnO₂ nanoparticles.

The protocol of the preparation of CuO nanoparticles is based on an organ-metallic approach to obtain small size nanoparticles [4].

CuO have been obtained by the reaction of ambient air at ambient atmospheric pressure for 24 hours on the precursor of copper acetin (Cu_{Acid}) in the presence of 16 molar equivalents of octylamine (OA). The excess OA in the medium is removed by several steps of precipitation by centrifugation and washing with THF.

The conditions of this reaction at room temperature and under ambient atmospheric pressure make it possible to obtain nanoparticles of small size.

The high resolution images clearly reveal the different sets of lattice planes inside the particles (See Fig. 2).

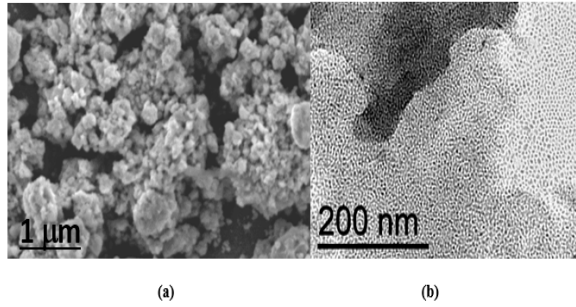


Fig. 2. Image of a CuO powder: a) SEM image, and b) TEM image.

The sensitive layers are then deposited locally on top of a silicon micro-hotplate [5] designed and realized in CNRS-LAAS in Toulouse and annealed slowly in ambient air from room temperature up to 500 °C in order to generate ionized oxygen species in atomic or molecular form at the oxide surface and therefore to improve the reactivity between the reacting gas and the sensor surface [4].

In this paper, we propose to study the specific behavior of a dual-layer sensor made of the successive deposition of SnO_2 (bottom) and CuO (top) material. For the demonstration, we compared the responses from this new sensor with “basic sensors” made by single layer of SnO_2 for one hand and with single layer of CuO for second hand.

All these three devices are operated with a pulsed temperature modulation applied to the integrated heater at a power of 12 mW and 50 mW, during 30 s each corresponding to temperature steps from 100 °C to 500 °C respectively (Fig. 3).

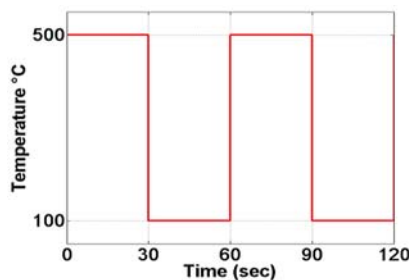


Fig. 3. Optimized temperature profile.

This optimal profile was obtained thanks to a specific study in order to determinate the operating temperatures that yield to the best sensitivity and their stage duration.

This pulsed temperature procedure reduces the power consumption of the device by almost 40 % compared to the standard continuous high temperature mode. In addition and above all, it offers a very specific response from the bi-layer SnO_2 -CuO sensor to formaldehyde exposure.

All the sensors have been tested under three different gases with an automatized test bench. Each sensor are placed in a 250 ml test chamber under 50 % of relative humidity obtained by bubbling synthetic air flow controlled by an analog mass flow controller.

All tests are following successive gas injections and summarized in Table 1.

Table. 1. Gas injection and their concentrations.

Step	Gas	Concentration (ppm)
1.	CH_2O	5
		10
2.	CO	50
		100
3.	NO_2	0.3
		0.5

3. Results and Discussion

The normalized resistance R_n of the sensors is calculated according the following expression (1) for a given temperature step.

$$R_n = \frac{R_{\text{gas}} - R_{\text{ref}}}{R_{\text{ref}}} * 100, \quad (1)$$

where R_{gas} represent the last point of the given step under target gas and R_{ref} the last point of the given step under air.

In the following part, we presents the normalized responses R_n measured at 100 °C and 500 °C of the different sensors to CO, NO_2 and CH_2O in two concentration for each gas injection.

In Fig. 4, we can recognize the response of SnO_2 layer that behaves as a n-type semiconducting oxide with a large resistance decrease (- 90 %) to 5 ppm formaldehyde at 100 or at 500 °C. This specific response could be explained by two possible reasons: the saturation of the sensitive layer during the detection of this gas or the insufficient of duration of the step at 500 °C to allow total desorption.

The SnO_2 sensor presents a moderate response to CO 100 ppm (- 25 %) and a huge resistance increase to 0.5 ppm NO_2 at 100 °C (+ 350 %). As we can see in Fig. 4, the measurements of the R_n must be taken at a

low temperature (100 °C) because it present the best sensitivity of this SnO₂ sensor.

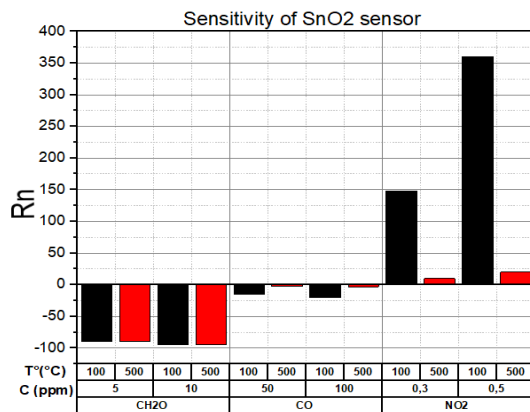


Fig. 4. Response of SnO₂ gas sensors operated with pulsed temperature modulation in different gaseous environment.

Without surprise, CuO sensor presented in Fig. 5 behaves as a p-type oxide with an increase of resistances under reducing gases injections (CO and formaldehyde) and a decrease of resistances under oxidizing one (NO₂). The very high response of this sensor to formaldehyde is remarkable (+ 450 % for 10 ppm at 100 °C). At high temperature (500 °C) the CuO sensor is not sensible for any gas.

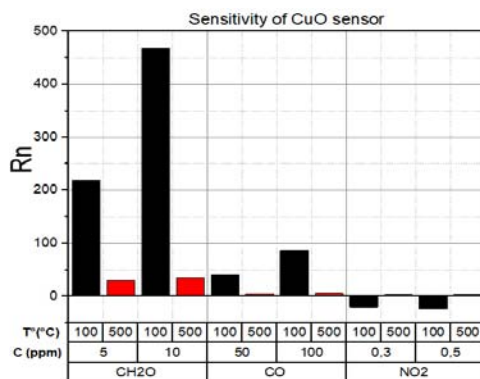


Fig. 5. Response of CuO gas sensors operated with pulsed temperature modulation in different gaseous environment.

The striking result of these experiments concerns the behavior of the bi-layer sensor made of a SnO₂ layer covered by a CuO layer illustrated in Fig. 6. This sensor presents a very complex response depending on the temperature of the operating cycle and the nature of the reacting gas.

Under CO, the response of this sensor is controlled by the p-type behavior of the top CuO layer where the electrical conduction is controlled by holes, whatever the temperature of the cycle (100 and 500 °C). But we obtained the best response at 500 °C (+40 % for 5 ppm and +60 % for 100 ppm).

Under NO₂, the sensor is not sensitive at 500 °C and only presents a weak n-type (bottom SnO₂ layer) response at 100 °C (+ 10 %) where the electrical conduction is controlled by electron whatever the

concentration. The bi-layer SnO₂-CuO reduce the response of this gas.

For formaldehyde concentrations (5 or 10 ppm), the sensor response is even more complex since the response type (n or p) varies depending on the measurement temperature step. The response of this sensor toward this gas can be controlled by the operating temperature step thank to the pulsed mode.

At 100 °C, the sensor response is driven by the SnO₂ layer with up to – 30 % of resistance drop for 10 ppm formaldehyde.

Surprisingly, at 500 °C the response type is reversed with around +25 % of the measured resistance increase to 5 ppm formaldehyde. The response in this step temperature follow the behavior of the top CuO layer.

Otherwise, the resistance are significantly decreased compared by the SnO₂ sensor and CuO sensor as well. It is also interesting to work with lower resistance for best signal-on-noise rate and to reduce global power consumption.

This complex response can be computed by PCA analysis (principal components analysis) to easily discriminate formaldehyde thanks to the pulsed temperature modulation.

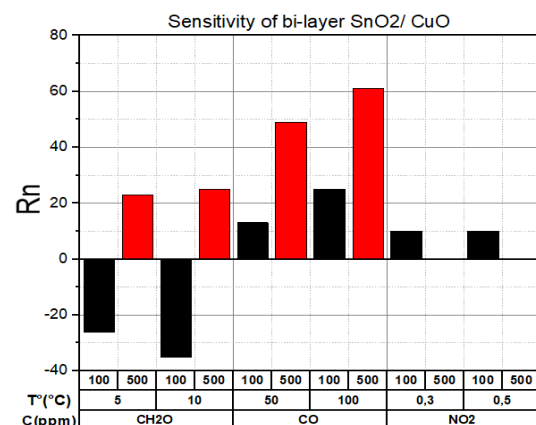


Fig. 6. Response of Bi-layer SnO₂/CuO gas sensors operated with pulsed temperature modulation in different gaseous environment.

4. Conclusions

We can conclude that the reversed response depending on the temperature has been only observed with the bi-layer SnO₂-CuO sensor and for CH₂O and then may represent a specific signature of the presence of this gas in the atmosphere. The combination of n and p type stacked layers coupled with pulsed temperature mode of the device represent an original and relevant way to improve the sensitivity and above all, selectivity of MOS gas sensors.

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Graphene-loaded Tin Oxide Nanofibers for Low Temperature Detection of Air Pollutants

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Summary: In this work we investigate the detection performance of nanofibers of tin oxide (SnO₂) towards sub-0.1 ppm nitrogen oxide (NO₂) and ozone (O₃) in air. It is found that SnO₂ nanofibers are more sensitive to O₃ than to NO₂ and the optimal working temperature differs significantly between the two gases: 200 °C (NO₂) and 350 °C (O₃). SnO₂ nanofibers loaded with reduced graphene oxide (RGO) show higher sensitivity than nanofibers of pure SnO₂ at low temperatures and the optimal working temperature diminishes by around 150 °C for the two gases. The air humidity does not modify the optimum working temperature of RGO-loaded SnO₂ nanofibers but it affects their sensitivity at temperatures below 150 °C. Irradiation with UV light is a promising method to activate gas desorption from RGO-loaded SnO₂ nanofibers so that this material could be used in chemoresistive sensors for low temperature detection of air pollutants.

Keywords: Tin oxide nanofibers, Reduced graphene oxide, Electrospinning, UV irradiation, Air pollution.

1. Introduction

Tin oxide (SnO₂) is probably the most well-known oxide and the most used material for gas detection. SnO₂ films are used as sensitive layers in commercial metal oxide semiconductor (MOS) gas sensors. Besides, SnO₂ nanostructures, particularly one-dimensional (1D) forms like nanowires, nanobelts nanorods and others, are currently studied because of their large surface-to-volume ratio and of nano-size effects, which can improve the sensing properties of MOS sensors [1]. However, the power consumption and fabrication cost of MOS sensors need to be further reduced to enable their massive deployment in wireless sensor networks for air pollution monitoring. In this regard, it has been found that the working temperature of graphene/MOS sensors is lower than that of MOS sensors, and eventually they may be operated at room temperature [2]. This result paves the way towards the industrial fabrication of nanostructured MOS sensors by using affordable substrate materials such as polymers, and simple and easy to scale-up techniques like electrospinning [3, 4]. The present work is aimed at assessing electrospun SnO₂ nanofibers as sensing material in chemoresistive sensors for detecting air pollutants. The detection performance of SnO₂ nanofibers towards low level (<100 ppb) nitrogen dioxide (NO₂) and ozone (O₃) in air is studied as a function of the sensor working temperature and air humidity. In addition, the effect of adding graphene to SnO₂ nanofibers is investigated with regard to their detection behavior at low temperature.

2. Experimental

2.1. Synthesis of Nanofibers

Nanofibers of SnO₂ are synthesized by electrospinning of an aqueous solution of polyvinyl-alcohol (PVA) and tin tetrachloride pentahydrate (SnCl₄·5H₂O) followed by thermal treatment [5]. The viscosity and electrical conductivity of the solution, and process parameters were optimized for controlling instability growth in electrospinning that affects the morphology of the nanofibers [6, 7]. The composition and the method for the preparation of the solution, and the parameters for electrospinning chosen in this work are summarized in Tables 1 and 2, respectively.

Table 1. Composition and preparation of precursor solution.

Compound	%wt	Solution	Preparation
DI H ₂ O Quimipur 231-791-2	80	#1 DI H ₂ O PVA	magnetic stirring (50 rpm) & heating (90 °C) 30 min
PVA Aldrich Chem. 341584	10	#2 DI H ₂ O SnCl ₄ ·5H ₂ O	magnetic stirring (50 rpm) 10 min
SnCl ₄ ·5H ₂ O Sigma Aldrich 244678	10	#1 + #2 DI H ₂ O PVA SnCl ₄ ·5H ₂ O	mixing & magnetic stirring (50 rpm) overnight

Electrospun PVA microfibers loaded with Sn were deposited onto a silicon substrate and calcined in air at 500 °C. Then, PVA decomposes and Sn oxidizes. The X-ray diffractogram of the as-prepared nanofibers is displayed in Fig. 1 (Malvern Panalytical Empyrean Cu LFF). They exhibit pure SnO₂ rutile structure with (110) and (101) as primarily exposed crystal faces.

Table 2. Electrospinning parameters.

Needle diameter (m)	Flowrate (m ³ ·s ⁻¹)	Voltage (V)	Distance (m)
5×10 ⁻⁴	3.3×10 ⁻⁸	19×10 ³	6×10 ⁻²

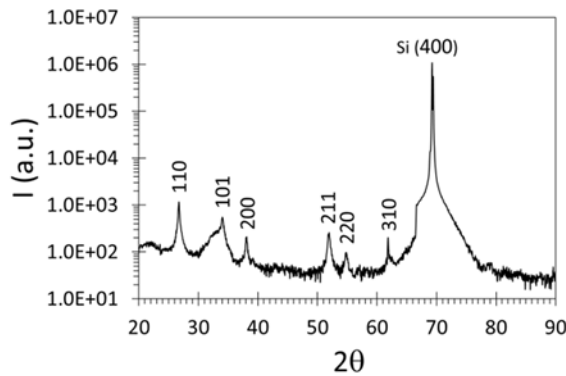


Fig. 1. XRD pattern of SnO₂ nanofibers on Si substrate.

The structure and the morphology of the SnO₂ nanofibers observed by SEM are shown in Fig. 2 (Hitachi S-4800, 0.5-30 kV). The nanofibers form a three-dimensional porous mat network on the substrate surface. They have diameters in the range of 30 nm to 100 nm and the diameter is uniform along the entire length of the nanofibers, which is in the order of tens of micrometers. Moreover, the nanofibers have a hollow core-shell tubular structure, with a granular shell consisting of grains of a few nanometers in size [8].

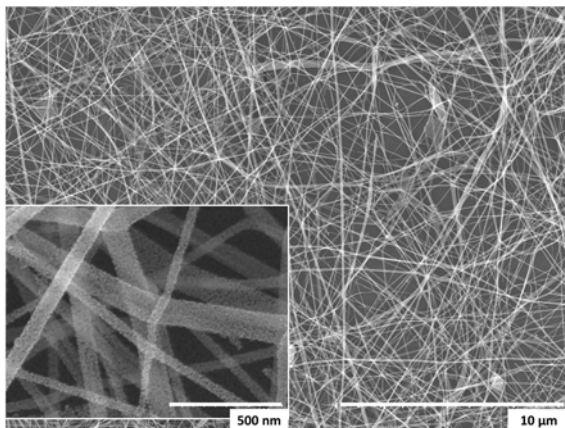


Fig. 2. SEM micrographs of electrospun SnO₂ nanofibers.

In addition, nanofibers of SnO₂ doped with graphene are produced by adding a dispersion of reduced graphene oxide (RGO, Abalonyx AS,

Norway) in isopropanol (0.1 mg·ml⁻¹) to the precursor solution and sonicating for half an hour. The content of RGO in the solution is 0.1 %wt with respect to Sn.

2.2. Preparation of Sensors

Sensors are prepared by depositing the electrospun PVA microfibers loaded with either Sn or Sn and RGO from the gas phase onto a square platform of micromachined silicon bonded to the surface of a TO8 housing. Fig. 3 shows a picture of the multisensor platform, which includes four microhotplates. A microhotplate consists of a circular dielectric membrane (SiO₂/SiN_x) and two spiral metallizations (Ti/Pt): interdigitated electrodes (IDTs) and a resistor or heater [9].

Tin loaded PVA microfibers are produced by electrospinning at room temperature in ambient air. The process maintains stable during periods of 10 minutes, when the fibers deposit onto the surface of the TO8 housing. Afterwards, the fibrous layer covering the IDTs (0.16 mm²) is calcined by heating the layer with the help of the microresistor lying underneath.

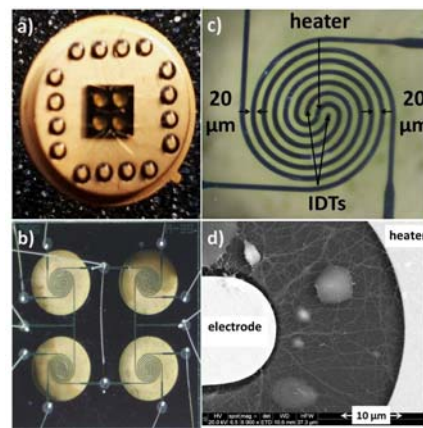


Fig. 3. a) Camera picture of TO8 housing with multisensor platform on the top; Optical microscopy images of b) Multisensor platform and c) Microhotplate; d) SEM micrograph of electrospun SnO₂ nanofibers onto microhotplate.

2.3. Characterization of Sensors for Gas Detection

Sensors based on nanofibers of SnO₂, both pure and doped with RGO, are characterized for the detection of nitrogen dioxide (NO₂) and ozone (O₃). Tests are conducted with gas cylinders of calibrated NO₂-air mixtures (~1 ppm) and O₃-air mixtures (≤0.3 ppm), the latter being generated by using a commercial ozone generator and a cylinder of synthetic air.

The facility used to characterize sensors for gas detection is depicted in Fig. 4. It shows a UV ozone generator (Fisher Scientific UVP SOG-1) and an ozone monitor (Aeroqual S500). The latter includes a semiconductor gas sensor (Aeroqual OZL) which measures the concentration of ozone in the mixture at

the outlet of the gas line. Other components are a gas mixer with three mass flow controllers and an electronic module. A gas bubbler at the exit of the dilution air line allows for humid air. The electronic module contains the sensor cell (0.85 cm³), a voltage supply to power the sensor heaters, a microprocessor, an analog-to-digital converter, and the electronics and communications needed to operate the sensors, record and filter the sensor signal, and display and store sensor data (operating temperature and resistance) and gas sample conditions (flowrate, temperature and humidity). The installation is fully automated and controlled remotely from a PC by using a dedicated software (LabView) program.

In the tests the sensors were exposed to the target gas mixture (adsorption) and then to dry/humid air (desorption). The gas sample was at room temperature, the relative humidity (RH) of the air varied between less than 5 % (dry air) and 80 %, and the flow rate was constant and the same in all tests (200 ml·min⁻¹). The sensor operating temperature was kept constant for any adsorption-desorption cycle and was varied from cycle to cycle from room temperature up to 350 °C.

3. Results and Discussion

3.1. Working Temperature

The sensor response is defined here as the ratio R_g/R_a where R_a stands for the resistance of the sensor under exposure to air, before the start of the adsorption phase; and R_g is the resistance of the sensor being

exposed to the target gas, at the end of the adsorption period. The response of sensors based on nanofibers of SnO₂ and on composite nanofibers of SnO₂ and RGO (SnO₂:RGO) towards NO₂ and O₃ in dry air is plotted as a function of the sensor operating temperature in Fig. 5.

The optimal working temperature (i.e. the temperature at which the sensor sensitivity is maximum) of the sensors using SnO₂ nanofibers is around 200 °C and 350 °C for NO₂ and O₃, respectively. These are consistent with the values of the optimum working temperature based on SnO₂ thin- films published in the literature. As a rule, the maximum sensitivity of SnO₂ sensors is observed at temperatures below 300 °C down to ambient temperature for NO₂ and above 300 °C up to 500 °C for O₃ [10, 11]. The kinetics of the interaction of these gases with the SnO₂ lattice and with the oxygen ions chemisorbed on the surface has been described elsewhere [12-15].

With the addition of RGO to SnO₂ nanofibers the sensitivity of the sensor in the low temperature range increases considerably and the optimal working temperature diminishes by nearly 150 °C, reaching ambient temperature for NO₂ and 200 °C for O₃. This result is attributed to the formation of local p-n heterojunctions between RGO nanosheets (p-type) and SnO₂ nanograins (n-type) on the surface and/or inside the nanofibers. A noticeable improvement of sensing performance towards NO₂ and other air pollutants has been observed with SnO₂-graphene composite materials by other authors too [16-18]. To the best of our knowledge, this is the first time that such behavior is reported for O₃.

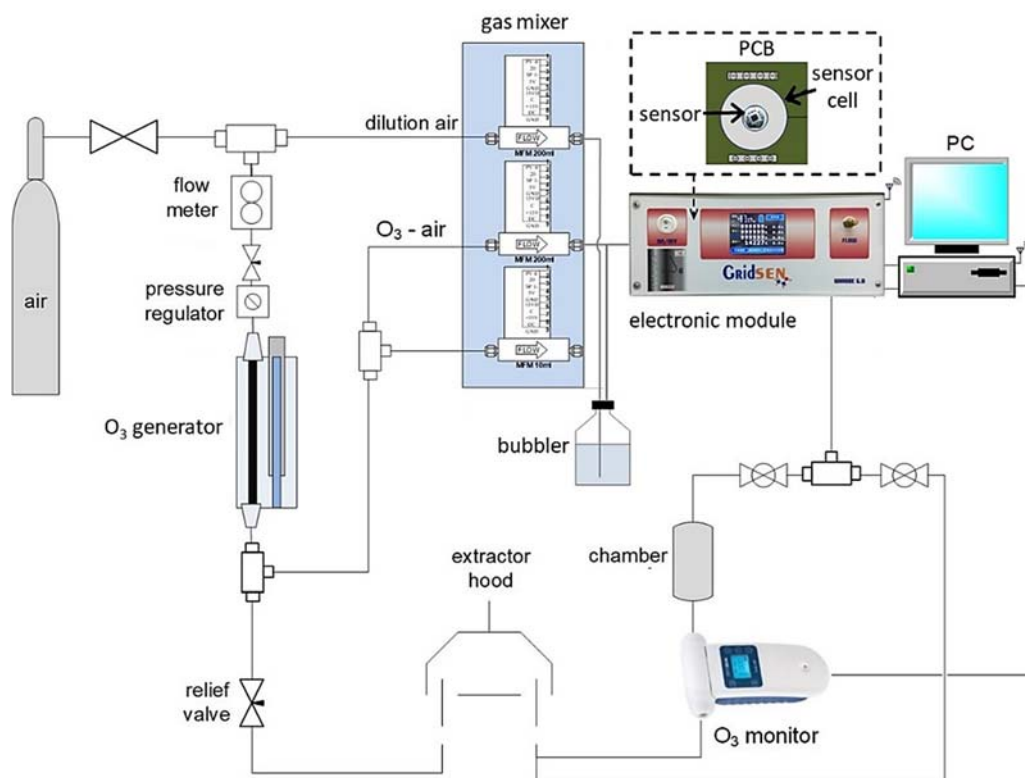


Fig. 4. Layout of the facility to characterize gas sensor detection behavior towards ozone.

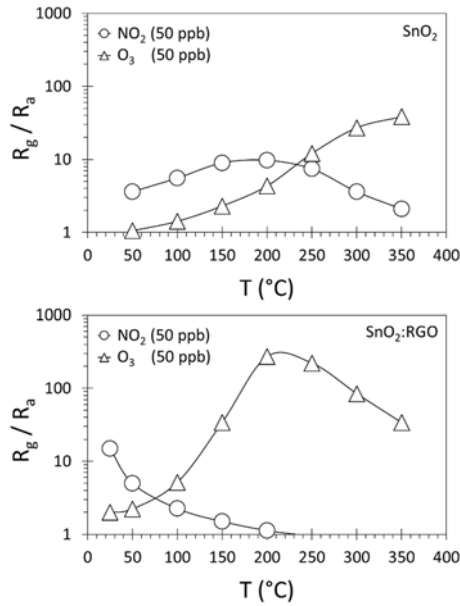


Fig. 5. Response of a sensor using SnO₂ nanofibers and RGO-loaded SnO₂ nanofibers to NO₂ and O₃ in dry air versus sensor working temperature.

Next, the effect of air humidity on sensor sensitivity is assessed. Fig. 6 compares the response of sensors using SnO₂ nanofibers doped with RGO towards NO₂ and O₃ in dry air and in humid air. It can be seen that the optimal working temperature is not affected by air humidity, but it does affect the sensitivity of the sensor at low temperatures. Below 150 °C the sensor response towards NO₂ decreases as air humidity increases and the opposite occurs for O₃. The effect of water vapor on the gas detection behavior of SnO₂ thin-films has been widely investigated [19, 20]. SnO₂ tends to absorb multiple layers of humidity from ambient air, forming a liquid water film on the surface. The active sensing layer in this case is the adsorbed water film, which responds only to gases that can easily dissolve and undergo electrolytic dissociation in water [21, 22]. Raising the substrate temperature towards 150 °C, the water coverage drops and large patches of free SnO₂ surface appear, providing adsorption sites for water vapor and other gases in the air (O₂, NO₂, O₃).

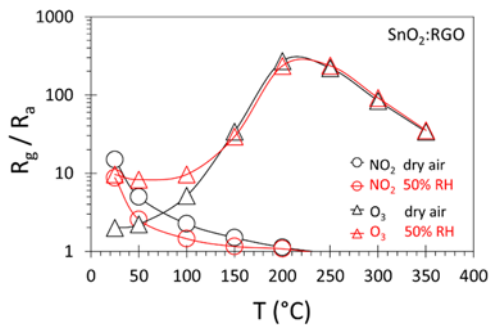


Fig. 6. Response of a sensor using RGO-loaded SnO₂ nanofibers to NO₂ and O₃ in dry air and in humid air (50 % RH) versus working temperature. The concentration of the gases is 50 ppb.

3.2. Recovery Speed

The results suggest using electrospun nanofibers of SnO₂ doped with RGO as sensing material for selective detection of NO₂ and O₃ in air. Acceptable operating temperatures are below 100 °C for NO₂ and between 100 °C and 200 °C for O₃. However, the slow recovery of MOS sensors at such low temperatures limits its suitability to air pollution monitoring.

Fig. 7 shows the recovery behavior of a sensor using RGO-loaded SnO₂ nanofibers after exposure to O₃ in humid air at various working temperatures. The recovery time τ_{rec} is the time it takes for the change in the sensor response when exposed to the gas to be reduced by 90 % once the gas is removed. The recovery time is depicted as a function of temperature in Fig. 7.

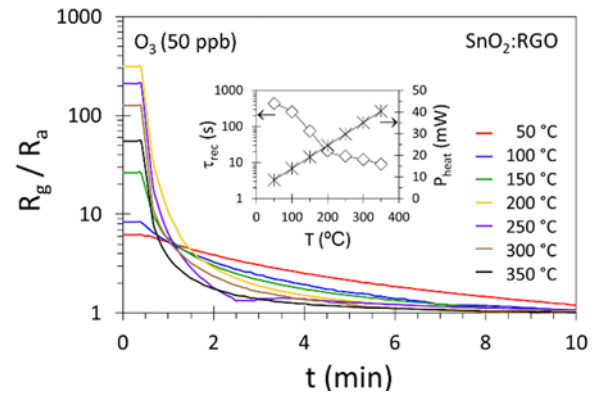


Fig. 7. Recovery behavior of a sensor using RGO-loaded SnO₂ nanofibers after exposure to O₃ (50 ppb) in humid air (50 % RH) at different working temperatures. Insert: Sensor recovery time (diamonds) and sensor heating power (crosses) versus working temperature.

The recovery time decreases with increasing temperature and reaches values acceptable for practical applications ($\tau_{rec} < 30$ s) at around 200 °C. Therefore, the sensors should be operated in dynamic (variable temperature) mode instead of isothermal (constant temperature) mode, i.e. at low temperature for gas adsorption and high temperature for gas desorption [23, 24].

Also, it has been demonstrated that the low temperature sensing performance of MOS sensors can be improved by irradiation with UV light [25-29]. Here we investigate UV irradiation for low temperature gas desorption from RGO-loaded SnO₂ nanofibers. To that aim, we use a UV-LED source with peak wavelength λ_{max} at 356 nm (OSA Opto Light, OCU-400 UB355). The UV lamp is placed centered on the top wall of the sensor cell facing the surface of the multisensor platform at a distance of nearly one cm. Sensors are subjected to adsorption-desorption cycles in which the sensor temperature is kept constant and the UV lamp is switched on only in the gas desorption phase.

Fig. 8 shows the response and the recovery time of a sensor in a test with a mixture of O₃ and humid air, the sensor operating and 100 °C, and the power of the

UV source varying from null (lamp off) up to 55 mW. It can be observed that UV irradiation causes an enhancement of the sensor response and a reduction of the sensor recovery time, even at the lowest power of 3.3 mW used in the tests. A kind of saturation stage is reached at around 20 mW with a recovery time in the order of 60 s.

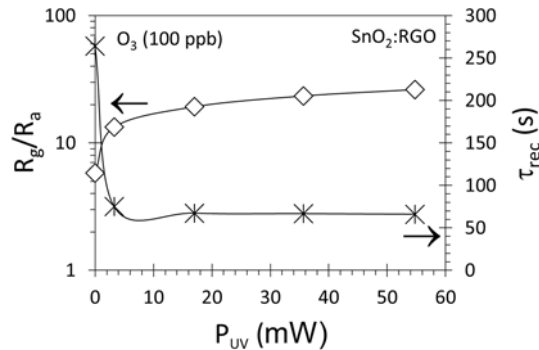


Fig. 8. Response (diamonds) / recovery time (crosses) of a sensor using RGO-loaded SnO₂ nanofibers under/after exposure to O₃ (100 ppb) in humid air (40 % RH) versus power of UV lamp in the desorption phase. The sensor operates in isothermal mode at 100 °C.

Finally, we assess the stability of the sensors working in 1) dynamic mode: 100 °C for gas detection (T_{ads}) and 200 °C for sensor recovery (T_{des}), no UV irradiation; 2) dynamic mode: 100 °C for gas detection and 350 °C for sensor recovery (T_{des}), no UV irradiation; and 3) isothermal mode: 100 °C for gas detection and sensor recovery, UV irradiation at 5 mA for gas desorption.

The tests run for 16 h and consist of 50 cycles, each of a duration of 20 min: 10 min for gas adsorption (T_{ads}), 5 min for sensor heating ($T_{ads} \rightarrow T_{des}$) and gas desorption (T_{des}), and 5 min for sensor cooling ($T_{des} \rightarrow T_{ads}$). In 3) both T_{ads} and T_{des} are set to 100 °C. The evolution of the sensor response in tests with O₃ in humid air is displayed in Fig. 9.

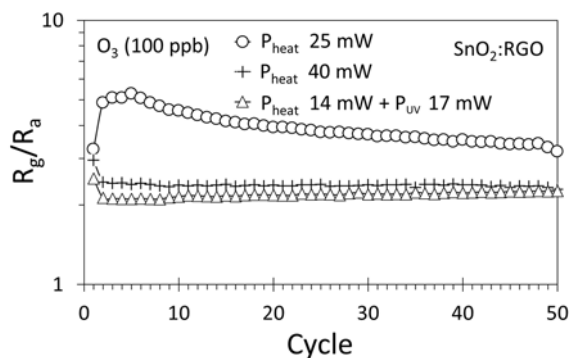


Fig. 9. Response of a sensor using RGO-loaded SnO₂ nanofibers under exposure to O₃ (100 ppb) in humid air (40 % RH) over adsorption-desorption cycles. Sensor temperature for gas adsorption is 100 °C and for gas desorption: 200 °C (P_{heat} 25 mW), 350 °C (P_{heat} 40 mW), and 100 °C (P_{heat} 14 mW) with UV irradiation at 5 mA (P_{uv} 17 mW).

The sensor shows a rather similar response, both in value and stability, in tests 2) and 3). However, the power consumption of the sensor in the gas desorption phase is lower in test 3). Assuming that UV irradiation affects the four sensors equally, each sensor consumes around 18.25 mW for gas desorption in test 3), compared to 40 mW in test 2).

4. Conclusions

It has been demonstrated the feasibility of electrospun RGO-loaded SnO₂ nanofibers for selective detection of NO₂ and O₃ in air at temperatures below 100 °C with good sensitivity in the sub-0.1 ppm range. It has also been proven that thermal activation and UV photoactivation can effectively speed-up gas desorption from the nanofibers of SnO₂:RGO at low temperatures. For practical applications it is still necessary to design and assess heating and UV irradiation strategies that guarantee the stability of the sensor in the long-term.

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A LVDT Based System with SW Synchronous Detection Capabilities

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Summary: This paper presents an automatic measurement system for positioning measurements. The system is based on an LVDT transducer and a LabVIEW program was developed for synchronous detection of the LVDT output signal. A complete characterization of the measurement system using a linear actuator was performed in order to obtain accurate metrological values of linearity and hysteresis. The paper will also include data reduction and statistical processing of measurement data to validate the measurement system performance and conformity. It is also important to underline that a preliminary version of the system presented in this paper has been successfully used in electronics automotive manufacturing plants to control the positioning of pointers base on the instrument panel that is a critical operation in cockpits assembly.

Keywords: LVDT, Linearity, Sensitivity, Hysteresis, Synchronous detection, Automotive industrial application.

1. Introduction

Automatic measurement systems [1] have been widely implemented in advanced technological industries. Its metrological capabilities, flexibility, reliability, and low cycle time, enable them to answer simultaneously to the high-quality product specification and high volume production requirements. The emergency of Industry 4.0 [2] highlights the importance of data connection between measurement systems with automatic production cells and information systems. This is, nowadays, a key factor of competitive advantage enabling real process quality control, the statistical analysis and product traceability through serial number barcode reading of RFID [3]. The availability of measurements big data incentives the development of process continuous improvement activities through Lean Six Sigma [4] assisting, at the same time, the engineering program development.

The automotive electronics industry is a typical case where this reality supports a highly competitive environment with tight product specifications and high demanded customer requirements.

This paper describes the development of an automatic measurement system for positioning measurement, based on an LVDT transducer, used during the manufacturing process of electronic instrument panels usually called instrument cluster. In spite of the proliferation of full digital clusters, the use of dc and stepper motors, with pointers assembled, to indicate speed and rpm's are still the preferred option for high-end series of car makers, due to its metrological attributes. During the electronic cluster manufacturing process, one of the most critical steps is the pointer placement, corresponding to the insertion of the pointer on the motor shaft. This is a meticulous process step that needs to be assisted with a real-time pointer height measurement. The control of the pointer

height (distance between the pointer base and graphical scale dial) prevents the occurrence of two distinct failure modes, sticky pointers and loose pointers. Sticky pointer failure mode occurs when the pointer base touches the graphical scale dial, blocking the pointer movement or causing the intermittent pointer displacement. The loose pointers failure mode is the results of a pointer with a high gap between the pointer base and graphical scale dial. In these conditions, the contact area between the pointer hub and the motor shaft is not enough to create retention force to support high torques imposed by motor accelerations.

2. Measurement System

The novelty of this automatic measurement system relies on the fact that it can be integrated into the typical design of automatic pointer placement machines, with a simple coupling of an LVDT sensor and a data acquisition card connected to the machine computer. All the signal processing and data interchange is done by SW without the requirement of any additional hardware integration. The automatic measurement system enables the accurate real-time measurement of pointer height, during the pointer insertion process, preventing the failure modes occurrence avoiding, that way, further testing when product is fully assembled.

2.1. LVDT

The LVDT transducer, interconnected with blocks of Hardware and Software, is the key element of the measurement system, converting a rectilinear motion of a tool to which is coupled mechanically into a proportional output electrical signal.

Due to de single-axis sensitivity, the LVDT responds to the displacement of the core along the coil's axis. That way, the LVDT can be integrated onto the production system, with the proper mechanical fixtures, sensing the gap between pointer base and graphic scale dial.

The LVDT is an inductive passive sensor with a primary winding centered between a pair of identical secondary windings, with phase inversion, forming the stationary element of the sensor. The moving element is the core of the LVDT made of magnetically permeable material.

The primary winding is supplied with an alternating current defined by

$$V_i = V_{im} \sin(\omega_0 t), \quad (1)$$

where V_{im} is the amplitude of the signal, $\omega_0 = 2\pi f_0$ and f_0 is the frequency of the input signal. The output signal is the differential ac voltage resulting from the magnetic flux induced by the primary in the two secondary windings, which varies with the axial position of the core within the LVDT coil. The output signal can be described by

$$V_o = V_{om} \sin(\omega_0 t + \phi), \quad (2)$$

where $\phi = 180^\circ$ if V_o is the phase shifted in regards to V_i or, $\phi = 0^\circ$ if V_o is in phase with V_i .

As shown in Fig. 1, when the core is located midway between the two secondary windings S1 and S2 the output voltage is zero, since voltage V1 across S1 is equal to voltage V2 across S2.

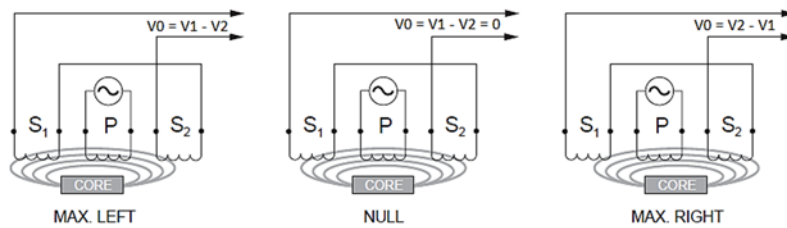


Fig. 1. LVDT core in different axial positions.

As core is moved closer to secondary S2 more flux is coupled in S2 resulting that V2 voltage increases while V1 voltages decrease. Within these conditions, V_o is 180° phase shifted with V_i . When the core is moved closer to S1 in regards S2, the V1 voltage becomes greater than V2. At this stage V_o is in phase with V_i .

2.3. System Hardware

Fig. 2 represents the hardware elements used in this measurement system. The LVDT Transducer is excited with an external Signal Generator capable of supplying the required LVDT primary input signal. The Oscilloscope enables the visualization of oscillograms of LVDT input and output signals as per equations (1) and (2). The Micrometric Screw works

as counter measurement element for the characterization of the LVDT output response function of displacement. The DC Linear Actuator imposes the required linear displacement to the LVDT shaft/core, simulating the precise movement during the pointer placement insertion, supporting, that way, the full characterization of the measurement system. The DC Driver Controller, supplied with a 12V external power source, supplies the required dc voltage to control the speed and direction (FW-Forward and BW-Backward) of the DC Linear Actuator. The National Instruments myDAQ acquisition card interfaces with LabVIEW [5] Software through PC/USB. The LVDT input and output signal are also acquired by NI myDAQ for signal processing. The myDAQ controls also the DC Driver Controller with a PWM signal generated with LabVIEW.

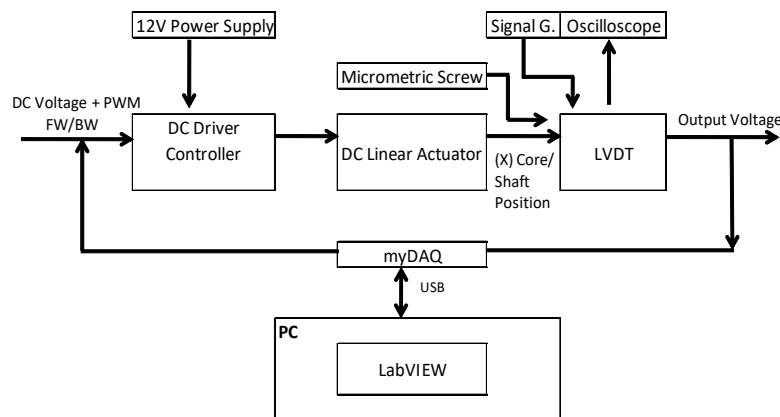


Fig. 2. Measurement System block diagram.

2.4. Software

The Software of this automatic measurement system was fully developed using the National Instruments LabVIEW. This SW, made of several LabVIEW subVI's (Virtual Instrument subroutines), aggregates all the elements of the measurement system: The signal acquisition, dc linear actuator and the data measurements collection. The main LabVIEW subVI, represented in Fig. 3, performs the acquisition

of the LVDT input and output signals, V_i and V_o , assisted by myDAQ acquisition card. Once V_i and V_o signals are acquired it is possible to perform an extended range of electric measurements required for the signal processing. For that purpose, it was used the following VI's: DAQ Assistant, Amplitude and Level Measurements, Spectral Measurements and FFT, and a Butterworth 2nd order LPF.

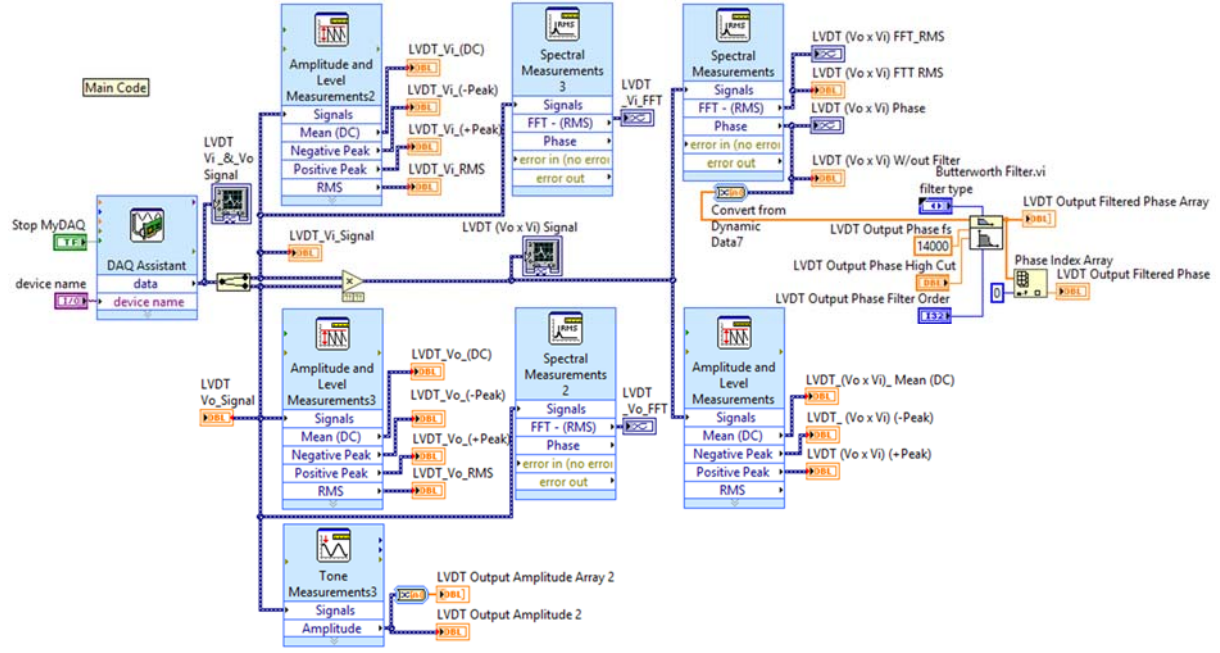


Fig. 3. Main stream software subVI for data acquisition and signal processing.

2.5. Synchronous Detection

The synchronous demodulation method was applied to transform the LVDT ac output signal into a dc signal proportional to the displacement of the LVDT shaft/core. To implement this method the LVDT input signal V_i was used as the synchronous signal, for shift lock phase, and its multiplication was performed with the LVDT output signal V_o as described by equation (3).

$$= \begin{cases} \frac{V_{im}V_{om}}{2} - \frac{V_{im}V_{om}}{2} \cos(2\omega_0 t); \phi = 0^\circ \\ -\frac{V_{im}V_{om}}{2} + \frac{V_{im}V_{om}}{2} \cos(2\omega_0 t); \phi = 180^\circ \end{cases} \quad (3)$$

where $\frac{V_{im}V_{om}}{2}$ represents the dc component of the output signal. Figs. 4 and 5 illustrate the oscillograms of the input signal V_i , the output signal V_o and the signal resulting from it's multiplication ($V_i \times V_o$).

Using the spectral measurement VI it was possible to obtain the FFT of the input and output signals multiplication and, thus, detect the phase transition of the ac LVDT output between 0° and 180° to obtain a

continuous dc output signal, proportional to LVDT shaft/core displacement.

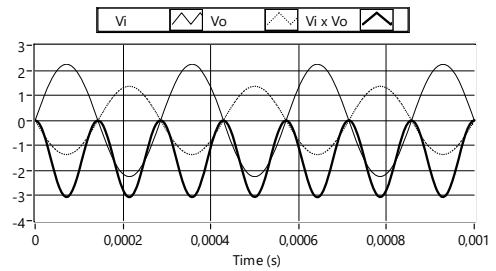


Fig. 4. Input voltage V_i , Output voltage V_o 180° shifted and $V_i \times V_o$.

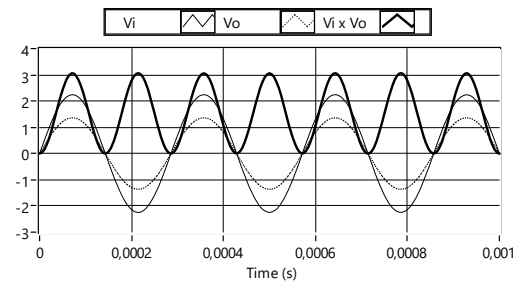


Fig. 5. Input voltage V_i , Output voltage V_o in phase and $V_i \times V_o$.

Therefore, re-writing equation (3) as,

$$f(t) = \pm \left(\frac{V_{im}V_{om}}{2} - \frac{V_{im}V_{om} \cos(2\omega_0 t)}{2} \right), \quad (4)$$

it can be expanded as a Fourier series [6],

$$f(t) = a_0 + \sum_{n=1}^{+\infty} [a_n \cos(n\omega_0 t) + b_n \sin(n\omega_0 t)], \quad (5)$$

being

$$\begin{cases} a_n = \frac{2}{T} \int_0^T f(t) \cos(n\omega_0 t) dt; n = 0 \text{ or } n \text{ even} \\ b_n = \frac{2}{T} \int_0^T f(t) \sin(n\omega_0 t) dt; n \text{ odd} \end{cases}, \quad (6)$$

for $n = 0$ and $n = 1$ we have respectively

$$\begin{cases} a_0 = \frac{V_{im}V_{om}}{2} \\ a_1 = \frac{V_{im}V_{om}}{2} \frac{2}{T} \int_0^T \cos^2(2\omega_0 t) dt \end{cases}, \quad (7)$$

where $T = 2\pi/\omega_0$. Fig. 6 represents a_0 and a_1 , equation (7), obtained with the FFT from signal

$(V_i \times V_o)$ and define, respectively, the dc component proportional to LVDT core displacement and the amplitude of the first harmonic of the signal. Extracting the rms value of a_1 from equation (7) we have:

$$a_{1 \text{ rms}} = \frac{V_{im}V_{om}}{2\sqrt{2}}, \quad (8)$$

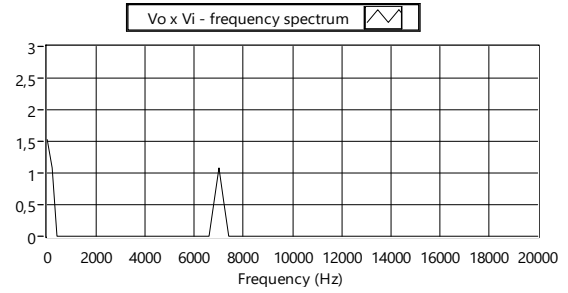


Fig. 6. Frequency spectrum of $V_i \times V_o$.

The Fig. 7 illustrates the synchronous phase-detection demodulating the LVDT output ac signal V_o into a dc continuous signal.

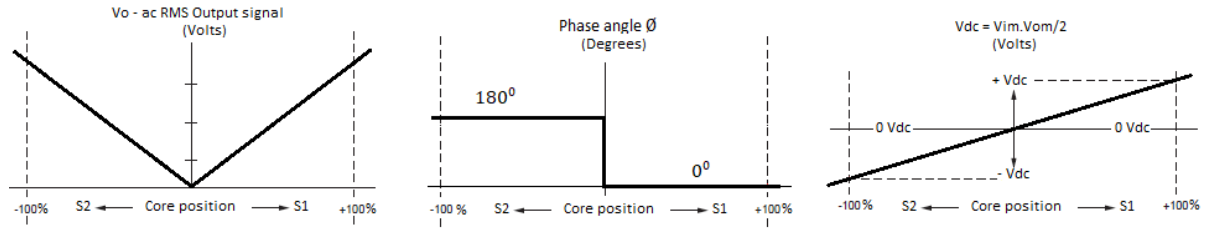


Fig. 7. LVDT output signal conversion from ac to dc.

3. Some Results

The dc LVDT output signal (V_{dc}) enabled the complete characterization of the measurement system that was done in two steps. At a first stage, a Micrometric Screw was used to manually actuate the LVDT shaft/core and, therefore, obtain a complete correlation between physical dimensions in mm and the linear ac LVDT output. At a second stage, the DC Linear Actuator was used, in an automatic sequential

way, exercising the LVDT shaft/core to travel back and forward along its full course.

3.1. Linearity and Sensitivity

Fig. 8 contains the graphic representation of the measurement data taken during the characterization with the Micrometric Screw.

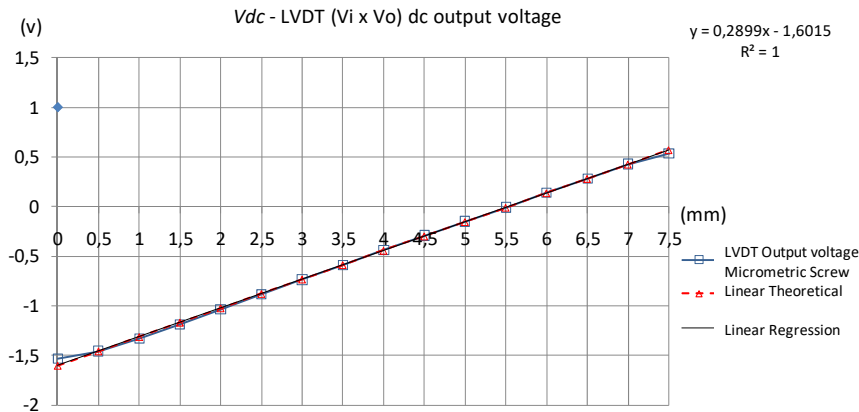


Fig. 8. LVDT characterization, dc output voltage measurements (V) along full shaft/core displacement (mm).

The data was treated statistically and the linear regression equation obtained is defined as:

$$V_{dc} = 0.2899x - 1.6015, \quad (9)$$

where the variables V_{dc} and x represents, respectively, the LVDT dc output in Volts (V) and the LVDT shaft/core displacement in (mm). The sensitivity obtained is 0.2899 V/mm and the linearity has a maximum local absolute error of 5.6 mV and a

maximum local relative error of 1.89 % within the 0.5 mm and 7.0 mm of the LVDT shaft/core course.

3.2. Hysteresis

Table 1 shows the hysteresis absolute and relative errors obtained with the 2 steps of the measurement system characterization described.

Table 1. Data of absolute and relative hysteresis errors.

Shaft/Core (mm)	Shaft/Core FW motion		Shaft/Core BW motion		Hysteresis Abs. Error (V)	Hysteresis Rel. Error (%)
	LVDT-Vdc ($V_i \times V_o$) average (V)	Step (V)	LVDT-Vdc ($V_i \times V_o$) average (V)	Step (V)		
0	-1.5301		-1.5315		0.0014	0.07
0.5	-1.4565	0.0735	-1.4580	0.0735	0.0015	0.07
1	-1.3290	0.1275	-1.3327	0.1253	0.0037	0.18
1.5	-1.1839	0.1452	-1.1845	0.1482	0.0006	0.03
2	-1.0338	0.1501	-1.0374	0.1472	0.0036	0.17
2.5	-0.8845	0.1493	-0.8891	0.1483	0.0046	0.22
3	-0.7359	0.1487	-0.7354	0.1537	0.0004	0.02
3.5	-0.5873	0.1486	-0.5856	0.1498	0.0017	0.08
4	-0.4399	0.1475	-0.4400	0.1456	0.0002	0.01
4.5	-0.2914	0.1485	-0.2909	0.1492	0.0005	0.02
5	-0.1496	0.1418	-0.1463	0.1446	0.0034	0.16
5.5	-0.0068	0.1429	-0.0015	0.1448	0.0053	0.26
6	0.1379	0.1446	0.1410	0.1425	0.0032	0.15
6.5	0.2831	0.1453	0.2894	0.1484	0.0063	0.31
7	0.4280	0.1449	0.4327	0.1433	0.0046	0.23
7.5	0.5323	0.1043	0.5322	0.0995	0.0001	0.01
Max	0.5323		0.5322			
Min	-1.5301		-1.5315			
Span	2.0624		2.0637			

As it can be seen, the maximum local absolute error is 6.3 mV, representing a relative error of 0.31 %.

$$\begin{cases} H_0: \eta_1 - \eta_2 = 0 \\ H_1: \eta_1 - \eta_2 \neq 0 \end{cases}, \quad (10)$$

3.3. Statistical Validation

In order to validate the LVDT characterization process described in Section 3, it was done the comparison of the linear absolute errors obtained with two actuation methods of the LVDT shaft/core, the manual method with the Micrometric Screw versus the automatic method with the DC Linear Actuator. For this propose it was used the sampling data collection obtained with the measurement system. Considering the number o data points used to characterize the linearity (Table 1) and the fact that its statistical distribution is at this stage unknown, it was proposed the use of a statistical non-parametric hypothesis test, the Mann-Whitney-Wilcoxon (MWW) [8].

The MWW test the differences between the medians of the absolute errors of the two methods. Therefore, being η_1 the median of the linearity error associated with the Micrometric Screw actuation and η_2 the median of the linearity error associated with DC Linear Actuator method, we can define $\eta_1 - \eta_2$ as the median difference between the two methods. Thus, we can define the Null Hypothesis H_0 and the Alternative Hypothesis H_1 as follows:

Using the statistical SW MINITAB [7] it was possible to apply the MWW bilateral test to the difference of medians with a confidence level of 95 % ($\alpha = 5\%$). As it can be observed on Table 2, it was obtained a P-Value = 0.8. As P-Value $> \alpha = 0.05$ Null Hypothesis H_0 can't be rejected, demonstrating that there are no evidences of statistical significant difference between the medians of the two actuation methods. This result validates the LVDT characterization process used.

Table 2. MWW test summary results.

Test	
Null hypothesis	$H_0: \eta_1 - \eta_2 = 0$
Alternative hypothesis	$H_1: \eta_1 - \eta_2 \neq 0$
W-Value	P-Value
209,00	0,800

4. Conclusions

The present paper explained how an LVDT can be successful implemented in the automotive electronics

industry. The implementation of this automatic measurement system based on an LVDT with SW Synchronous Detection Capabilities demonstrated, for one side, the good metrological capabilities, with lower errors of linearity and hysteresis and, on the other side, the exceptional flexibility and adaptability to industrial environments technologically advanced, where the interconnectivity of data measurements is fundamental. This measurement system enables the real-time process control of the pointer height during the manufacturing of the instrument panels, contributing to the required high quality levels in the industry. Due to the positive results obtained the system has been implemented across the world in multiple automotive electronic plants and is considered one of the best-known methods.

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Lectin Biosensors for the Selective Detection of Cancer Glycobiomarkers

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Summary: A common feature seen in different types of cancer, namely in carcinomas, is the alteration in post-translational modifications of proteins, particularly in protein glycosylation [1]. Due to increased or decreased expression of particular enzymes participating in glycosylation of proteins, different glycan structures are formed, which are typical of tumoral cells. When the respective glycoproteins are secreted into the blood stream, these aberrant structures can be used as valuable cancer biomarkers, since they are not synthesized by normal cells.

Glycan structures are efficiently and selectively detected by lectins, which are proteins of natural origin with high affinity for a particular or a very small group of glycan epitopes. Lectins are used for glycan detection as antibodies are used for protein recognition.

Lectin biosensors are attractive devices for the detection of cancer-associated glycobiomarkers in serum since they combine the advantageous aspects of biosensors (portability, easy use in point-of-care analysis, low sample requirement) with the high selectivity of lectin biorecognition. This work presents three lectin-based impedimetric biosensors for the selective detection of specific aberrant cancer-associated *O*-glycans, namely STn, Tn and T antigens, which are well-established pan-carcinoma biomarkers. For these three biosensors, *Sambucus nigra* agglutinin, *Vicia villosa* agglutinin and *Arachis hypogaeae* agglutinin were used as biorecognition elements, with specificity for STn, Tn and T antigens, respectively.

Keywords: Cancer, Glycobiomarkers, Truncated *O*-glycans, Lectins, Biosensors, Electrochemical impedance spectroscopy.

1. Introduction

Cancer is a group of diseases characterized by a wide diversity in location, cellular features and aggressiveness. However, carcinomas present some common aspects, one of them being alterations in protein glycosylation [1, 2], generating aberrant glycoproteins with glycostructures which are typical of carcinoma cells and that can be used as carcinoma-associated glycobiomarkers.

Detection of aberrant glycoproteins in serum can be successfully performed by using lectin biosensors, as lectins show high selectivity towards particular glycan structures. Lectin biosensors are attractive devices for the detection of cancer-associated glycobiomarkers in serum since they combine the advantageous aspects of biosensors (portability, easy use in point-of-care analysis, low sample requirement) with the high selectivity of lectin biorecognition [3].

This work presents three lectin-based impedimetric biosensors for the selective detection of specific aberrant cancer-associated *O*-glycans, namely STn, Tn and T antigens, which are well-established pan-carcinoma glycobiomarkers. For these three biosensors, *Sambucus nigra* agglutinin, *Vicia villosa* agglutinin and *Arachis hypogaeae* agglutinin were used as biorecognition elements, with specificity for STn, Tn and T antigens, respectively.

1.1. Reagents, Materials and Measurement

The lectin biosensors were constructed on screen printed gold electrodes (Dropsens). *Sambucus nigra*

agglutinin, *Vicia villosa* agglutinin and *Arachis hypogaeae* agglutinin (in pure form, all from EY Laboratories) were used as biorecognition elements. Lectin immobilization was performed by covalent binding to self assembled monolayers (made with 16-mercaptohexadecanoic acid and/or 6-mercaptohexanoic acid) via amine coupling, using N-(3-dimethylaminopropyl)-N'-ethylcarbodiimide and N-hydroxysulfosuccinimide. Ethanolamine was used to block unoccupied terminal carboxyl groups of alkanethiols [4-6]. Several glycoproteins were used as positive or negative controls, according to the glycan epitopes carried by them. For the SNA biosensor, bovine submaxillary mucin and human transferrin were the model glycoproteins; for VVA biosensor, asialofetuin and asialo-bovine submaxillary mucin were used and for PNA biosensor asialofetuin was the model glycoprotein.

The binding event between each lectin and the corresponding aberrant *O*-glycan was monitored by electrochemical impedance spectroscopy. The biosensor impedance was measured before incubating the sample, then the sample was incubated for 10 min and, after a washing step, the biosensor impedance was measured again. The difference in the impedance (final minus initial) divided by the initial impedance was the analytical signal. The increase in impedance was related to the lectin-glycan complex formation.

All impedimetric measurements (performed by electrochemical impedance spectroscopy) were carried out in an Autolab electrochemical system (Metrohm model PGSTAT 302N) equipped with a FRA32M module and controlled through Nova software version 2.0. Impedance measurements were performed at the

formal potential of the $\text{Fe}(\text{CN})_6^{4-}/\text{Fe}(\text{CN})_6^{3-}$ pair, with a 5 mV sinusoidal excitation amplitude, within a full frequency range from 0.010 Hz to 100 kHz.

2. Results

2.1. Selectivity Evaluation

Selectivity of the three lectin biosensors was assessed towards different glycoproteins carrying the glycan epitope of interest and other glycan epitopes (Figs. 1 and 2).

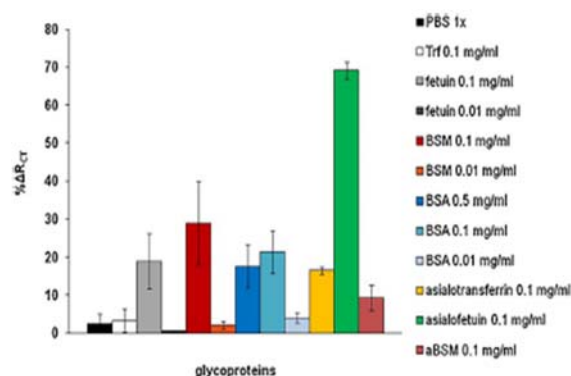


Fig. 1. Responses obtained for diverse glycoproteins incubated with the VVA biosensor for 10 min, at room temperature [5].

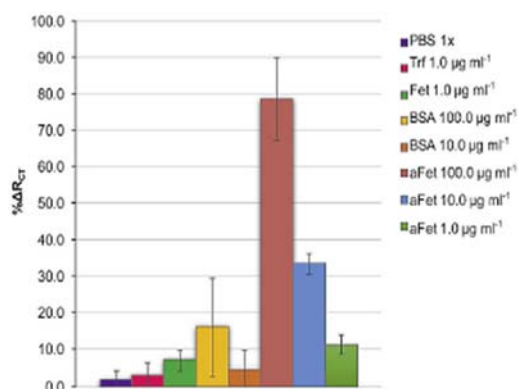


Fig. 2. Responses obtained for diverse glycoproteins incubated with the PNA biosensor for 10 min, at room temperature [6].

2.2. Sample Analysis

The three biosensors were validated through the analysis of serum samples from carcinoma patients (Figs. 3 and 4).

3. Conclusions

All biosensors were constructed following the same general procedure, demonstrating its high versatility. The three biosensors correctly discriminated samples from healthy donors and from

cancer patients with different carcinomas, showing high selectivity towards the antigens STn, Tn and T. Using the three lectin biosensors in the analysis of the same sample allowed to characterize the glycosylation profiles of glycoproteins in the diverse types of carcinomas.

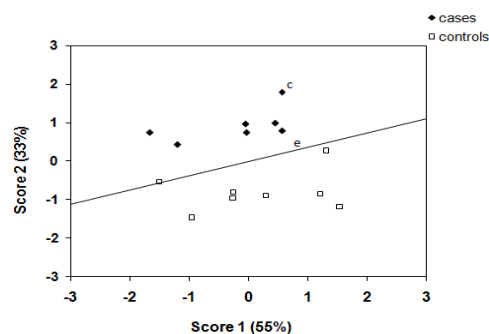


Fig. 3. Graphical representation of the first two scores of a PCA performed on the impedimetric data from sample analysis using the SNA biosensor [4].

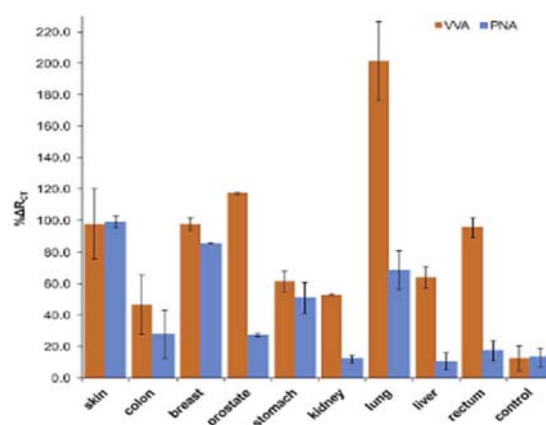


Fig. 4. Results obtained in sample analysis for VVA and PNA biosensors [6].

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(014)

Optoelectronic Nose with a High Chemical Resolution for the Detection of Volatile Organic Compounds in the Gas Phase

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Summary: Artificial noses have a great application potential in various industrial fields. Many of these require to detect a target volatile organic compound (VOC) with certainty in a complex environment, sometimes at a very low concentration. However, the limit of detection and the chemical resolution of today's artificial noses are still far from matching those of their biological inspiration. In this context, our team developed an optoelectronic nose for the detection of VOCs in the gas phase. It relied on an array of peptide sensing materials and surface plasmon resonance imaging. Here, we present different strategies to discover novel sensing materials tailored to the detection of a target VOC with a high selectivity. By combining these novel specific materials with wide-spectrum ones, we were able to discriminate between molecules with very similar structures, differing by only one carbon at low concentration.

Keywords: Electronic nose, Volatile organic compound, Surface plasmon resonance imaging, Odorant binding proteins, Phage display, Virtual screening.

1. Introduction

The detection and identification of odors is a real industrial need. A device that mirrors the sense of smell would address many applications: in the food or cosmetics industry, for quality control or pathogen detection, in the environmental field, or even for medical diagnosis. In this context, a new class of biomimetic sensors is emerging: artificial noses. Inspired by their biological counterpart, they rely on the combination of an array of cross-reactive sensors and a pattern recognition system. They offer a portable, low-cost, easy-to-use analysis tool, which can virtually detect any volatile organic compound (VOC), the main constituents of odors. However, despite recent progress, the performance of artificial noses is still limiting its use. Especially, to adapt to industrial needs, the device must be customizable and allow for the detection of a VOC of interest in a complex and variable environment, even at low concentration [1].

In this context, our team developed an optoelectronic nose based on surface plasmon resonance imaging (SPRi) transduction and biomimetic sensing materials. Here, we evaluated the interest of three different strategies: protein engineering, phage display and virtual screening, for the development of novel high-performance sensing materials, with tunable binding properties. We aimed at improving the chemical resolution of our device, that is to say its ability to discriminate between VOCs.

2. An Optoelectronic Nose for the Detection of VOCs by SPRi

The Fig. 1. presents the optoelectronic nose we have been developing since 2012 for the detection of VOCs in the gas phase.

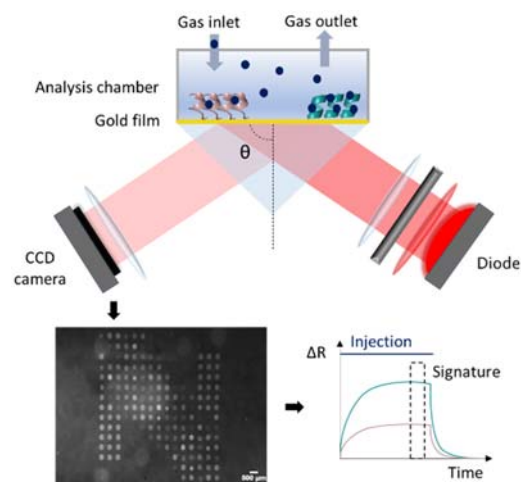


Fig. 1. Working principle of the optoelectronic nose.

Historically, several wide-spectrum sensing materials based on short peptides were responsible for VOC recognition [2]. They were immobilized on a gold-coated prism in a microarray format. The use of a micro-deposition robot made it possible to obtain a high density of sensing materials (300 spots on 64 mm²). The prism was then mounted in an analysis chamber connected to a fluidic circuit. This circuit generated the VOC samples and conveyed them to the analysis chamber in air in a controlled manner. These samples were then analyzed by SPRi. Our SPRi bench is homemade and was the first to work in the gas phase. This transduction method allowed the real-time monitoring of the interaction of the VOC with the various sensing materials simultaneously. All the resulting signals formed the signature of the VOC. The discrimination ability of the optoelectronic nose was assessed by Principal Component Analysis (PCA). The

transfer of our technology led to the creation of a company that has successfully miniaturized and commercialized a portable SPRi for gas sensing.

3. Strategies for the Discovery of Peptides with a High Affinity for a Target VOC

We tested three strategies to discover sensing materials with a high affinity for target VOCs, that could be used in association with the historical ones.

First, we used odorant binding proteins (OBPs) modified by site-directed mutagenesis as sensing materials. OBPs are small, soluble proteins from the biological nose. Their good affinity and selectivity for VOCs, plus their robustness, facilitated their integration to the device. OBP mutants with different binding properties (toward aromatic or aldehyde VOCs) were expressed by *E. coli* and deposited on the SPRi chip. We showed their interest for the trace detection of VOCs in solution [3]. Under controlled conditions (relative humidity >8 %), they also retained part of their selectivity in the gas phase. Peptides inspired by the binding pocket of OBPs could be developed in the future.

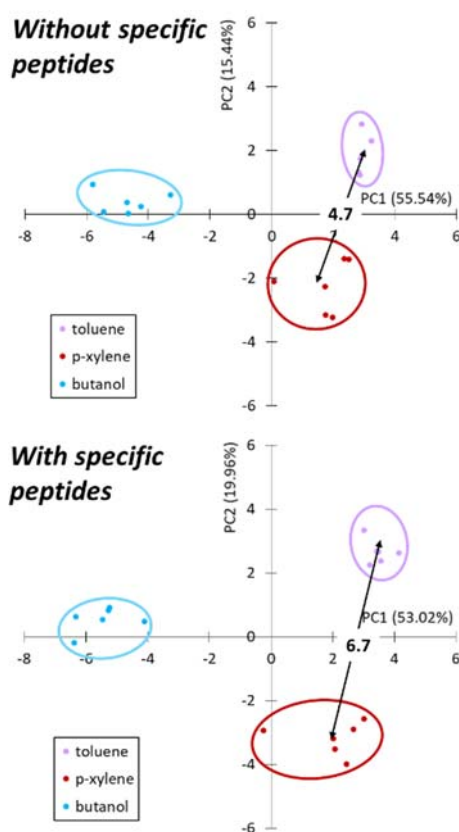


Fig. 2. PCA on different VOC samples with only the historical wide-spectrum peptides and with the addition of specific peptides.

Second, we used phage display, an *in vitro* selection method, to find peptides with a high affinity for aromatic compounds [4]. A commercial library of

M13 phages was screened against a phenyl-decorated surface. Once immobilized on the SPRi chip, the selected peptides increased the chemical resolution of our device, which enhanced the distance between the toluene and xylene clusters on the PCA (Fig. 2). The same technique could be used on other VOC targets.

Third, we evaluated the input of virtual screening for the discovery, the characterization and the maturation of novel sensing materials. With commercially-available softwares (PEP-FOLD, AutoDock Vina), we obtained a good prediction of the experimental affinity between a sensing material and a target molecule. Moreover, a large peptide database could be screened against a wide range of VOCs to identify the most selective binders [5].

4. Conclusions

We developed an optoelectronic nose based on peptidic sensing materials and gas-phase SPRi transduction. This project addressed different solutions to further improve the performance of our system, including its customizability and chemical resolution.

We showed that, under optimized conditions, protein engineering, phage display and virtual screening were all suitable and complementary to discover novel sensing materials to reach this goal.

This work paved the way for the development of the next generation of artificial noses that would meet tomorrow's application needs.

Acknowledgements

The PhD scholarships of C.H., J.S.W. and M.E.K. were funded respectively by CEA, Fdn. Nanosciences, and Labex LANEF and ANR. ANR supported this work (ANR-18-CE42-0012). SyMMES laboratory is part of Labex Arcane (ANR-12-LABX-003) and Labex LANEF (ANR-10-LABX-51-01) programs, both funded by the ANR.

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(017)

Determination of Best Feature Combination for Healthcare Monitoring System Based on Evaluation via Four Distinct Machine Learning-based Methods of Analysis

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Summary: In this study, we used four distinct machine learning-based methods of analysis to evaluate the most effective feature combination for a previously proposed system that detects changes in the physical conditions of human beings using eight sensors placed on a bed and floor beside the bed. First, the proposed system was applied to evaluate the presence or absence of restrictions in the movement of the knee joint in two subjects of different ages, four classes, under distinct scenarios with 10 case of each four classes to construct a set of measurements. Next, 30% of measured feature values were randomly extracted for testing. Scikit-learn was used as the machine learning library, and the k-nearest neighbor method, logistic regression, perceptron, and multilayer perceptron were used as the four learning models. The k-nearest neighborhood method assigned the highest accuracies to all three feature combinations that were investigated in this study. However, sufficient data were not available, and therefore, the accuracies exhibited wide variability, range between 0.33 to 0.75, with an average accuracy of 0.54 for all 10 samples. The low accuracy was primarily attribute to the variations of specific subjects.

Keywords: Pressure sensor, Healthcare monitoring system, Machine learning, Scikit-learn.

1. Introduction

It is important for aged or differently abled people with to be aware of worsening health conditions as early as possible, particularly when they are living by themselves. Therefore, it is beneficial challenge to develop an automatic system that can monitor the health conditions of senior citizens on a daily basis and issue warnings before their health conditions deviate from normal standards. Various type of health and welfare equipment, primarily reliant on the Internet of Things, have been developed over the past years [1-5]. In particular, human motion monitoring and analysis can be used to detect unusual conditions in people with disabilities. Based on this principle, we proposed a system to detect changes in the physical conditions of human beings using eight-sensor-system and reported the outcome of machine learning [6, 7].

In this study, three of the measurements introduced in the works which presence of restrictions in the movement of knee joint were repeated on two subjects of different ages, and four types of machine learning-based methods of analysis were used to determine the accuracies of the three types of measurements. These accuracies indicated the different evaluations of the respective possibilities of detecting the presence of restrictions in knee movement in the subject.

This research was conducted with approval of the Teikyo University of Science's Ethics Committee.

2. Experiment

2.1. Healthcare Monitoring System and Data

As depicted in Fig. 1, the system proposed in our previous report consists of eight flexible force-sensing resistors, a data acquisition device, and a personal computer. One sensor (s1) is placed on a pillow, three sensors (s2-s4) are placed apart from each other at the edge of a bed, and four more sensors (s5-s8) are placed on the floor beside the bed. The output signals were selected from four of the eight sensors.

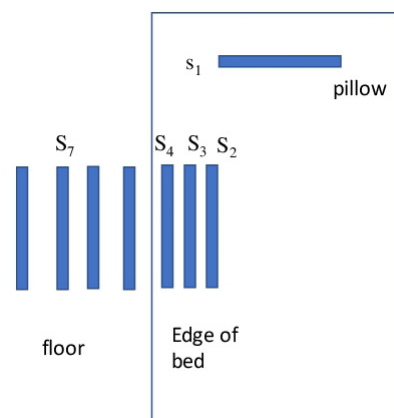


Fig. 1. Position of flexible force-sensing resistors.

Future values t_2 to t_7 , corresponding to sensors s_2 to s_7 , were defined based on the respective time differences between the arrival of the output signals from s_1 , which specified the initiation of the measurement, and the corresponding sensors as depicted in Fig.2. Existing studies have established the best candidate combinations of feature values to be t_2 - t_7 , t_3 - t_9 and t_4 - t_7 [5, 6]. In this study, scikit-learn was used as a machine learning library, and the k-nearest neighbor method, logistic regression, perceptron, and multilayer perceptron were used as classification modules.

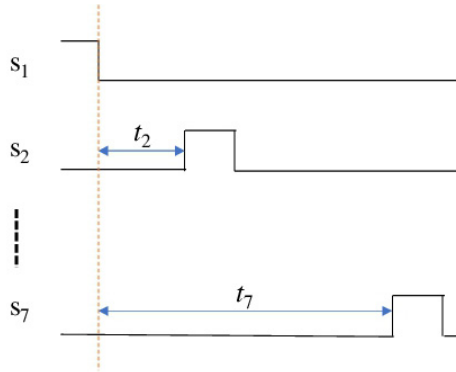


Fig. 2. The definition of the time difference as future values.

2.2. Survey Method

Measurements data were collected by applying the proposed system to two men in their 60s and 70s, under two separate conditions: normal range of motion and simulated limited right knee motion. In all instance, each set of learning and test data was created by randomly sampling 70 % and 30 % of the measurement data as learning data and test data, respectively.

3. Results

3.1. Dependence of Analytical Modes

Fig. 3 depicts in the accuracies of the t_2 - t_7 feature value combination under default setting, as achieved using various machine learning-based methods of analysis. Corresponding to each method, 10 sets of learning and test data were constructed by randomly sampling 70 % and 30 % of measurement data as the learning data and test data, respectively, on 10 separate instances. Subsequently, each analysis was conducted corresponding to each set of learning data and test data, producing 10 accuracy values corresponding to each method of analysis. In the figure, KN, Logi, Perc, and MLP denote the k-nearest neighbor method, logistic regression, perceptron and multiple perceptron, respectively. The k-nearest neighbor method assigned the highest accuracies for all predicted values, with 0.54.

3.2. Accuracies of the Three Feature Combinations

Fig. 4 shows the accuracies of the three feature values combinations, t_2 - t_7 , t_3 - t_7 , and t_4 - t_7 , evaluated using k-nearest neighbor method corresponding to the randomly extracted datasets. The mean accuracies of feature combinations t_2 - t_7 , t_3 - t_7 , and t_4 - t_7 over the sample datasets were 0.54, 0.51 and 0.49, respectively. Thus, according to the k-nearest neighbor method, the t_2 - t_7 feature combination was the most accurate.

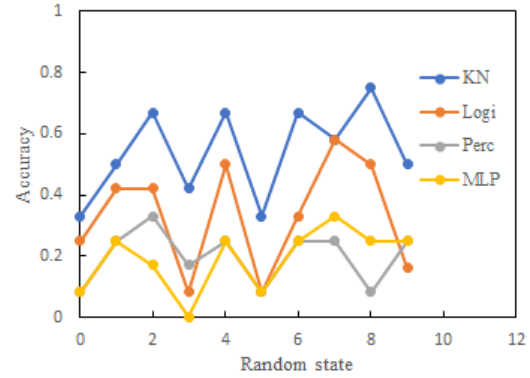


Fig. 3. Difference by analytical model.

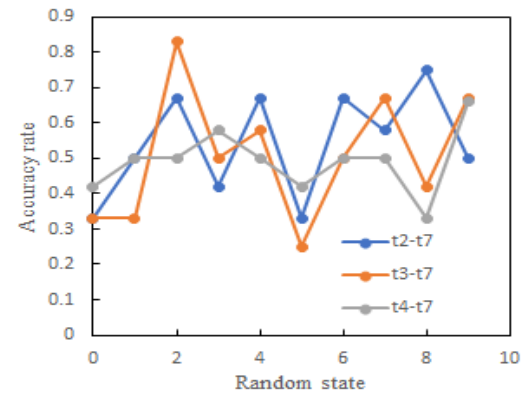


Fig. 4. Difference in accuracy due to differences in feature values.

3.3. Analysis of Accuracy Distribution

Because the three types of combinations for the feature values demonstrated a significant variation in accuracy for the test data, factors affecting the accuracy were investigated. At this stage, t_2 - t_7 was considered. First, we examined the extraction rate of the data used for the tests. Although a value of 30 % is commonly used, we adopted a value of 20 % to increase the number of learning data because of the measurements, and we also adopted a value of 40 % because of the substantial variability of the data.

Fig. 5 illustrates the three extraction rate accuracies evaluated by the k-nearest neighbor method corresponding to randomly extracted datasets. Subsequently, each extraction rate showed almost same trend.

The dependence of the percentage of the test data on accuracy, including standard errors, is shown in Fig. 6. The mean values of accuracy corresponding to the test data selection percentages of 20 %, 30 %, and 40 % are 0.68, 0.54, and 0.58, respectively. In the case of 20 %, there were cases in which the extracted test data did not include data points corresponding to the subject's conditions. It was concluded that it was better to adopt data by extracting 30 % as a test data from the relationship between accuracy and standard error.

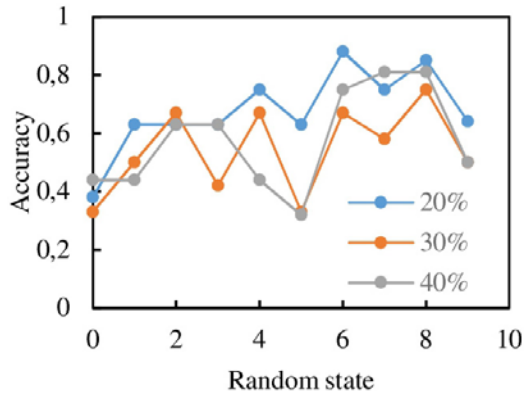


Fig. 5. Three extraction rate accuracy.

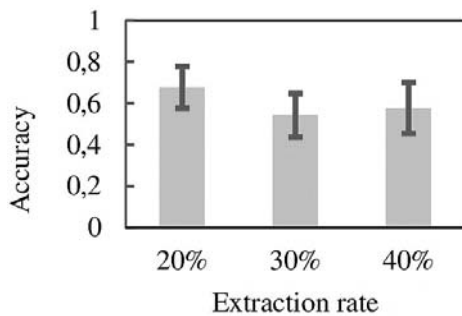


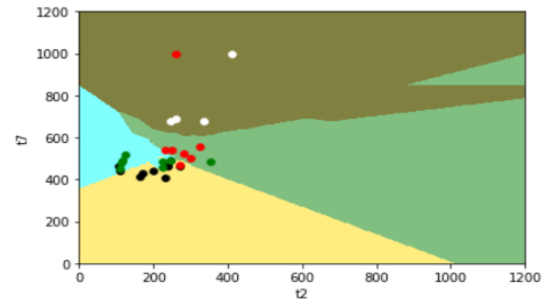
Fig. 6. The dependence of the extraction rate.

3.4. Analysis of Accuracy Using 2-D Distribution

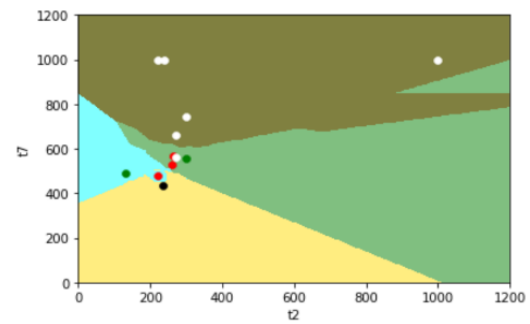
For future value combination t_2 - t_7 , we investigated the main causes for high and low accuracy using a two-dimensional distribution diagram. First, the distribution of points corresponding to each exercise limitation condition was examined for the high-accuracy case. Fig. 7 depicts the case where a high accuracy of 0.75 is achieved. In the learning data, subject A, represented by the black circle for normal range motion and the green circle for simulated limited right knee motion in the figure, demonstrated no specific variation in the presence or absence of motion limitation. For subject B, represented by the red circle for normal range motion and the white circle for simulated right knee motion demonstrate no specific variation.

Fig. 8 shows the learning data for low accuracy for features t_2 - t_7 , t_3 - t_7 and t_4 - t_7 . We found that the shape

of the segmented area differed significantly depending on the feature values. Furthermore, in this case, we believed that the difference was largely dependent on the variation of the data for the unrestricted red and white circles of subject B. We observed that the segmented area differed considerably because of the difference in the feature amount and depended significantly on the accuracy of the data.

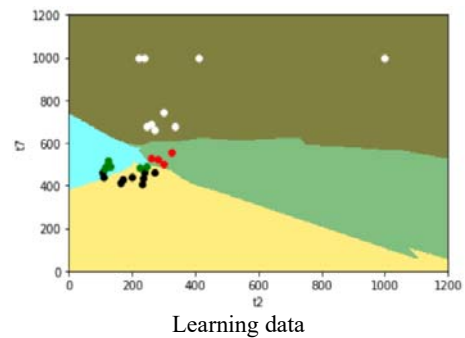


(a) Learning data

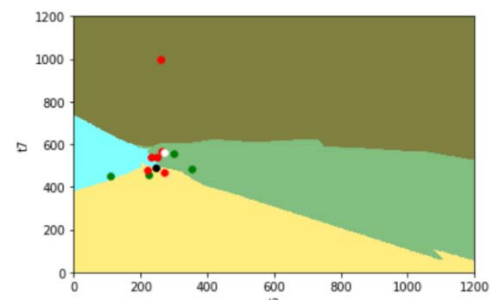


(b) Test data

Fig. 7. High accuracy case.



Learning data



Test data

(a) t_2 - t_7

Fig. 8 (a). Low accuracy for learning date of three feature values.

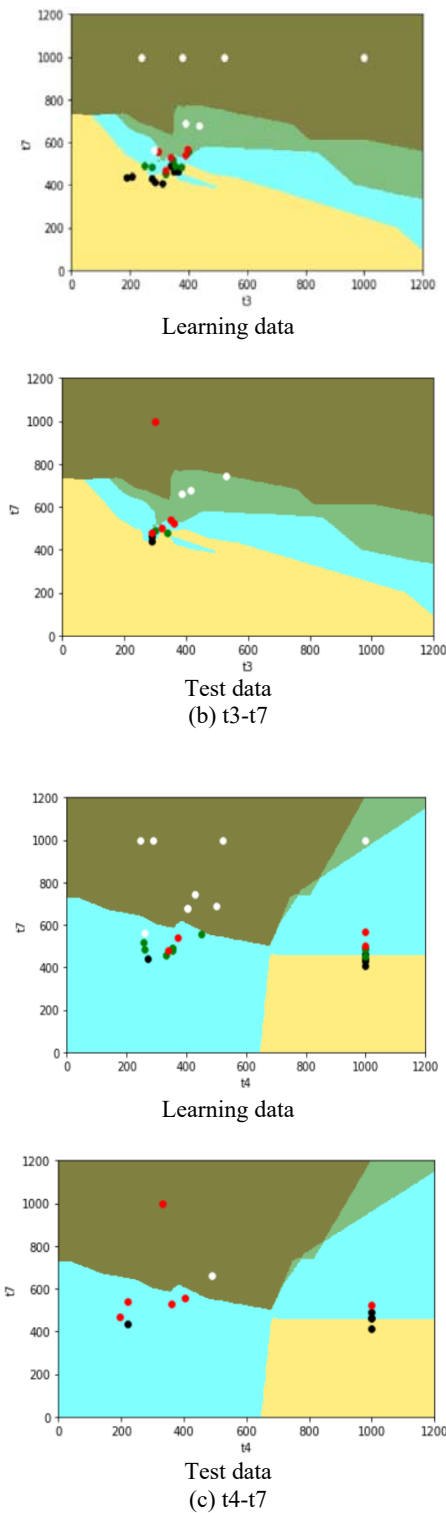


Fig. 8 (b-c). Low accuracy for learning data of three feature values.

4. Conclusions

We evaluated the performance of a system to detect changes in the human health conditions using four distinct machine-learning-based methods of analysis.

First, the aforementioned system was used to monitor two subjects of different ages under two separate scenarios -no exercise restriction and restricted exercise for the right knee- obtain a set of measurements. Based on multiple learning and test datasets constructed using the recorded measurements, it was determined that the k-nearest neighbor method attributed the highest mean accuracy of 0.54 to the t2-t7 feature combination. The accuracies corresponding to all 10 sets of training and test data exhibited considerable variability -ranging between 0.33 and 0.75- owing to an insufficient quantity of training data and the lack of optimization via other methods. Further analysis of low accuracies demonstrated the significant impact of specific variability on accuracy. It cannot be applied to actual patients owing to the influence of COVID19. The effectiveness of this system can be demonstrated in the future by integrating the pressure sensor system and the health condition detection system incorporating machine learning.

Acknowledgements

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(018)

R-testbench: A Python Library for Instruments Remote Control and Electronic Test Bench Automation

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Summary: Numerous applications require instruments remote control and electronic test bench automation. Software solutions exist but are either proprietary or low-level programming. In this paper, a high-level open-source software solution, R-testbench, is presented. It is a Python package that allows to manage and control instruments remotely and to fetch data for saving and post-processing. The communication with instruments is built on the well-known and reliable Virtual Instrument Software Architecture (VISA) standard. The main advantages of the proposed solution are the high-level interface and the automatic instrument recognition, which facilitate the user experience. The Python package has been validated by performance characterization, unit tests, and by application to a real use case involving precise current measurements in the framework of a feasibility study in the nuclear domain.

Keywords: Remote control, Test bench, Instrumentation, Automation, Software, Python, VISA.

1. Introduction

Critical applications, such as experimental research involving nuclear material or precise characterization of integrated circuits, require remote control of measurement instruments and electronic test bench automation for safety and/or accuracy reasons.

In this respect, the most popular proprietary tools that provide instruments control and automation are LabVIEW [1] and MATLAB/Simulink [2]. Both offer (partial) graphical programming, but scripting is based on their own programming languages, which are less known and practiced than generic purpose languages such as C or Python. On the open-source side, Octave and Scilab, which are MATLAB-like programs, have both a toolbox dedicated to instruments control [3, 4]. The main drawback is the low-level programming, which requires experience from the user. Finally, a Python package, PyMeasure, is an existing solution that tries to offer high-level programming [5], but lacks of properties such as automatic instrument recognition.

When comparing the main features of the existing software solutions for instruments remote control and test bench automation, one can observe that the major constraints are (1) the price, (2) the low-level programming, and (3) the need to learn a proprietary language. One solution to solve these problems consists in developing a Python-based automation tool. Indeed, Python has experienced a quick development over the past years, especially for numerical computation, data mining and artificial intelligence [6, 7]. Hence, it can be considered for both academic and industrial technical solutions.

Considering the needs for a high-level open-source tool offering automatic instrument recognition, we have developed R-testbench, an open-source Python package that allows to create a software remote test

bench to control distant instruments from a computer and automate a complete electronic test bench. The proposed library relies on the Virtual Instrument Software Architecture (VISA) standard [8], which enables to communicate with instruments by software through many types of interfaces (e.g., GPIB, USB), and that is now implemented on most operating systems (OS). Our approach aims to reach at least the same performance and universality as the aforementioned solutions, while offering a high-level interface as well as improved automation capabilities.

This paper is organized as follows: Section 2 describes the software architecture of the proposed Python package, Section 3 presents results that validate the package for real use case, whereas Section 4 summarizes the key points.

2. Software Architecture

The object-oriented paradigm has been used to develop a package which offers a unique manager class, a generic instrument interface, and automatic instrument recognition.

In addition, to make the proposed package robust and reliable, the main Application Programming Interfaces (APIs) available for scientific computation have been used.

2.1. VISA Communication API and Automatic Instrument Recognition Implementation

The manager class is named RTestBenchManager. Its main objectives are (1) to embed the VISA protocol detecting and connecting instruments to the software testbench, and (2) to manage all connected resources regarding communication channel access, memory

allocation and tasks queuing. The VISA libraries are accessed through the PyVISA Python package [9].

As shown in Fig. 1, the base class to control an instrument is Tool, which defines the basic functions required for all instruments, such as send() for configuration or query() to get data. Then, one interface per type of instrument, defining the main functions, is created as an abstract class (a class from which no object can be instantiated), inheriting from Tool. This architecture allows to implement the automatic recognition of instruments by using polymorphism, i.e., an access to a specific class through the generic mother class. The user experience is then facilitated as the interface is always the same, whatever the brand and the model of the instrument.

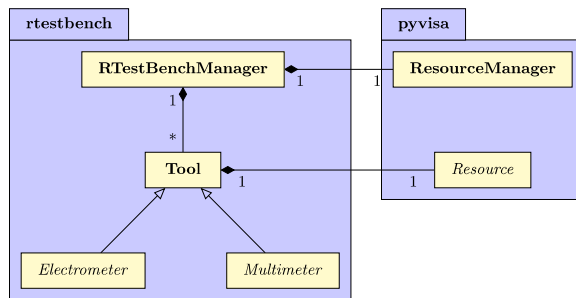


Fig. 1. R-testbench Unified Modeling Language (UML) class diagram: manager, tool interfaces mechanism, and interactions with PyVISA.

2.2. Data Containers APIs and Data Management Implementation

All data gathered by R-testbench can be stored by using Python built-in classes, such as lists, or by using scientific computations libraries such as NumPy and pandas. This provides flexibility and possibility to use any data mining or processing framework.

For efficiency reasons, the default data containers used to store data coming from any Tool is the well-known NumPy's N-dimensional array: ndarray [10]. It is the current best compromise between the memory and speed efficiencies of C/C++ and the high-level and user-friendly data manipulation of Python.

For its ease of use and its rich interface, the pandas API [11] is used to gather data all together and to save them to a specific type of file. R-testbench can save data either as human-readable or binary files. Currently, human-readable files types comprise the largely used CSV, which stands for comma-separated values. On the other hand, supported binary files formats are the common Python pickle, as well as HDF5, that is extensively used for big data and artificial intelligence. Other files types are supported by the pandas API, such as the more and more popular JSON, can be used to extend R-testbench in the future.

3. Validation

The R-testbench package has been used to carry out measurements in the frame of experiments that

required real time monitoring of semiconductor devices irradiated by β -rays, as described in [12]. Taking advantage of these experiments, the software has been intensively tested together with the B2985A electrometer from Keysight [13]. In addition, automated unit tests have been implemented by using the unittest and PyTest modules. Finally, the package has been made publicly available on GitHub (<https://github.com/Ark42/rtestbench>), in order to be shared with and reviewed by the open-source community.

All of these constitute the validation tests of the proposed R-testbench package. Details are presented further in this section.

3.1. Performance Measurements

The main measurements to characterize the performance of the proposed software are: 1) the time required to fetch data from the instrument to the computer, 2) the time required to save data locally to a file, and 3) the size of the file. All measurements have been done on a 64-bit computer, Windows 10 Pro OS, 16-GB RAM, Intel Core i7-8550U CPU.

The Keysight B2895A can be controlled by USB and LAN (TCP/IP), among others. The maximum data transfer speed of the USB interface is 488 Mbit/s, whereas it is 100 Mbit/s for the LAN interface (Ethernet cable) [13]. Each datum can be stored either on 32 or 64 bits.

As expected, the transfer through USB is faster than by TCP/IP (Fig. 2). This is, of course, explained by the limitations due to the interfaces of the electrometer. Moreover, data transferred by TCP/IP are embedded with several addresses (MAC, IP), which increase the number of bits to be sent. Data stored on 32 bits are also fetched faster than data stored on 64 bits, which is expected for the transfer speed but could be jeopardized by the fact that the computer operates in 64 bits.

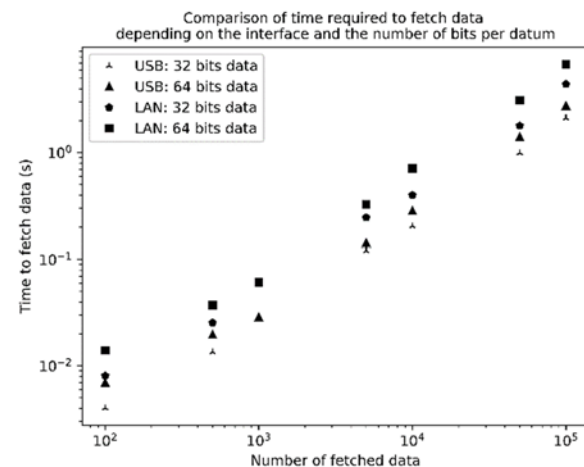


Fig. 2. Comparison of time required to fetch data from the Keysight B2985A electrometer to the computer. Data are stored locally in a numpy.ndarray container.

In addition, the fetch time increases linearly with N , the number of data. The linear regression performed on the four sets of measured data (Fig. 2) provided the following equations:

$$t_{\text{fetch,USB } 32} = 0.0204 N + 2.1, \quad (1)$$

$$t_{\text{fetch,USB } 64} = 0.0281 N + 4.8, \quad (2)$$

$$t_{\text{fetch,LAN } 32} = 0.044 N + 4.6, \quad (3)$$

$$t_{\text{fetch,LAN } 64} = 0.064 N + 6.7 \quad (4)$$

Times are given in milliseconds.

As explained in Section 2, the pandas API is used to save data to files, locally. The time required to save data (Fig. 3), as well as the file size (Fig. 4), have been studied with an increasing number of data.

As expected, CSV is the worst format, considering both speed and memory, especially when the number of data increases. However, if the number of data to save is lower than 1,000, the HDF5 is surprisingly less efficient. A possible explanation might be that this file format has been created to target big data applications and could suffer from a lack of optimization for a small amount of data.

Regarding the binary files, the pickle format seems to have better performance than the HDF5 format. This assertion is true on the tested range of data number. Considering the speed, it seems that the gap tends to reduce for large sets of data (Fig. 3). Conversely, the evolution of the size seems to be linear, except for small datasets.

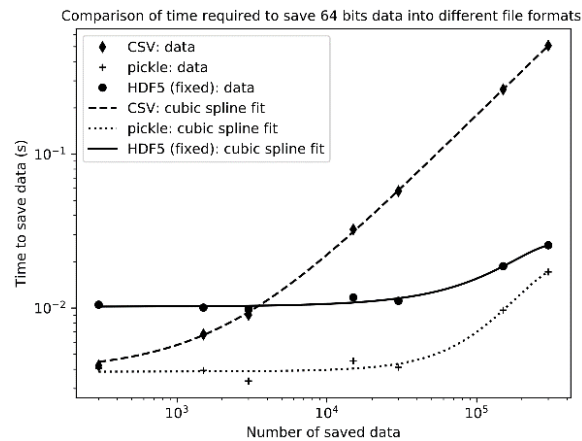


Fig. 3. Comparison of the evolution of the time required to save data into CSV, pickle and HDF5 files formats with the number of data. Data are saved to files using pandas.

3.2. Continuous Integration Automated Testing

Reliability is of utmost importance in instrumentation. This is the reason for which the proposed R-testbench package has been built on top of industrial protocols, such as VISA. In order to achieve the same objective, continuous integration and

automated tests have been deployed together with the package. Both rely on the well-known PyTest library and GitHub workflows.

The functions of R-testbench are tested extensively (different conditions, success and fail cases) by creating unit tests. Each test ensures that a function behaves as expected for a specific condition. Once a test passes, it should never be modified, so that the user application is guaranteed to continue working even if the development of the package goes further.

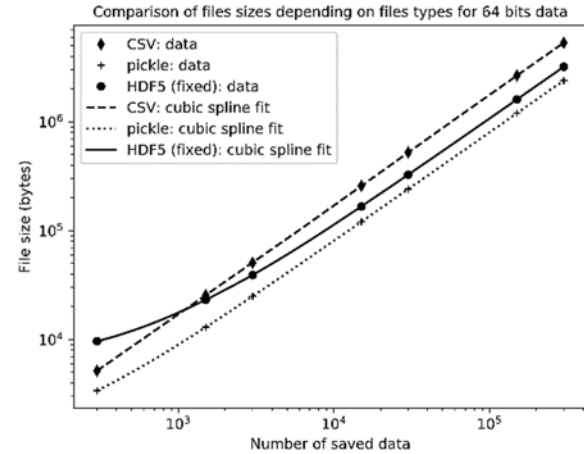


Fig. 4. Comparison of the evolution of sizes of CSV, pickle and HDF5 files with the number of data. Data are saved to files using the pandas API.

Currently, more than 80 % of the code is covered by tests. The remaining 20 % is mainly the part specific to an instrument, for which automated testing is difficult to implement. Nevertheless, this part has been covered by the performance measurements exposed in Section 3.1, as well as during the experiments conducted in [12]. Hence a high confidence level in the proposed tool.

3.3. Open-source Community

R-testbench is publicly available on GitHub, an open-source platform used by researchers and developers from industry and academia to share and contribute to open-source projects. It is licensed under the Open Software License v. 3.0, a copyleft license that emphasizes the possibility to share the code.

There is currently one contributor and maintainer: A. Quenon. As the package has been made publicly available for the SEIA' 2020 conference, the interest for this project is expected to grow up.

4. Conclusions

R-testbench is a Python package that allows to control instruments remotely and to automate test benches. Its main advantages over other available solutions are the open-source share and the automatic instrument recognition (Table 1). It does not support

communications by User Datagram Protocol (UDP), but this is not really important for instrumentation as UDP provides a connection less communication model

and does not guarantee the packet delivery, so the quality of the data transfer. However, it minimizes the latency during the data transmission.

Table 1. Main characteristics of software solutions for measurement instruments remote control and automation.

Tool	License	Programming			Compatible protocols	Automatic recognition
		Level	Language	Graphical		
LabVIEW	Proprietary	all	MathScript	yes	all	no
MATLAB	Proprietary	low	Matlab	partial	all	no
Octave	GPL	low	Octave	no	no PXI	no
Scilab	GPL	low	Scilab	no	no UDP	no
PyMeasure	MIT	high	Python	no	no UDP	no
R-testbench	OSL	high	Python	no	no UDP	yes

The library has been tested and validated on a real critical use case, i.e., experiments involving nuclear materials. Several parameters have been measured to test the performance of data transfer speed and data saving speed and memory. Other performance parameters, such as the memory size necessary during execution time, can be monitored in later studies.

The project is now awaiting for reviews by the open-source community to be improved.

Acknowledgements

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Optical Design Approach of a Fully Integrated Miniature Spectrometer

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Summary: Optical simulation software, which is based on the ray tracing method, offers fast results in the imaging optics. This simulation method can also be applied in other fields of light propagation, such as non-imaging optics for design of optical sensor or metrology elements.

In this paper a design approach of a fully integrated miniature spectrometer by means of ray tracing simulation software is demonstrated. The basis of the presented element is a Rowland spectrometer that is optimized in crucial points, e.g. shape of mirror and grating, to guide the maximum of light. The classical Rowland spectrometers exhibit a flat design in one dimension, leading to a bad signal-to-noise ratio especially for high numerical aperture sources. Therefore, this spectrometer is designed to guide the complete light in the sagittal and tangential plane, see Fig. 1.

Further on, a monolithic approach is presented with a blazed grating based on an aspheric mirror to spectrally separate and focus light on the image layer. The aspheric mirror is designed in a way that most of the aberrations are suppressed. In general, the element should be designed in a way that it can be produced with a mass production technology like injection molding in order to offer a reasonable price. The paper will introduce the application fields of miniature spectrometers and will step by step describe the development of this fully integrated monolithic miniature spectrometer by means of ray tracing simulation.

Keywords: Raytracing, Spectrometer, Optical sensor, Visible spectrum, Miniature, Injection molding

1. Introduction

WDM, wavelength division multiplexing, has the power to increase the data rate of standard POF, polymer optical fibers, communication dramatically [1-14]. Therefore, it is essential to develop an inexpensive (de-) multiplexer component [15]. In this paper development steps for this component based on a Rowland spectrometer are described.

Several preconditions must be fulfilled to create a functional (de-) multiplexer for POF. First, the divergent light beam, which escapes the POF, must be focused. This is done by an on-axis mirror. In the first attempt, a spherical mirror is used. To get perfect results without any spherical aberrations, an ellipsoid mirror should be preferred.

The second function is the separation of the different transmitted wavelengths. This principle is illustrated for three wavelengths (red, green, blue) in Fig. 1, respectively. This is not a limitation for possible future developments, but rather an experimental basis from where to run the various simulations described below. The diffraction is done by a diffraction grating. The diffraction is split into different orders of diffraction. The first order is the important one to regain all information. A detector line can be installed there to detect the signals.

Because the grating is attached to a bended basement only one element can cover both functions, focusing and diffracting. Hence the light is not afflicted with any aberrations or attenuations of a focusing lens or other diffractive elements, which are necessary for any other setup.

Another unique characteristic of the presented key element for POF communication is the three-dimensional approach. Key elements of glass fiber communication are usually designed planar. This simplification cannot be adopted for POF communication, because of the large Numerical Aperture and therefore large opening angle of the POF. A planar setup for POF key elements would lead to high losses.

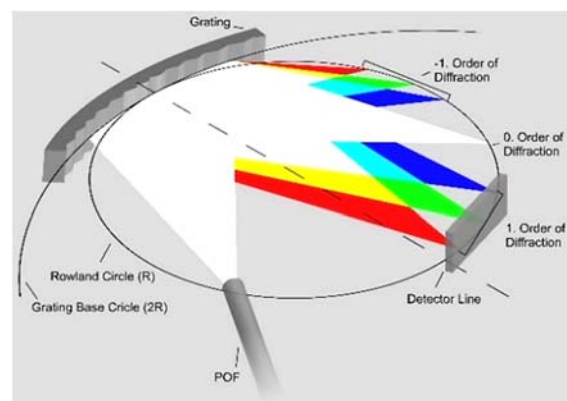


Fig. 1. Principle sketch of a Rowland Spectrometer.

2. Results of the Simulation

In the following steps, a software program is used to design a demultiplexer based on the general concept of a Rowland Spectrometer. For the current task, the software OpTaLiX provides all required functionalities [16]. This approach offers different

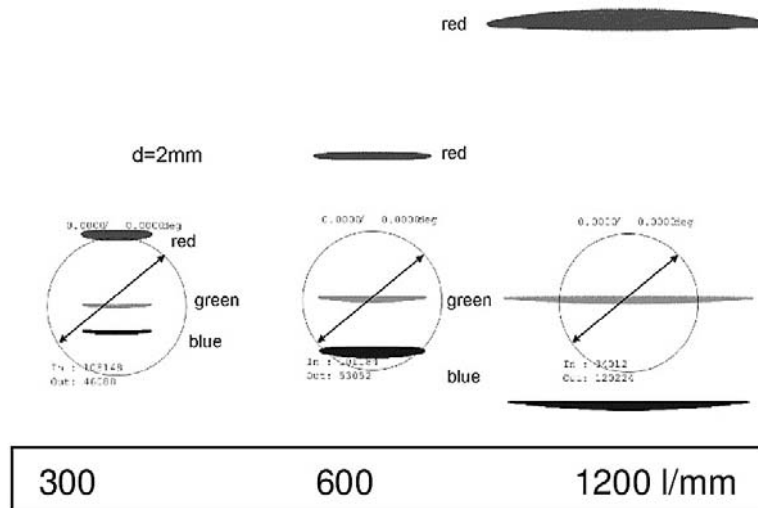


Fig. 4. 2D Plot of the demultiplexer with a spherical mirror.

The TRA (Fig. 5) shows a slight defocusing for the meridional (tangential) section, but a strong defocusing for the sagittal section. The graph of the function in the meridional section exhibits a predominant third order Seidel coefficient. Therefore, the slight defocusing in the meridional section compensates the astigmatism. The OPD (Fig. 6) shows as expected strong deviation from the ideal waveform, especially in the sagittal section. This defocusing leads to high losses for the coupling efficiency for the POF- or detector- array in the image layer.

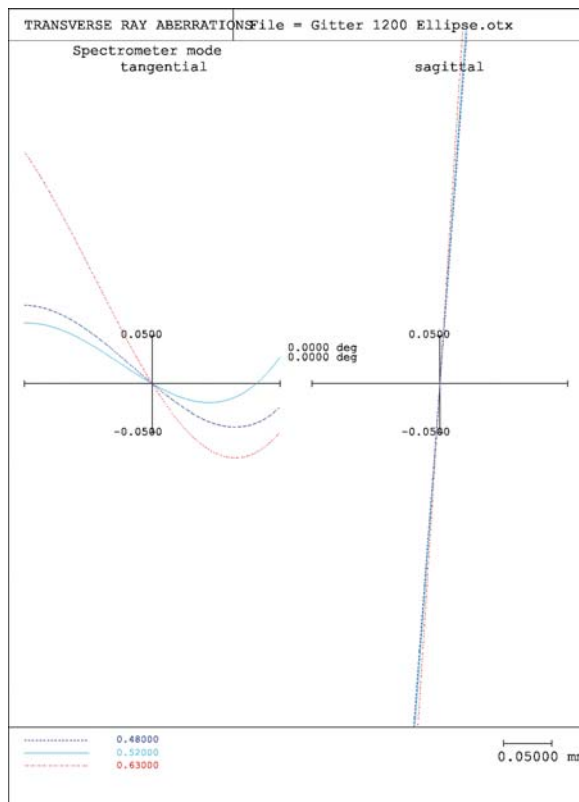


Fig. 5. TRA for the ellipsoid demultiplexer with 1200 lines/mm.

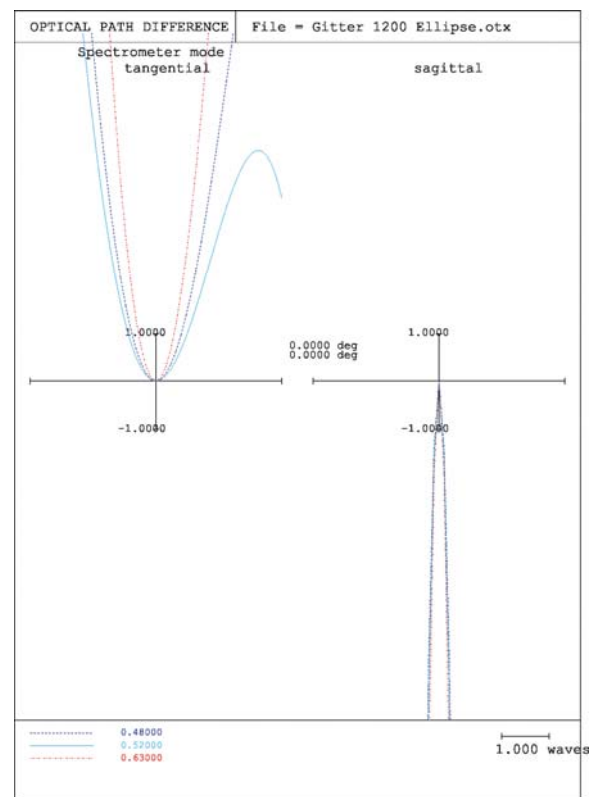


Fig. 6. OPD for the ellipsoid demultiplexer with 1200 lines/mm.

It is obvious that the grating changes the focal length especially of the sagittal section; therefore, the shape of the mirror must be improved. It is necessary to change the radius of curvature notable in the sagittal section. Hence the basic shape of the mirror is no longer a sphere or ellipsoid. To meet the demands a higher order shape, which is nearly cylindrical, is used.

The change of the mirror shape improves the imaging quality substantial. The Spot Diagram and the TRA for the improved demultiplexer are shown in Figs. 7 and 8.

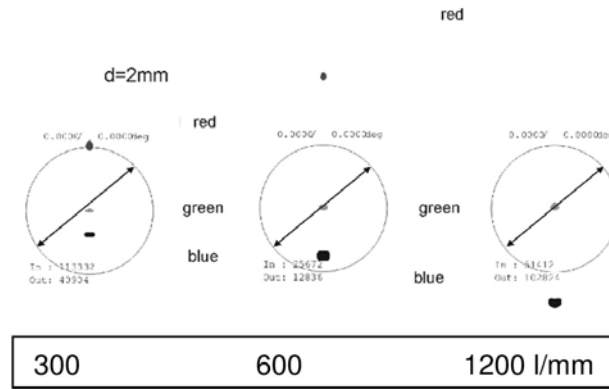


Fig. 7. Spot Diagram (circle diameter 2 mm) for the improved demultiplexer.

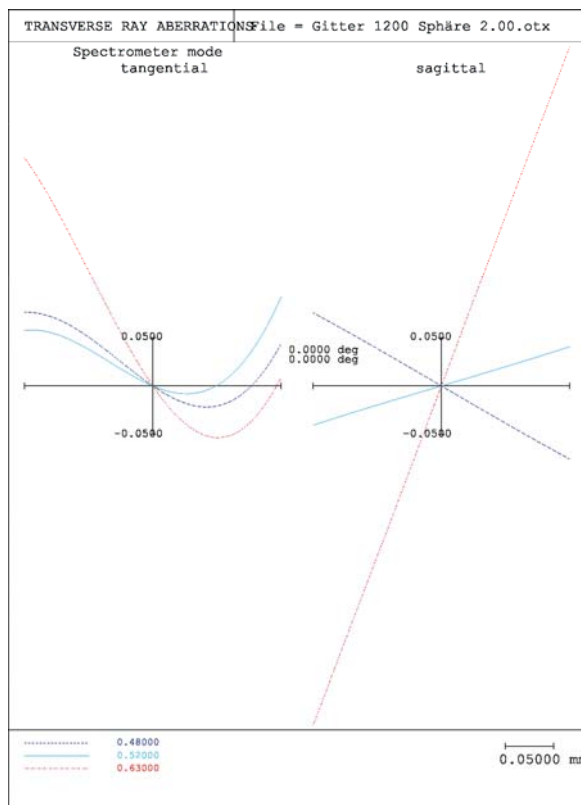


Fig. 8. TRA for the improved demultiplexer.

The Spot Diagram shows three dividable colors for all line densities. The gap improves with the line density and at 1200 lines/mm the gap between every color is larger than 2mm. The TRA shows a marginal shift of the focus of all wavelengths to offset the astigmatism in the meridional section. Because of the spectrometric function of the demultiplexer it is not possible to focus all three colors simultaneously. There is always a combination of over and under correction for the different colors. Hence the radius of the mirror in the sagittal section is optimized to focus the colors in whole as much as possible.

This improved spectrometer can separate three colors with enough space between them to regain the information with a POF- or detector-array. Hence it

can be used as a demultiplexer for POF communication. The shapes of the foci feature low coupling losses and the shape of the mirror should be easy to produce by injection molding.

3. Conclusions

State of the art for POF communication is the use of only one single channel. This means a limitation of bandwidth. The solution for this bottleneck is WDM over POF, as not only one channel is used to transmit information over a single fiber. To use this technique two key elements must be designed completely new: a multiplexer and a demultiplexer, because the key elements of the established WDM for glass fibers in the IR cannot be applied.

The simulation results show that it is possible to build up a demultiplexer by means of a Rowling Spectrometer. A special shape of the mirror is required to suppress most of the aberrations caused by the grating. The improved demultiplexer can separate all three colors with a gap of 2mm and a crosstalk of less than 30 dB. So, in combination with injection molding this configuration can be produced with sufficiently costs. This spectrometer has the chance to increase data rate of standard POF communication also with a broad range of usability in optical spectroscopy for sensor systems in automotive and medical applications due to its low-cost realization.

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Model Based Signal Processing for Angle Measurement with a Magnetoresistive Sensor Array

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Summary: Robust magnetic sensors are given preference regarding automotive mechatronics. Innovative sensor technologies and signal processing concepts are the key to increasing requirements for magnetic sensors. This paper shows a signal processing concept for magnetic sensor arrays using the tunnel magnetoresistive effect (TMR) with the very recent TMR-Vortex design. The effect is used in separated dots with a size of 1 μm . Many of these dots can be arranged as a matrix on a chip. The matrix can be combined with two-dimensional signal processing approaches. For example, the system can act as an angle sensor. A dedicated signal processing is demonstrated. It consists of a training phase for teach-in parameters and an operation phase using the learned parameters. This method is able to compensate mounting and production tolerances.

Keywords: Automotive magnetic sensors, Sensor array, Two-dimensional signal processing, TMR-Vortex Sensors.

1. Introduction

More than ten magnetic sensors are used in modern vehicles. For example, the angle detection for throttle of a combustion engine or the rotor position for electric motors are typical mechatronic applications [1]. The sensor detects the position of the magnetic field of a rotary magnet, see Fig. 1. Recently, automotive sensor development is driven by increasing reliability and accuracy requirements. In contrast, large mechanical tolerances must be allowed for mass production at a very low cost. Currently, the focus is on quality management. In the future, new sensor concepts and the use of digital signal processing will increasingly complement this.

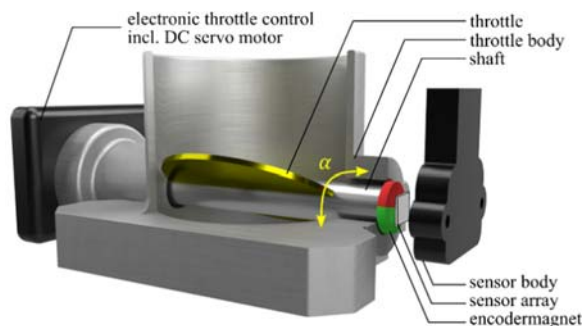


Fig. 1. Example of the potential application of a sensor array as a throttle position sensor. The encoder magnet in the end-of-shaft-position indicates the angle information.

Magnetic sensor arrays are known for biomedical applications [2] or specialized measurement equipment [3, 4]. This paper explains how the TMR-Vortex technology [5, 6] can be used as a sensor array structure. We propose the array concept to support the automotive requirements. The array arrangement in combination with suitable signal

processing achieves remarkable tolerance compensation capabilities. The necessary processing parameters can be trained by reference data in a teach-in procedure.

2. Simulation and Experimental Setup

Most angle sensors only have two output signals (cosine and sine). In contrast, matrix-shaped sensor arrays offer a large number of output data, which form a vector magnetic field image. The pixels of this image are output values of TMR-Vortex dots. Each pixel represents the X-component (H_x) and Y-component (H_y) of the magnetic field. The dots are insensitive in the Z-component.

We generate the vector images by calculated synthesis and with an experimental setup. The synthesis follows two steps.

First, the simulation of magnetic fields is done for all positions of the pixels. The simulation takes into account the rotation of the encoder magnet, see Fig. 2. The result of the simulation are vector fields. An example is shown in Fig. 3.

Second, the magnetic data are the input of characteristic maps. The maps contain two-dimensional data measured from a TMR-Vortex sensor. This method works with a computer controlled cross coil setup [7, 8]. The results of real sensor devices are stored as two dimensional arrays which are recorded with an ADC resolution of 12 Bits. The dimensions of the array are the vector components of the magnetic field H_x and H_y . The completely filled arrays represent the characteristics of the sensor including nonlinearities. The next step is the simulation of the pixels in the sensor array. A magnetic field simulation based on the magnetic dipole equation is performed for each pixel. As result of the local

magnet field parameters the arrays provide the voltage values for all pixels outputs. This effective combination of measurement and simulation generates “synthetic” – but realistic – input values for the presented investigations.

Additionally, we validate the synthetic results with experimental setups. Here discrete conventional sensors are placed as an array on a PCB. These setups are upscaled 25:1 as functional models compared to future integrated arrays. This is shown in Fig. 2 right.

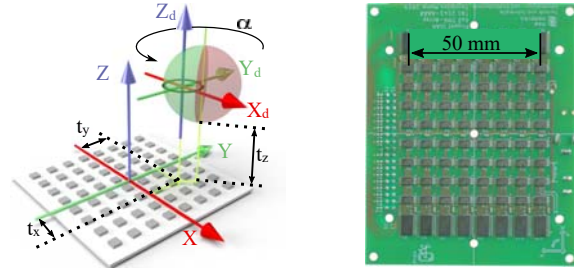


Fig. 2. Left: Example of an 8×8 array arrangement with 64 sensor pixels in combination with a magnetic dipole as rotary encoder magnet. To demonstrate tolerance problems the dipole is shifted from the origin axis. Right: Scaled validation model with discrete TMR-Vortex sensors.

As a simple approach for the simulation of a spherical encoder magnet the well-known magnetic dipole equation is used [7]. Fig. 3 shows an exemplary result of a 2×2 mm² large array with 8×8 pixels. The rotating axis of the encoder magnet is shifted by $t_x = t_y = 0.5$ mm out of the center of the array. The center of the magnet has a distance of $t_z = 2$ mm. Besides the representation as a vector field, the cosine and sine components are plotted three-dimensionally, which emphasizes the inhomogeneity of the detected encoder field.

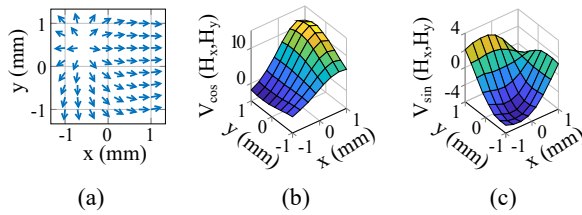


Fig. 3. Synthesized output data from the array: Vector field (a) from the cosine matrix data (b) and sine matrix data (c) in mV/V.

3. Model Based Signal Processing

For processing of the vector matrix data we propose a model based method. The method operates in two stages – angle extraction and displacement error correction. The required model parameters are derived during a training phase and are used for tolerance error compensation in an operation phase.

For training and operation phase a polynomial regression forms the pre-processing of the array data in the same way. After this pre-processing step, the two phases differ. In the training phase the position of the

misaligned axis is calculated by the first partial derivative. The result (x_0, y_0) is stored for the operation phase. The angle is determined for this position. The model parameters are calculated by the known reference angle and the components (x_a, y_a) of the angle.

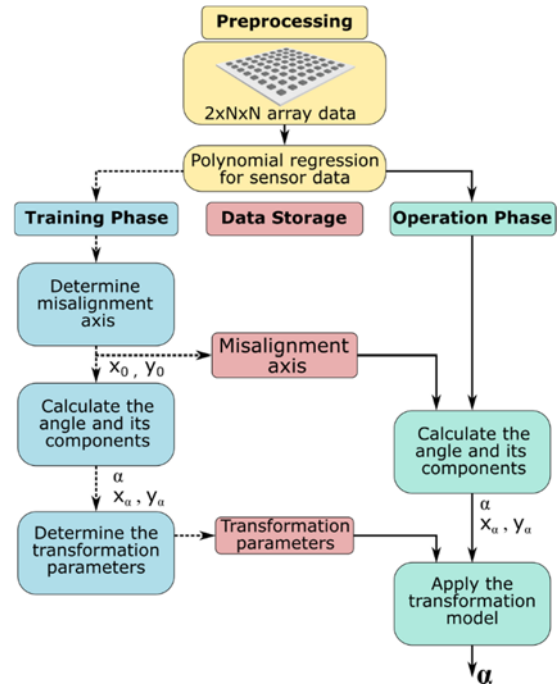


Fig. 4. Overview of the processing steps of the training and operation phase. The pre-processing step is identical for both phases. In the last steps the appropriate model is used (projective or polynomial transformation).

Jointly with the known reference angles the model parameters are derived and stored for the operation phase. We favorite two transformation models: the projective transformation with the projection matrix as parameter or alternatively the bivariate polynomial with their coefficients as parameters.

The steps of the operation phase after the pre-processing are, firstly, angle calculation for the predetermined axis position (x_0, y_0) and, secondly, the application of the model and its stored parameters. As before, the projective transformation or the polynomial transformation apply respectively.

3.1. Pre-processing Angle Extraction

The desired angular position of the magnet can be calculated differently based on the array data. In the simplest way, this angle is determined from the mean values of the two vector components of all pixels. However, this does not solve the problem of misalignment. Furthermore, the calculation with discrete derivations of the pixel values is possible, whereby problems of digital resolution and strong boundary value effects can occur.

We propose to avoid these problems by approximating the bivariate polynomial (1) to the sensor data.

$$f(x, y) = a_0 + a_1x + a_2y + a_3x^2 + a_4xy + a_5y^2 + a_6x^3 + a_7x^2y + a_8xy^2 + a_9y^3 \quad (1)$$

The dipole is located at the position where the orientation of the vector field is homogeneous. The calculation is done by searching the minimum of the sum of the amounts of gradients G in X- and Y- direction, where G_{cx} , G_{cy} are from the cosine matrix and G_{sx} , G_{sy} are from sine matrix.

$$\min\{x_0, y_0\} = \sqrt{G_{cx}^2 + G_{cy}^2} + \sqrt{G_{sx}^2 + G_{sy}^2} \quad (2)$$

The angle of the vector field can be determined by the second partial derivatives.

$$\tan(\alpha) = \frac{\partial^2 V_{\sin}(x_0, y_0)}{\partial y^2} / \frac{\partial^2 V_{\cos}(x_0, y_0)}{\partial x^2} = \frac{y_\alpha}{x_\alpha} \quad (3)$$

Equation (3) delivers sufficient results in simulations. For practical use the second derivative is very sensitive to numerical resolution and influenced by non ideal behavior of the sensor elements. Therefore, an additional correction method is proposed.

3.2. Compensation through Data Transformation

After angle extraction, gain and offset errors, phase errors and higher harmonics can still occur. The harmonics, in particular, impair the angular accuracy of the sensor [9]. For correction, a transformation towards the ideal encoder-sensor system is carried out [10]. This has in the X-Y-representation a trajectory of the unit circle. In the training phase, the parameters of the transformation model are calculated by assigning measurement data to reference values, see Fig. 5.

Two models for transformation are presented.

As a first model, the projective transformation can calculate translations, rotations and shear [11]. This method originating from image processing and camera calibration is now applied to magnetic sensors. The projection is done by multiplying the input coordinates $\mathbf{p}$ with the matrix $\mathbf{H}$. This matrix is calculated in the training phase. The corrected sensor signals $\mathbf{p}'$ are calculated in the operation phase by applying equation (4).

$$\mathbf{p}' = \mathbf{H}\mathbf{p} = \begin{pmatrix} h_{11} & h_{12} & h_{13} \\ h_{21} & h_{22} & h_{23} \\ h_{31} & h_{32} & h_{33} \end{pmatrix} \begin{pmatrix} x \\ y \\ 1 \end{pmatrix} \quad (4)$$

As a second model, the polynomial transformation can compensate errors, using equation (5). The polynomials must be formed for each individual

quadrant. The training phase approximates the coefficients a_n .

$$f(x, y) = a_0x + a_1y + a_2xy + \sum_{n=1}^{n_{\max}} (a_{2n+2}x^{2n} + a_{2n+3}y^{2n}) \quad (5)$$

In the operation phase, the parameters are fixed for the steps described before. In this phase all the calculation could be performed with integer numbers. In the future it can be strongly optimized in on-chip hardware for speed.

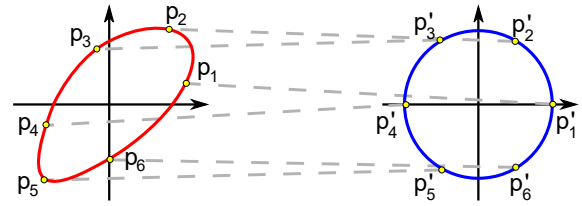


Fig. 5. Training procedure for model based angle correction. Distorted trajectory from the sensor signals of the real system are assigned to ideal coordinates.

4. Evaluation Results

The simulation setup in Fig. 2 is used for the evaluation of the methods. A typical but significant example axis misalignment of 25 % of the array size in both directions will be shown in the following examples. The evaluation follows the steps of Fig. 4.

4.1. Extraction of Position and Angle Components

The position detection of the magnet takes place on the surface of the array. Since the sensors only detect the magnetic field components in the X-Y-plane, three-dimensional position detection is not possible. Therefore the position is only a projection of the magnet onto the surface of the sensor array. The position calculation is carried out by equation (2).

For a 360 ° rotation, the percentage deviation of the ideal and calculated dipole position has a range from -8.5 % to 1 %, see Fig. 6. From these values the mean value over all rotation angles is determined for further steps. For the example shown, after averaging the position, the deviation is -3.31 % for x_0 and -3.44 % for y_0 .

The angular components extracted from the second derivatives at the point x_0, y_0 of the dipole show a cosine and sine waveform, see Fig. 7. However, these components show a distortion, which can be seen in the X-Y diagram.

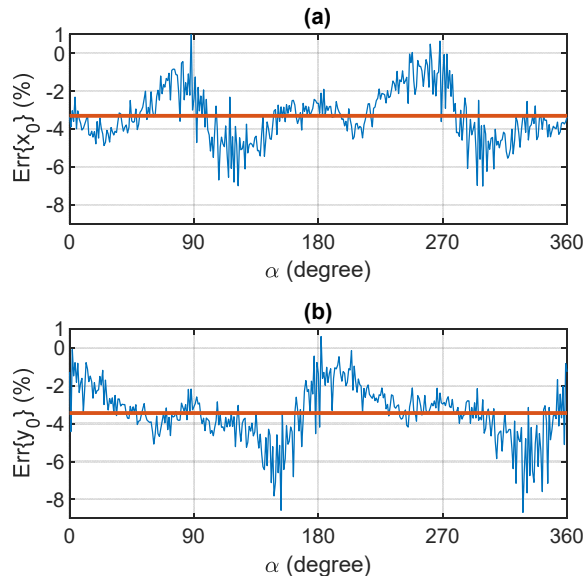


Fig. 6. Relative deviation of the model based encoder position with the ideal position for both vector components x (a) and y (b). The noise is mainly generated by the quantization of the sensor values in the array and corresponds to an ADC quantization.

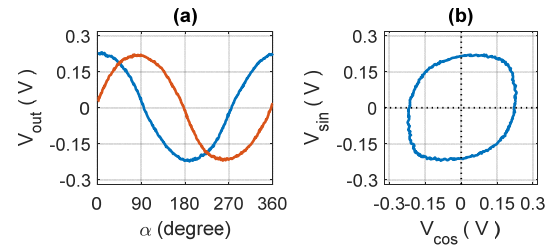


Fig. 7. Example of the angle extraction with bivariate polynomials, see equation (3). Extracted cosine and sine component over the encoder angle (a). Same as X-Y-diagram (b). These values are inputs of the error compensation, see Section 4.2 and Fig. 8.

4.2. Evaluation of Data Transformation

A residual error remains after extraction of the angular components in the first step of the operation phase, see Fig. 4. This can be significantly reduced in the second step by a transformation of the angular components. As a reference an ideal system is assumed, which has a circle as a trajectory in the X-Y-diagram, see Fig. 8.

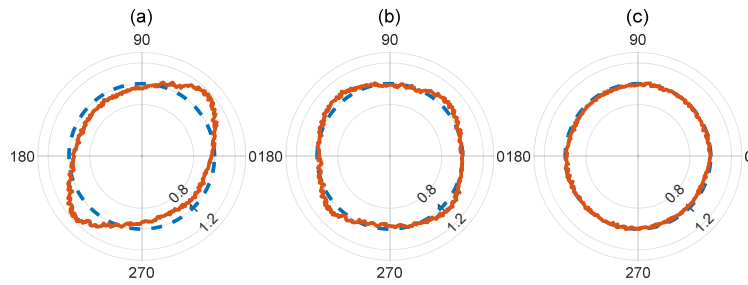


Fig. 8. Results of the compensation methods in polar coordinates. The unit circle ($R = 1$) shows the ideal angle position of the encoder as blue dashed circles. The diagram (a) shows the deviation before any compensation. The diagram (b) shows better results after projective transformation. The diagram (c) shows much better results by bivariate polynomial transformation.

In addition to the polar plot the angular error is plotted over the encoder angle. The result of the calculation can be seen in Fig. 9. The results from Fig. 8 and Fig. 9 are provided in the following Table 1 as numbers.

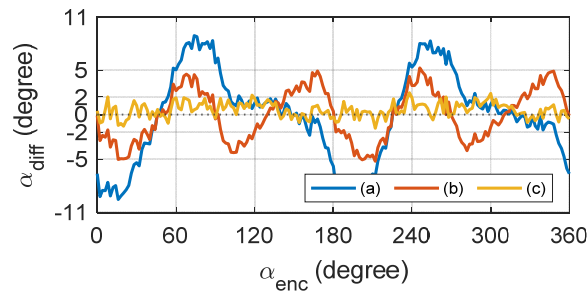


Fig. 9. Example shown in Fig. 8 as remaining angle error. (a) not compensated, (b) projective transformation and (c) bivariate polynomial transformation (c).

Table 1. Maximum and mean angular errors ($E_{\max}$ and E_{mean}) before and after compensation by projective transformation or polynomial transformation.

	No Compensation	Projective Transf.	Polynomial Transf.
$E_{\max}$ (degree)	9.5760	5.2330	2.6312
E_{mean} (degree)	3.9006	2.4980	0.7897

In general we find a significant improvement with the proposed methods. The polynomial transformation has better results but slightly more calculation effort, because of separate application for each quadrant.

5. Conclusion

We discussed the application of TMR sensor arrays for angle measurement in technical applications with

mechanical tolerances. The large number of pixels of the array is used for error compensation. We focus on two model based methods in our paper. The model parameters are extracted in an initial training phase. An effective use of the parameterized calculation model follows during the operation phase. The methods are evaluated by simulation and hardware experiments. The misalignment is limited to the boundaries of the sensor array, to prevent the ambiguity of the models. Larger misalignments could be handled with a model free statistical method, which are currently under research [12].

As a dedicated feature of magnetic sensor arrays, we propose a model based method, which reduces the remaining angular error due to the mechanical deviations by up to one order of magnitude.

Acknowledgements

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(022)

Evaluation of Electrochromic Properties of Polypyrrole Films Modified by Phenothiazine Derivatives

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Summary: The aim of this research was to investigate three different syntheses of films based on polypyrrole and three phenothiazine derivatives - methylene blue, azure A, thionine. Films were synthesized on indium tin oxide (ITO) coated glass electrode by cyclic voltammetry (CV) method at scan rate of 50 mV/s between -0,2 and +1 V for 25 cycles, step potential 2,44 mV. The obtained films were analyzed by cyclic voltamperometry, chronamperometry, AFM methods. The surface morphology of the coatings was determined by AFM and it was found that the thickness of the surface irregularities ranges from 1 to 3 μm . The investigated polymeric coatings exhibited electrochromic properties (color change in different media), reacted to pH changes.

Keywords: Electrochromism, Polypyrrole, Methylene blue, Azure A, Thionine, Electropolymerization

1. Introduction

Electrochromism is the phenomenon, where the color of a material changes by applying a voltage [1]. This feature is very important for a wide range of materials such as smart windows, screens, thermal modulators and others [1]. Electrochromic properties are characteristic for some metal oxides and also some organic materials such as polymers too [2]. Polymers have a wide range of possibilities for both synthesis and application, and therefore can be used in the production of electrochromic materials [3, 5].

Phenothiazine derivatives (Fig. 1), which are biologically active [4] and widely applicable may also be polymerized.

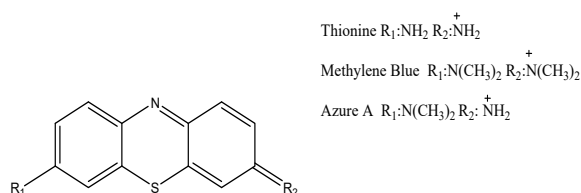


Fig. 1. Structure of phenothiazine derivatives (thionine (TH), methylene blue (MB), azure A (AA)).

The aim of this work was to synthesize polypyrrole films modified with phenothiazine derivatives (MB, AA, TH) and to investigate the properties of the obtained materials.

2. Electrochemical Deposition of Polypyrrole-phenothiazine Derivative Film on Galss/ITO Electrode

Fig. 2 represents cyclic voltammograms recorded during electropolymerization of Ppy-MB (Fig. 2A), Ppy-MB (Fig. 2B), Ppy-TH (Fig. 2C).

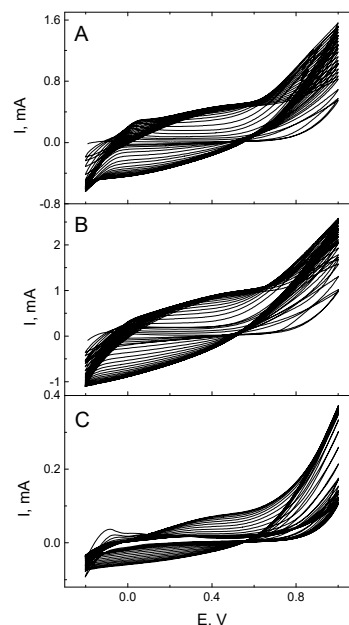


Fig. 2. Electropolymerization of Ppy-MB (A), Ppy-AA (B), and Ppy-TH (C) films. Polymerization solution: 10 mM phenothiazine derivative and 50 mM pyrrole (in water). CV at scan rate of 50 mV/s, from -0.2 V to +1 V, 25 cycles, step potential 2.44 mV.

A polymerization solution was prepared of 10 mM phenothiazine derivative and 50 mM pyrrole in water. Films were synthesized on indium tin oxide coated glass electrode by cyclic voltammetry (CV) at scan rate of 50 mV/s between -0.2 and +1.0 V for 25 cycles, step potential 2.44 mV.

3. Evaluation of Polypyrrole-phenothiazine Derivative Film Properties

The obtained films were analyzed by cyclic voltammetry (CV), chronoamperometry (CA), AFM methods.

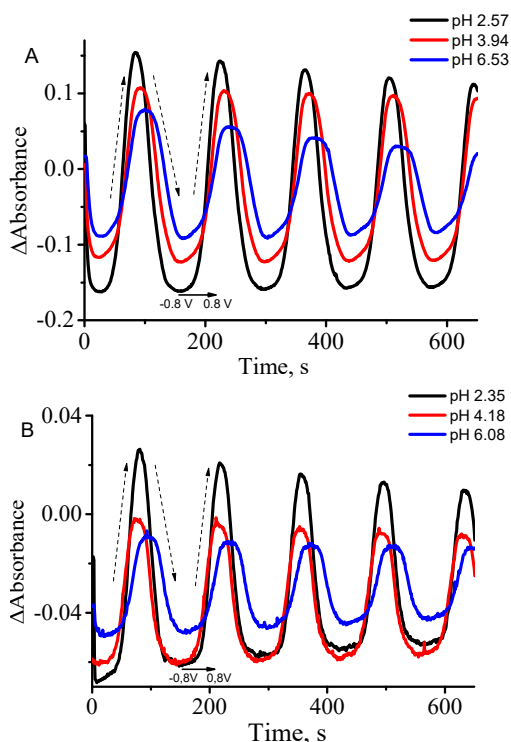


Fig. 3. Dependence of absorbance on pH value. Ppy-MB (A) and Ppy-TH (B). Absorbance changes were followed in BRB solution. pH was adjusted with NaOH solution (0.1 M). CV (-0.8 V to 0.8 V, scan rate 50 mV/s, 5 cycles).

Fig. 3 demonstrates optical absorbance (Fig. 3A at 668 nm and Fig. 3B at 750 nm), which points the variation of optical absorbance of the Ppy-MB or Ppy-TH film as a function of the pH of BRB solution and potential value. Layer of Ppy-AA was too thick and the optical absorbance was not measured. Optical absorbance was measured during potential cycling from -0.8 V until 0.8 V, at potential sweep rate of 50 mV/s by 5 potential cycles at the specified pH of solution. The highest absorbance shift from leuco form to colored form was obtained in the most acidic BRB solution. Due to the increased pH of the BRB solution shift of absorbance of leuco form of phenothiazine to

the colored form was decreasing. It can be assumed that optical absorbance value decreases with increasing pH value and electrochromic properties of Ppy-MB or Ppy-TH are well observed until pH 6.5. When pH exceeded value of pH = 6.5, then the destruction of the polymeric film was observed.

4. Conclusions

In the synthesis with pyrrole and phenothiazine derivative, the preferred method for the synthesis was cyclic voltammetry, with a potential range of -0.2 V to 1 V, 50 mV / s, 25 cycles with a step lift of 2.44 mV and 10 mM in the polymerization mixture. phenothiazine derivative, 50 mM pyrrole.

The obtained films were investigated, and their properties evaluated. Polymer films exhibited electrochromic properties, reacted to pH changes. Colour change could be seen in acidic medium and the change of the colour was reversible. It can be assumed that electrochromic properties of Ppy-PT-based coatings are well observed until pH 6.5.

Acknowledgements

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The Use of Sensors to Measure Soil Moisture, Plant Temperature and Vegetation Index for Turfgrass Vigor Evaluation

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Summary: Urban grasslands have a great capacity to reduce air pollution, as well as to produce a positive impact on the quality of life of citizens. Nonetheless, their maintenance and sustainability requires costly inputs, and it is difficult to achieve in regions with a lack of economic resources. In this context, the use of low-cost sensors for monitoring vegetation and water status has revealed a various advantage to improve the quality and vigor of the grasslands. In this study, the application of three different categories of sensors to estimate turfgrass growth under different level of irrigation was evaluated. Parameters evaluated were the vegetation index using the GreenSeeker sensor, the plant water status using the Infrared Thermometer and the soil moisture measured by the Time Domain Reflectometry Sensor. This study highlights the usefulness of low-cost tools based on sensors to assess turfgrass growth and water status. Moreover, this study also showed a higher correlation between data of sensors used, allowing then their use in comparable manner.

Keywords: GreenSeeker sensor, Infrared thermometer, Soil moisture, Turfgrass vigor, Vegetation index, Canopy temperature.

1. Introduction

Currently, the use of technologies as sensing techniques, wireless sensors networks (WSN) and remote sensing are essential in agriculture. They are englobed in the concept of Precision Agriculture (PA). The use of those technologies to reach irrigation efficiency and to save water is becoming a principal critical way to enhance crop sustainability [1].

In turfgrass management, the irrigation is centralized and well-planned to minimize the required inputs for its maintenance. Thus, the use of WSN and remote sensing has become fundamental tool to improve grassland maintenance and preservation [2]. Many turfgrass studies have demonstrated that parameters derived of spectral reflectance sensors are highly associated with visual turf quality, shoot density and has been used to PA applied to turfgrass management [3].

The main objective of this study was to evaluate the useful of low cost remote sensing techniques to assess turfgrass vigor, under different level of irrigation. The sensors data were used to estimate turfgrass canopy green biomass (by the GreenSeeker), the canopy temperature (by the Infrared Thermometer) and the soil moisture (by Soil Humidity Sensors). Furthermore, we will also examine the relationship between the different sensing techniques in order to use them in a comparable manner for turfgrass irrigation management.

2. Material and Methods

Field trials were conducted at the Madrid Institute for Rural, Agrarian and Food Research and Development (IMIDRA). Six experimental plots cultivated with a turfgrass mixture composed by *Poa pratensis* (C₃ plant) and *Buchloe dactyloides* (C₄ plant) were tested. The equipment used to collect the data are presented in Fig. 1.

The first parameters evaluated is the Normalized Difference Vegetation Index (NDVI). It was measured with a portable spectroradiometer, the GreenSeeker handheld crop sensor (Trimble, USA). The NDVI is obtained using the following equation: $(NIR-R)/(NIR+R)$, where R is the reflectance in the red band and NIR is the reflectance in the near-infrared band. The distance between the sensor and the plots was constant and around 50-60 cm above and vertical to the canopy. Moreover, the canopy temperature (CT) was measured using an infrared thermometer (Fluke 561 sensor, China). Finally, the soil moisture (SM) was measured with a Time Domain Reflectometry (TDR) using the Field Scout TDR 350 sensor (Spectrum Technologies, Inc., USA). Data were subjected to factorial analyses of variance (anovas) to test the effects of irrigation on all parameters studied. A bivariate correlation procedure was constructed to analyze the relationships between NDVI, CT and the SM.



Fig. 1. The different devices used in this study.

3. Results

Results showed that all variables derived from the different sensors used presents significant differences between the different irrigation regimes (Table 1).

Table 1. Means values of NDVI, CT and SM. Significance levels: *** $p < 0.001$.

Irrigations	NDVI	CT	SM
Low	0.64	21.00	15.60
High	0.83	13.50	43.18
ANOVA	0.001***	0.000***	0.000***

Lower values of turf NDVI were observed under low irrigation, whereas under the well irrigation NDVI was higher. According to our results, the authors of [4] have demonstrated that NDVI was highly correlated with visual turf quality, and had been used to measure drought stress in turfgrass.

Moreover, CT has been long recognized as valuable index to detect differences in plant water status [5]. In this context, the results of this study showed differences in plants CT under different level of soil moisture.

Furthermore, both groups of plant vegetation and water status indices present a high and positive correlation between them (Fig. 2). This result confirms that both sensing methods, GreenSeeker, which measures NDVI and infrared thermometer, which measures CT inform similarly about plant water stress and vigor.

Likewise, highest correlation was also found between SM and both NDVI and CT. SM measured by the TDR 350, presented a highest and positive correlation with NDVI and negative with CT. Thus, the soil moisture, can help to maintain the irrigation efficiency and then turfgrass quality and vigor.

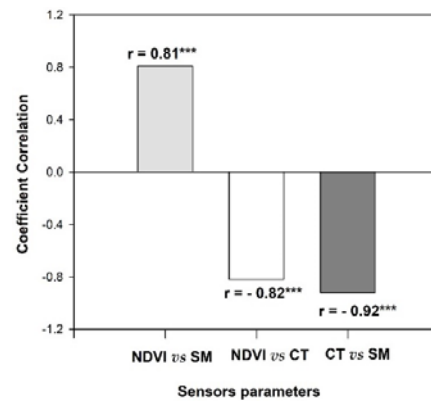


Fig. 2. Relationships of NDVI with CT and SM and the CT with SM across all turfgrass plots assayed.

4. Conclusions

Results of this study highlight the usefulness of the three sensors used for turfgrass growth assessment. The low-cost and easy-to-handle characteristics of the three sensors makes them a useful device for turfgrass management. The GreenSeeker and the infrared thermometer are a perfectly device for turfgrass canopy water status and green biomass estimation. Moreover, soil sensor can inform perfectly of the soil water availability and thus permit the control of water irrigation. Finally, the use of the easy remote sensing technique can help to detect water stress at real time and to take a rapid decision in field for the grassland sustainability.

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High Sensitivity Digital GMI Sensor Based on FPGA

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Summary: This paper presents of a high sensitivity digital off diagonal Giant Magneto Impedance (GMI) sensor using Field Programmable Gate Array (FPGA). The proposed implementation is mainly made of a Direct Digital Synthesiser (DDS) for the excitation of the GMI wire and a digital receiver composed of an Analog-to-digital converter (ADC) and a Digital Down Converter (DDC) for quadrature digital demodulation of the voltage across the GMI wire. The proposed approach allows to give the key implementation rules of the digital electronic conditioning in order to minimize the equivalent magnetic noise in the white noise region. The results show that equivalent magnetic white noise level of 1 pT/√Hz is obtained in the white noise region with respect to the proposed rules. These results are achieved using a sensing element composed of a 400 turn coil wound around a Co-rich GMI wire.

Keywords: Giant Magneto Impedance (GMI), Digital electronic conditioning, Low noise.

1. Scientific Background

Magnetic sensors, as a device for converting magnetics signals into electrical signals, play an essential role in almost all sectors of engineering and industry, such as biomedical applications, non-destructive testing, spatial, and industrial applications to measure pT bio-magnetic fields, magnetic fields of nanoparticles at the nT level or geomagnetic fields at the μT level [1]. Nowadays, there are a wide variety of sensors with different physical principles. A non-exhaustive list can for example include Hall effect sensors, Fluxgates, inductive sensors, Anisotropic MagnetoResistance (AMR), Giant MagnetoResistance (GMR) sensors, etc. Over the past two decades, the Giant Magneto-Impedance (GMI) effect has been intensively studied due to its attractive potential for realizing a highly sensitive magnetic sensor [2]. The GMI characterizes the huge variation of impedance of a ferromagnetic conductor subjected to an external magnetic field and supplied by an high frequency excitation current i_{ac} . It has been shown that GMI sensors offer several advantages compared to conventional sensors. Some of the decisive advantage is the high sensitivity, the potential for miniaturization and the low energy consumption [2].

In a GMI magnetic sensor, the sensing element which can be composed of an amorphous ferromagnetic wire for example, is supplied by a High Frequency (HF) excitation current obtained by an oscillator associated with a voltage-to-current converter. When the sensing element is submitted to an external magnetic field to be measured, the voltage across the GMI element is amplitude modulated by this field due to the modification of the ac impedance. An amplitude demodulator is then used so as to have information of the measured field. The relation between the measured field and the sensor output is characterized by the sensitivity.

In the context of high sensitivity, the equivalent magnetic noise (EMN) level has to be improved since it characterizes the smallest magnetic field measurable by the sensor [3]. EMN is defined by the ratio between the noise spectral density (NSD) and the sensor sensitivity. Usually, this noise spectral density or the equivalent magnetic noise level can be separate into a low-frequency excess (1/f) noise attributed to the fluctuations of the magnetization movement and a white noise region due to the electronic conditioning [3, 4]. In [5-7], we have introduced the digital concept in GMI sensor in order to reduce the electronic conditioning noise in the white noise region. Moreover, the sensitivity has to be improved also so as to enhance the equivalent magnetic noise. The adopted solution to improve the sensitivity in GMI sensors is to use the sensing element in an off-diagonal configuration that consists of using a pick-up coil around the GMI sensing element [3]. Name such as orthogonal Fluxgate in fundamental mode is also used in the literature [8].

This paper presents a digital electronic conditioning for the high sensitivity GMI magnetic sensor based on a Field Programmable Gate Array (FPGA) and rules to predict white noise performance. It is organized as follows. The hardware architecture of the digital GMI sensor is described in Section 2. Section 3 presents the implementation rules from noise point of view and the prediction of the noise performance in the white noise region. Preliminary results are given in Section 4. Finally, Section 5 gives a general conclusion.

2. Digital Off-diagonal GMI Sensor

Fig. 1 shows the hardware architecture of the proposed high sensitivity digital off diagonal GMI sensor. It is mainly made of the excitation system

(which includes a Direct Digital Synthesizer (DDS) and the voltage-current converters), the sensing element in off-diagonal configuration and the detection stage (including a preamplifier, an Analog-to-digital converter (ADC) and a Digital Down Converter (DDC)).

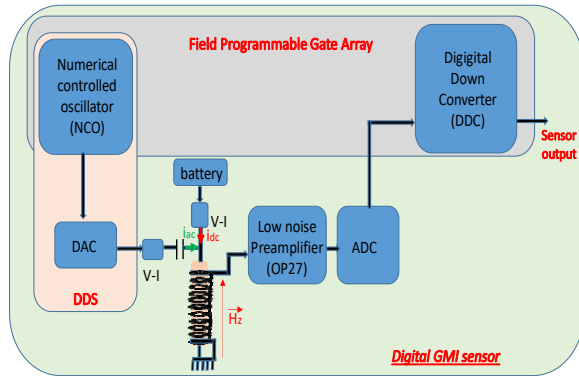


Fig. 1. Hardware architecture of the off-diagonal GMI sensor using a Direct Digital Synthesizer (DDS) and a Digital Down Converter (DDC) implemented using a Field Programmable Gate Array (FPGA).

The proposed sensor is based on a Field Programmable Gate Array [9, 10]. FPGA are configurable integrated circuits used to design digital functions. Their configuration is usually specified using hardware description languages such as VHDL. The Configurable Logic Block (CLB) is the basic element of an FPGA circuit. These are fundamental logical resources used for implementing combinatory and sequential functions. They mainly consist of Look-up Table (LUT), multiplexers and D flip-flop. The CLB can be used to create complex logic functions but also as variable storage elements. This structure makes the FPGA architecture effective since any digital design uses these resources simultaneously.

The novelty of this work, lies mainly in the fact that the DDS and DDC, required in the sensor implementation, are fully implemented on the FPGA.

2.1. The Excitation System

As presented in our previous work [5, 6], the DDS supplied the High Frequency (HF) excitation current for the off-diagonal GMI sensing element. This DDS is mainly made of a Numerical Controlled Oscillator (NCO) implemented in the FPGA and associated to a Digital-to-Analog Converter (DAC). The DDS is a technology that uses digital circuits to generate a programmable frequency output signal from a reference clock (125 MHz in our design) and a binary number (image of the desired frequency). The core of the DDS is the NCO, which integrates a phase-to-amplitude converter (usually a LUT) and a phase accumulator implemented to generate an appropriate phase argument. This phase argument is mapped to the address bus of the LUT which carries out the phase-to-amplitude conversion [9]. The DDS output voltage is

applied to a voltage-to-current converter (V-I) generally based resistor for supplying the HF excitation current, i_{ac} , for the GMI element.

2.2. The Sensing Element

When the sensing element is subjected to the magnetic field to be measured, a voltage, V_{coil} , is induced in the pick-up coil. This induced voltage results in fact from a cross magnetization process based on the appearance of an ac longitudinal magnetization in a ferromagnetic material (the GMI wire in this study) which is excited by a circumferential magnetic field created by the HF excitation current [11]. This longitudinal magnetization explains the appearance of the voltage across the pickup coil. To observe this longitudinal component of magnetization, static magnetization must follow a helical path around the direction of the excitation current. This helical static magnetization can be obtained in two ways. The first one consists in introducing a helical magnetic anisotropy by torsion or torque annealing the magnetic material. Such structures are generally used in sensors based on the reverse Wiedmann effect [12, 13]. The second way to obtain this magnetic structure in the case of a GMI wire with circumferential anisotropy, as is the case in our sensor, is to use a continuous static current, I_{dc} , superimposed to the i_{ac} excitation current. The static circumferential field produced by this direct current combined with the axial measured field, H_z (see Fig. 1), forces the magnetization to follow a helical path. That's why, in the current design, a static current, I_{dc} , is supplied to the GMI wire using a battery associated with a V-I converter generally based resistor. The use of this static current has also been shown to reduce magnetic noise [14].

2.3. The Detection Stage

The voltage across the pick-up coil is optionally amplified by a low-noise preamplifier before being digitized by a high speed analog-to-digital converter (ADC). At the output of the ADC, data are then transmitted to the Digital Down Converter implemented in FPGA for a quadrature lock-in detection that allow to realize a sensor without using a bias field and consequently an offset cancelling devices which are additional noise sources. This field biasing could also be affect by temperature [15] and stress [16]. This DDC include digital mixing, rounding circuit, and a cascade of two low pass decimation filter: a Cascaded Integrator Comb decimation filter (CIC) and a Finite Impulse response filter (FIR). Rounding circuits are necessary for real-time implementation since only a finite number of bits can be processed by the hardware. This DDC also includes Shift-up circuit that allow to set the digital gain.

3. Noise Analysis

The quantification of the voltage noise spectral density of the sensing element associated with its conditioning electronics has been studied [6] in our recent work. The different noise sources involved in a digital GMI sensor are: the excitation system noise, the noise of the sensing element and its preamplifier, the ADC noise and the DDC noise.

3.1. The Excitation System Noise

The noise of the excitation system include the DDS noise and the two voltage-current converters noise. This noise source is seen at the sensor output by the transfer function of the off-diagonal configuration [7]. Fig. 2 shows an example of curve obtained using the off-diagonal configuration.

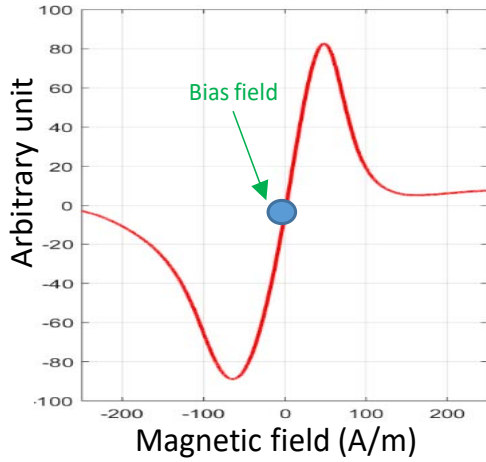


Fig. 2. Intrinsic response of the off diagonal configuration.

Since this magnetic configuration has the advantage of having an intrinsically zero response of the impedance at null field (i.e. odd function), choosing a bias field $H_{Z0} = 0$ A/m allow to roughly reduce the excitation system noise at the sensor output.

3.2. The Sensing Element Noise and the Preamplifier

In the sensing element, two noise sources have to be considered: the magnetic noise linked to fluctuations of the magnetization movement of the GMI wire and the thermal noise linked to fluctuations of the active power of the GMI wire and the pick-up coil [7]. For the preamplifier which acts as an impedance adapter between the sensitive element and the ADC, it must be chosen so that its equivalent input noise is much lower than that of the ADC [19]. In our design, the low noise OP27 preamplifier is used.

3.3. The ADC noise

At the output of the preamplifier, the voltage is transmitted to the ADC which is one of the major elements of the digital concept. The ADC is

characterized by: its input dynamic corresponding to the full scale, V_{FS} , and its resolution corresponding to its effective number of bits $ENOB$. Currently, the used converter is an 11-bit converter with a dynamic range of 2 V. The quantum q corresponding to the Least Significant Bit (LSB) designates the smallest input change that allows obtaining a bit change at the output. It is given by:

$$q = \frac{V_{FS}}{2^{ENOB}} \quad (1)$$

During the analog-to-digital conversion a difference at the origin of the quantization noise appears between the digitized voltage and the actual analog voltage. Assuming that this noise is uncorrelated to the input signal and it is uniformly distributed between the quantization levels, the RMS value of the ADC noise, V_{noise} , is given by [17]:

$$V_{noise} = \frac{q}{\sqrt{12}} \quad (2)$$

In practice, this noise is uniformly distributed in the Nyquist bandwidth $[0; F_s / 2]$. Thus, the voltage noise spectral density (in $V/\sqrt{Hz}$) of the ADC, $e_{n_{ADC}}$, can be evaluated by:

$$e_{n_{ADC}} = \frac{V_{FS}}{2^{ENOB} \sqrt{6F_s}} \quad (3)$$

3.4. The DDC Noise

Data at the ADC output is then transmitted to the DDC for digital synchronous demodulation. In the digital concept, the considered noise source is the noise due to the suppression of certain bits also called computation noise or rounding noise essential during digital processing on FPGA. Fig. 3 shows a block diagram of the different noise sources of the sensor.

So that noise from digital processing in the FPGA implementation does not degrade the overall noise performance of the sensor, the size of the computing accumulators in the DDC at the output of each digital circuit must be chosen so that the LSB of each circuit is much lower than that of the ADC [19]. As a result, the voltage noise spectral density and hence the equivalent magnetic noise level could be limited by the ADC.

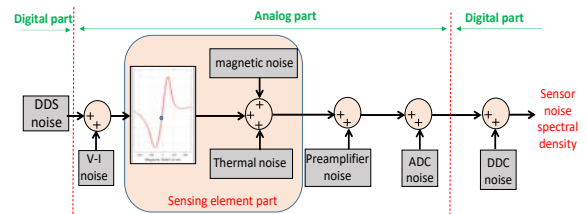


Fig. 3. Bloc diagram of different noise sources involved in a digital GMI sensor.

4. Results

To validate the proposed approach, noise measurement have been conducted. The sensitive element was a CoFeSiB wire (30 mm long and 100 μm diameter) with a 400-turn pick-up coil. The excitation current, i_{ac} , was 1 MHz of frequency and 7 mA of amplitude. A reference signal (70 nT_{rms} peak amplitude, 30 Hz frequency) was applied to the sensor to ensure its proper functioning. A DC current of 4 mA was also used. The digital gain of the DDC was set to around 40. With this setting and using an ADC sampled at 125 MHz with an effective number of bit of 11, a voltage noise spectral density of approximately $1 \mu\text{V}/\sqrt{\text{Hz}}$ is measured in the white noise region. Fig. 4 shows measurement of the noise spectral density of the sensor. With respect to the proposed implementation rules, we can observe that the white noise floor of the sensor is mainly due to the ADC which has an equivalent noise contribution of $0.98 \mu\text{V}/\sqrt{\text{Hz}}$ according to equation (3).

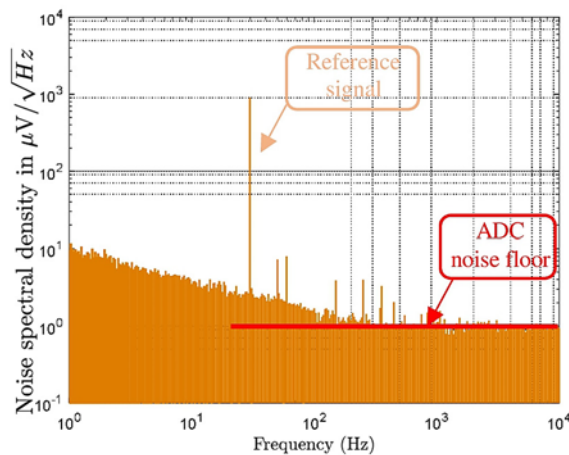


Fig. 4. Voltage Noise spectral density of the digital sensor; The excitation parameters are $I_{ac} = 7\text{mA}$, $I_{dc} = 5\text{mA}$. The digital gain of the DDC was about 40. Bandwidth at -3 dB = 41 kHz.

The equivalent magnetic noise (EMN) is then measured. Recall that the EMN is given by the ratio between the voltage noise spectral density and the sensitivity of the sensor. In this configuration, a sensitivity of about 0.9 MV/T was measured. This yields to a measured equivalent magnetic noise level of $1.1 \text{ pT}/\sqrt{\text{Hz}}$ in the white noise region. From these noise levels perspectives, the integration of the digital concept in GMI sensors opens up a wide range of applications. These performances can be improved by decreasing the noise spectral density of the ADC since it limits the white noise performance. According to equation (3), this noise spectral density can be reduced by using a converter with a higher resolution (i.e. a larger effective number of bits) or a converter sampled at a higher sampling rate. However, this brings us to a technological limitation to overcome since, generally, the increase of an ADC resolution results to a decrease of the sampling rate.

5. Conclusion

We have presented in this paper the hardware architecture of a digital GMI sensor based on FPGA and some implementation rules in order to minimize the equivalent magnetic noise level in the white noise region. The concept combines the digital concept and the sensing element off-diagonal configuration. This magnetic configuration has the potential of having an intrinsically zero response of the impedance at null field that allows to minimize the noise contribution of the excitation system at the sensor output and to design a sensor without using a bias field. The obtained results showed that, following the proposed rules, the equivalent magnetic noise level is limited by the ADC. A performance of $1.1 \text{ pT}/\sqrt{\text{Hz}}$ was obtained in the white noise region. Finally, further methods to improve the white noise performance were proposed.

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(025)

Possibility for a Feature Value Extraction Method Using Photoacoustic Imaging to Evaluate Predictive Maintenance on Blood Clotting in an Extracorporeal Circuit

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Summary: We measure blood clotting by using a pressure sensor in blood circuit, but the sensor doesn't respond quick enough because the blood clotting can only be detected by an increase of the pressure caused by the clotting itself. This research is to evaluate predictive maintenance on MATLAB, based on imaging data from a new sensing method for blood clotting (photoacoustic imaging with LED light) that we reported at SEIA2019. When we analysed the luminance of the imaging data (0-255 levels) randomly from photoacoustic imaging in a grid pattern, we saw that the luminance was increased or decreased at certain levels before and after the clotting. These significant changes suggest a possibility that predictive maintenance can be done on the state of blood clotting.

Keywords: Blood clotting, Photoacoustic imaging, LED, Circuit in an extracorporeal device, Temporal observation, MATLAB, Predictive maintenance.

1. Introduction

Blood purification therapy is a method of treatments represented by hemodialysis and cardiopulmonary bypass (CBP). As the technology of the blood purification therapy advances, an apheresis treatment for autoimmune diseases and liver failure covers more indications and there are more cases that require a high-level management of an extracorporeal circulation device [1-3]. The therapy is a method of treatment that uses extracorporeal circulation and, as shown Schematic diagram of the blood flow in the blood purification circuit is shown in Fig. 1. Fig. 1, the blood is drawn out of the body by a roller pump during the treatment and comes into contact with artificial foreign materials such as a blood circuit and dialyzer. The blood that makes contact with an extracorporeal circulation device causes an acceleration of blood clotting.

Heparin was discovered as an anticoagulant treatment in 1915 and a sustainable anticoagulation therapy was established. At present Heparin is commonly used as an anticoagulant for blood purification therapy [4-7]. However, because Heparin that is one of the anticoagulants behaves as a foreign body against blood, it's difficult to prevent the blood clotting completely on the surface of the circuit or in the dialyzer.

A pressure sensor is used to detect blood clotting in an existing extracorporeal circulation circuit. The sensor reacts only after the circuit is clogged due to the blood clotting or when the pressure rises because the blood is difficult to flow. Therefore, there's a demand that we need a more sensitive method than conventional methods to be able to observe the change

of blood clotting state in an extracorporeal circulation circuit.

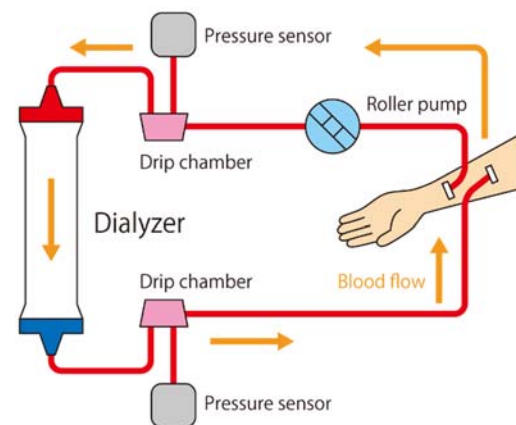


Fig. 1. Schematic diagram of the blood flow in the blood purification circuit.

Recently photoacoustic method was tried out as a measurement method to solve the aforementioned problem [8-10]. However, there's no report yet about the process of thrombus growth during the blood clotting in the blood purification circuit [11, 12].

Photoacoustic imaging is a promising structural, functional, and molecular imaging technique for a wide range of biomedical applications, with the potential for high contrast and high spatial resolution.

Recently, its application to blood flow has been reported. Elastic waves, generated by volume expansion due to the heat, are received and imaged by an ultrasound probe [13]. Our research aims to observe the change of blood clotting over time that occurs in

the extracorporeal circulation circuit during the blood purification therapy. When measuring the extracorporeal circulation circuit over time, a solid laser device that's conventionally used in photoacoustic measurement isn't suitable to use as a light source because of its weight and size in order to constantly attach it to the extracorporeal circulation device and to monitor blood clotting over time. Therefore, this research used photoacoustic imaging with LED as a light source.

The extracorporeal circuit includes an air trap to prevent air contamination. In addition, a mesh filter is installed to prevent generated blood clots flowing into the body from the external circuit. The air trap and mesh filter are installed in the drip chamber. Blood flow in the drip chamber is almost stationary due to the presence of the air trap and mesh filter as shown in Fig. 2. Since this is a basic experiment, we will use a microtube simulating the condition of the drip chamber.

Similarly to the previous research [14], we took a blood from a mouse and put it into a microtube simulating a drip chamber and we performed photoacoustic measurement on blood clotting over time by the Acoustic X using an LED light as a light source.

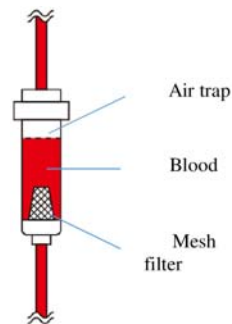


Fig. 2. The drip chamber.

In this presentation, we propose a method for converting images from photoacoustic imaging that was obtained by the photoacoustic phenomenon above to a smaller size. And we also show the result that obtained the change of the luminance level before and after clotting on mode B from the results of photoacoustic imaging analysis are able to use machine learning.

2. Method

The experimental setup is as reported in SEIA2019. The wavelength on LED that we used was 850 nm to obtain sufficient volume expansion of the light absorber. The light energy at the time was about 200 μ J/pulse. We used a 10 MHz ultrasonic probe and also ImageJ was used to convert images from photoacoustic imaging to luminance data [15-17]. Blood to be measured was put into a microtube simulating the drip chamber after it was taken out of a

mouse and we observed the change of blood clotting state through a transparent plastic.

Fig 3 (a) and (b) show the results: (a) shows immediately after mouse blood was filled; and (b) shows 10 minutes after mouse blood was filled and the state of blood clotting being confirmed. Fig. 3, on the image of mode B, when photoacoustic phenomenon responds stronger more strongly reacts, the image goes white and when it does weaker, it goes black. We divided the luminance into 256 levels – the brightest photoacoustic phenomenon level is set as 255 and no response at all is set as 0.

If the observation of the changes on blood clotting state in the extracorporeal circulation circuit was carried on continuously, the data volume would be enormous, and to avoid that, in order to reduce the data volume, we selected narrow area that has particular changes on luminance level before and after the blood clotting. Extraction of the luminance level was performed as follows. Using ImageJ to enlarge the Fig. 3 shows the grid line (green line) and sets analysing range (red window) as shown in Fig. 4. By enlarging the image the most, it shows 4×4 pixel on the grid line 1 square. The area where the photoacoustic phenomenon remarkably appeared was set as the analysis range. In this analysis range (in the red frame), 11 grid lines are arranged side by side, which is 176 pixels. Luminance per pixel sets the 256 levels from 0 to 255.

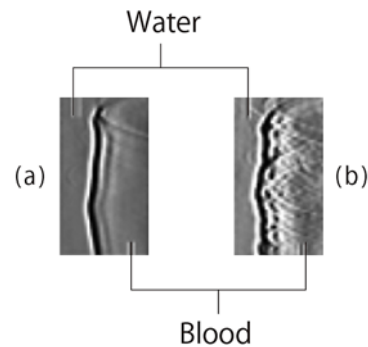


Fig. 3. Images from photoacoustic imaging (a) 0 mins after blood injection (b) 10 mins.

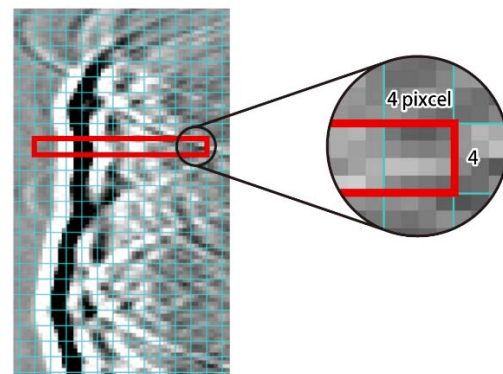


Fig. 4. Method of luminance level measurements at each pixel.

We counted luminance levels that was shown in the pixels within the measurement range (red window) and showed them as Table 1. Measurement was performed at 24 points along the microtube. In creating Table 1, MATLAB Diagnostic Feature Designer (MATHEWORKS) was used to extract a feature value from the temporal change of the obtained mode B image.

Table 1. Luminance analysis on the pixels of pictures obtained by photoacoustic imaging: (a) before blood clotting (0 mins after blood was filled), (b) after blood clotting (10 mins after blood was filled).

	Number of measurements							
	1	2	3	4	5	6	7	...
0	6	1	6	7	6	8	10	...
1	0	1	1	2	1	0	0	...
2	1	0	2	1	1	2	1	...
3	0	1	1	1	0	1	1	...
4	0	0	1	2	2	0	0	...
5	1	1	0	0	0	0	0	...
250	0	0	0	0	1	1	0	...
251	0	0	0	0	1	1	1	...
252	0	0	0	2	1	1	2	...
253	0	0	0	2	1	1	1	...
254	0	0	2	1	0	3	0	...
255	1	1	2	6	9	8	9	...
Total	176	176	176	176	176	176	176	...

(a) Luminance analysis immediately after the mouse blood was filled into microtube (Luminance levels: 0 to 5, 250 to 255).

	Number of measurements							
	1	2	3	4	5	6	7	...
0	8	8	8	9	11	11	7	...
1	1	2	1	1	1	2	3	...
2	0	1	2	2	2	2	2	...
3	0	3	1	2	0	0	0	...
4	0	2	0	1	1	1	2	...
5	2	3	2	0	2	1	0	...
250	2	3	1	3	1	2	2	...
251	0	0	0	2	1	1	0	...
252	4	1	1	0	4	5	4	...
253	2	2	3	3	4	6	5	...
254	1	2	0	2	4	3	5	...
255	18	15	17	20	23	14	14	...
Total	176	176	176	176	176	176	176	...

(b) Luminance analysis immediately after the mouse blood was filled into microtube (Luminance levels: 0 to 5, 250 to 255).

Based on the Table 1 measured as described on Fig. 4, Fig. 5 shows a comparison of the counts at each luminance Table 1 (a) before blood clotting and (b) after blood clotting.

The creation process of Fig. 5 is shown below:

① As shown in Fig. 5, one measurement range (red frame) is divided into 176 pixels.

② Specify the brightness of each pixel and count it. Do this at 24 points in the microtube.

③ Graphing the counts at each measurement point is shown in Fig. 5.

④ The vertical axis is the probability of each count total in the 0 mins after blood injection graph and 10 mins after blood injection graph(contribution ratio). The contribution ratio is by dividing each graph in Fig. 5 by total number. The contribution ratio shows how much each primary element accounts the rate of variation of the total. Graphs in orange and in blue show (a) before blood clotting (0 mins) and (b) after blood clotting (10 mins), respectively. It's shown that the greater the difference between graphs in colored by orange and blue, which represented before and after the blood clotting.

As an example, Fig. 5 shows a graph of Luminance level at 255 and 254.

3. Results

In order to identify notable luminance levels for using future values for machine learning, Fig. 5 was subjected to T-test in MATLAB and ranked. The t-test is a test of the difference between the means of samples taken from two independent populations. Based on the ranking, Fig. 6 shows results of analysis by 5-point moving average to compare changes at each luminance level (0~255). The vertical axis and horizontal axis are the T-test and luminance, respectively. It can confirm that the larger the T-test values are, the more remarkable the changes are.

Also, Fig. 7 shows the result of T-test for feature extraction. Results of the T-test at each luminance are colour coded. Below 4 to 5 is in blue; below 5 to 6 is in dark green; below 6 to 7 is in light green; below 7 to 8 is in yellow and 8 or more is in orange.

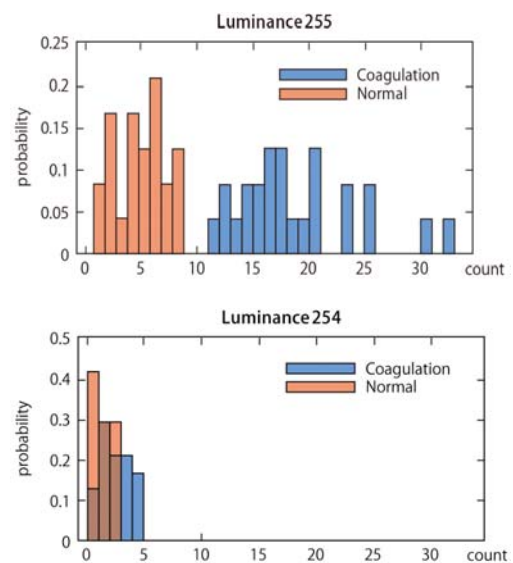


Fig. 5. Photoacoustic imaging at each luminance (comparison before blood clotting to after blood clotting in Fig. 3).

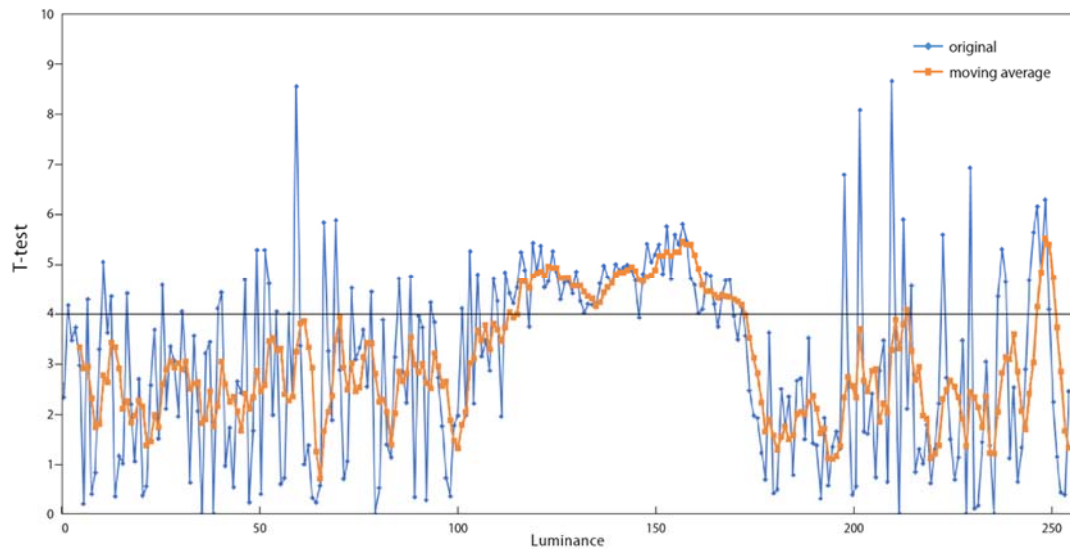


Fig. 6. 5-point moving average (based on Fig. 4).

0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
16	17	18	19	20	21	22	23	24	25	26	27	28	29	30	31
32	33	34	35	36	37	38	39	40	41	42	43	44	45	46	47
48	49	50	51	52	53	54	55	56	57	58	59	60	61	62	63
64	65	66	67	68	69	70	71	72	73	74	75	76	77	78	79
80	81	82	83	84	85	86	87	88	89	90	91	92	93	94	95
96	97	98	99	100	101	102	103	104	105	106	107	108	109	110	111
112	113	114	115	116	117	118	119	120	121	122	123	124	125	126	127
128	129	130	131	132	133	134	135	136	137	138	139	140	141	142	143
144	145	146	147	148	149	150	151	152	153	154	155	156	157	158	159
160	161	162	163	164	165	166	167	168	169	170	171	172	173	174	175
176	177	178	179	180	181	182	183	184	185	186	187	188	189	190	191
192	193	194	195	196	197	198	199	200	201	202	203	204	205	206	207
208	209	210	211	212	213	214	215	216	217	218	219	220	221	222	223
224	225	226	227	228	229	230	231	232	233	234	235	236	237	238	239
240	241	242	243	244	245	246	247	248	249	250	251	252	253	254	255
T-test															
8	7	6	5	4	3	2	1								

Fig. 7. Summary of T-test of before and after blood clotting on each luminance (0 to 255)

If the luminance of 6 or more on the T-test is considered as a feature value of blood clotting, the data would be too spread and it's difficult to comprehend but if the luminance of 4 or more on the T-test, there would be more data available to analyse. Therefore, 4 or more on the T-test and the range where luminance is gathered to some extent was used as a targeted data for analysis so that the volume of analysis data could be reduced.

As a result, on this research, photoacoustic phenomenon at luminance level of 109 to 169 before and after blood clotting is likely to produce particular changes.

4. Conclusion

In the process of blood clotting within a microtube simulating an extracorporeal circulation circuit in a blood purification therapy, we extracted the feature values from volume of changes in photoacoustic phenomenon and identified the large range of luminance for the contribution ratio. This result made it possible that when a temporal change of blood clotting in an extracorporeal circulation circuit is measured, the volume of the data handled is reduced.

We need to consider a measurement using an actual extracorporeal circulation circuit and optimization for an ultrasound probe frequency in the future.

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A Miniature Multi-sensor System and it's Application in a Condition Monitoring

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Summary: This paper presents a 10 degree-of-freedom sensor system with embedded computing power and wireless interface. The motivation is to provide a high-performance sensor system for condition monitoring, which requires minimal effort for retrofitting in existing plant and machinery. The embedded computing power enables on board calibration of the sensor module. A computational method is presented for the independent identification of the different coordinate systems, which are implicit in all real application, i.e. sensor axis to sensor housing, sensor housing to machine in manufacturing environment and the machine to the local earth coordinate system during operation in the field. The embedded computational power is used to perform geometric elliptical approximations for vibrational orbits. This eliminates the possible Gibbs error associated with elliptical Fourier descriptors. Additionally, a new statistical approach is presented for identifying periodic signals in a highly perturbed environment. Results are presented from the monitoring of large vibrating sieves; these results verify the correct functionality of the complete system and demonstrate the advantage of the new methods.

Keywords: Multi sensor system, Uncertainty analysis, Probability distribution, Condition monitoring, Retrofitting.

1. Introduction

This paper presents a 10 degree-of-freedom (DOF) sensor system with embedded computing power and wireless interface. The goal is to develop a fully integrated measurement module, which simplifies the implementation of condition monitoring in plant and machinery. To this end, the module combines a 10 DOF MEMS [1] sensor with an ARMv6 [2] architecture processor. This combination enables the sensor to deliver calibrated measurement results and to evaluate application specific algorithms, without the necessity for an additional computing device. Calibrated sensors offer the advantage of interchangeability, a very important issue in industrial applications and currently uncommon with accelerometers.

Furthermore, an IEC-61131 [3] conformant power supply specification was implemented to facilitate simple integration in industrial equipment. The wireless interface supports a low effort retrofitting to existing machinery.

The primary motivation for this development is condition monitoring of orbital vibration sieves used in material processing. However, the approaches and methods presented are applicable to a much wider range of monitoring applications. To keep the system as general as possible a MEMS device was selected with 3-axis-accelerometer, -gyroscopes, -magnetic field sensors and a temperature sensor. Whereby, the focus on this paper is on the accelerometers, calibrated measurements and evaluation algorithms.

The efficiency of orbital sieves is dependent on the path and magnitude of oscillation and their orientation wrt. gravity. Classical accelerometers [4] measure with respect to the mounting on the machine.

Here, we present an important extension to vibration monitoring, which enables the simultaneously measurement of the oscillation wrt. multiple coordinate frames, e.g. wrt. gravity and/or the machine. The challenge here is to develop models and computation methods which enable the simultaneous identification of the respective coordinate frames from the different components of the acceleration data.

The computational power within the device has made it possible to embed geometric elliptical approximations as an alternative to elliptical Fourier descriptors [5], the new approach eliminates the difficulties associated with Gibbs error [6]; the new statistical method is presented which uses the deconvolution of the PDFs for a sine wave and a Gaussian to classify the vibrations.

Results, obtained from monitoring a large industrial orbital sieve Fig. 4, are presented which demonstrate the correct functionality of the device and verify the proposed computational models.

2. Measurement Module

2.1. Module Specifications

The measurement module, shown in Fig. 1, integrates the MPU-9250 chip [7] together with an ARMv6 processor. The accelerometer provides a full-scale range of ± 16 g. The zero-g error is ± 60 mg for the x-axis and y-axes and ± 80 mg for the z-axis. The zero-g change w.r.t temperature is specified as ± 1.5 mg/°C in the temperature range of -40 °C to $+85$ °C. For further details about the specifications we refer to [7]. The embedded computer is based on an ARMV6 architecture with 1 GHz clock speed and 512 MB of memory. Power is supplied via a standard

24V IEC61131 conformant interface. It also provides a serial communication interface for initial flashing of the boot software. The sensor system provides a wireless connection via Bluetooth or Wi-Fi, this reduces cabling effort during installation and disturbance from electromagnetic fields in an industrial environment.



Fig. 1. The 10 DOF multi-sensor system. The aluminum base plate is approximately 12 cm 4.5 cm width. It is used to flange the module to the machine.

3. Reference Coordinate Frames and Calibration

The goal here is to determine the respective vibration magnitudes and shape of the vibration orbit, in the application presented here it is the vibration of the sieve. To obtain these results, in a calibrated manner, it is necessary to have calibration coefficients for the sensors and to register the different coordinate frames present in the system as a whole (see Fig. 2).

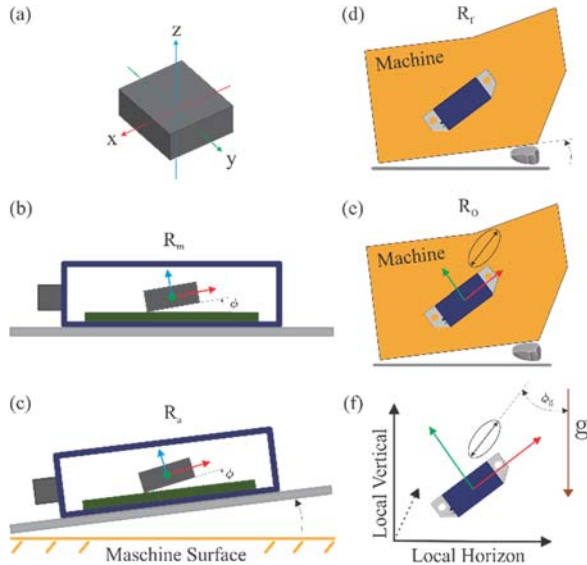


Fig. 2. The coordinate frames in the system: (a) The coordinates of the sensor module (see Fig. 3 for a more detailed drawing); (b) R_m orientation of the sensor chip within the module housing, this is due to tolerances during manufacturing of the sensor module, it must be determined during calibration of the sensor; (c) A rotation R_a due to tolerances in the orientation of the mounting flange on the machine, this should be determined during assembly of the machine; (d) A rotation R_r emanating from the machine not being perfectly level on site; (e) Orientation R_o of axes of oscillation wrt the sensor during operation. The sensor must be calibrated and the above rotations registered, so that (f) the orientation ϕ_g of axes of oscillation wrt. the earth's gravitational field can be determined.

3.1. Calibration Model

The goal of the calibration model is to correct for: the possible rotation of the MEMS device within the module R_m ; the gains g_i and offsets c_i of the individual accelerometers axis (Fig. 3). Only the calibration of the accelerometers is considered in this paper. The required calibration coefficients are determined during a calibration procedure and stored in the device. A calibrated measurement result m_c is computed from the uncalibrated data m in the following manner:

$$m_c = GR_m m + c, \quad (1)$$

whereby R_m compensates for the rotation; G and c are defined as:

$$G = \begin{bmatrix} g_x & 0 & 0 \\ 0 & g_y & 0 \\ 0 & 0 & g_z \end{bmatrix} \quad c = \begin{bmatrix} c_x \\ c_y \\ c_z \end{bmatrix} \quad (2)$$

3.2. Compensation for R_a

The tolerance of the mounting flange on the machine can lead to a rotation R_a ; however, since the MEMS accelerometers can measure down to DC it is possible to determine R_a via a static test during assembly of the machine. In the static case, the calibrated result m_c should map exactly to the orientation of the earth gravitational field (measurements are in m/s^2), i.e.

$$m_a = \begin{bmatrix} 9.81 \\ 0 \\ 0 \end{bmatrix} = R_a m_c \quad (3)$$

3.3. Non-level Machine in the Field R_f

Similarly, the compensation of the orientation of the machine, when placed in an application, can be computed as

$$m_f = R_f m_a \quad (4)$$

3.4. Orientation of the Vibration Orbits

The primary axes of the vibration orbits are computed later from the oscillatory data.

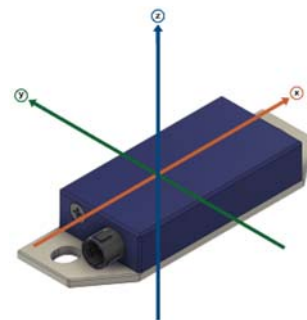


Fig. 3. Sensor module and the sensor axes orientation wrt. the module housing.

4. Reference Application

The module was mounted on the orbital vibration screen sieve shown in Fig. 4. At this test the module was positioned on the inner side of the sieve. Linear, elliptical and circular vibrational orbit modes were recorded. Each measurement was observed with a sampling frequency of 2048 Hz for 1 s in units of m/s².



Fig. 4. Left: orbital vibration screen sieve placed in an experiment area which was monitored with the sensor module in Fig. 1. Right: Position of the sensor module on the inner side of the sieve (yellow circle) and orientation of sensor axis according to Fig. 3. In this test one axis of the sensor module was aligned close to the local horizon. Photo courtesy of IFE Aufbereitungstechnik GmbH, Waidhofen an der Ibbs, Austria.

5. Data Processing

All computations presented here start from calibrated or referenced data; for simplicity we shall denote the measurement vector here as $\mathbf{m}$. In the case of n measurements they are concatenated to form the matrix $\mathbf{M}$, whereby the x , y and z values for the columns of $\mathbf{M}$. Measurements for three different oscillation modes of the sieve can be seen in Fig. 5.

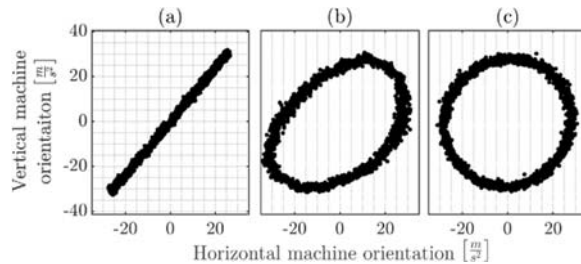


Fig. 5. Three different acceleration data sets acquired by the sensor module. A principle component analysis [8] (PCA) was performed for dimensionality reduction from 3D-axis acceleration to the main forced axes in 2D. The circular vibration screen sieve was driven in the optimal condition: (a) linear shape; in error conditions: (b) elliptical shape and (c) circular shape.

5.1. Check for Active Sensor Axes and Oscillation

The dominant direction of oscillation should be restricted to the x and y axes; the z direction should be predominantly machine noise. This is due to the manner in which the sensors are mounted on the sieve.

Consequently, an initial determination which axes are active can be determined from the ratio r_R of the standard deviations σ_x , σ_y and σ_z of the accelerations, in the following manner

$$\max \left[\frac{\sigma_x}{\sigma_z}, \frac{\sigma_y}{\sigma_z} \right] > r_R \quad (5)$$

The machine is considered to be dormant if this condition is not fulfilled.

5.2. Periodicity Trimming

The goal here is to obtain a precision separation of the DC components from the alternating component. The DC component can be used to determine the orientation of the sensors with respect to the earth's gravitational field and the alternating component to determine the frequency f_0 and magnitudes a_0 of oscillations. However, we must deal with the Gibbs error, since the oscillation may not be perfectly periodic in the measurement period. For clear oscillations we can use the zero crossings to identify the periodicity and trim the data to obtain an integer number of oscillation cycles. The DC components of the trimmed data give a better estimate than for the untrimmed data. Furthermore, the trimmed data permits an exact determination of the frequency via the fast Fourier transformation (FFT), which is numerically highly efficient.

5.3. Rotation $\mathbf{R}_a$: Module Housing – Machine (Assembly Registration)

To compensate mounting tolerances of the sensor module on the machine surface, the rotation $\mathbf{R}_a$ has to be determined. A requirement for reliable result is, that the machine is perfectly levelled wrt. gravity.

A static acceleration measurement is performed to determine $\mathbf{R}_a$. In a static measurement only the DC-components $\mathbf{M}_{DC}$ are measured by the accelerometers. Computing a PCA of the DC components $\mathbf{m}_{DC} = [x_{DC} \ y_{DC} \ z_{DC}]$ from the static measurement $\mathbf{M}_{DC}$, delivers the rotation $\mathbf{M}_a$ as follows:

$$\mathbf{m}_{DC} = \mathbf{U}_{DC} \mathbf{S}_{DC} \mathbf{V}_{DC}^T, \quad (6)$$

where $\mathbf{V}_{DC}$ corresponds to $\mathbf{R}_a$, the assembly registration. Now, the measurement rotation due to the mounting tolerances can be formulated as

$$\mathbf{M}_a = \mathbf{M} \mathbf{R}_a \quad (7)$$

The data are now exactly aligned towards the gravitational field.

5.4. Rotation R_f : Machine – Field

Rough conditions in the field can make it difficult to level the machine wrt. to the gravitational field and cause a misalignment of the machine, which is defined as a rotation R_f . The rotation R_f can be similar determined as R_a by computing the PCA of measurements M_f taken in the field of the calibrated data:

$$M_f = U_f S_f V_f^T, \quad (8)$$

where V_f corresponds to R_f , the field alignment registration.

5.5. The Inclination of the Machine in the Field During Operation wrt. Gravitational Field

To detect unwanted inclinations of the sieve during operation the rotation of the DC-components of the oscillating signal during operation can be monitored and compared to R_f . A change in the components can indicate an unintended movement of the machine or a loosening of the measurement device.

6. Data Analysis

6.1. Determine Orbits: Dimensionality Reduction

As stated, the axes of oscillation of the sieve are restricted to either a linear or closed 2D orbit. Dimensionality reduction can be applied to the 3-axes data to determine if it is a 1D- or 2D-oscillation. A principle component analysis is applied to the calibrated measurement data:

$$M = USV^T, \quad (9)$$

whereby the singular values s_i are contained in the diagonal of S . The singular values are the two norm's of the projections of the data onto the orthogonal axes, $s_i = \sigma_i \sqrt{n}$ that is, we have a measure for the activity in each principle axes. Consequently, we can determine if it is a linear vibration or an orbital vibration. The principal components are obtained by:

$$P = US, \quad (10)$$

whereby the principal components p_i are the single columns of P .

An example of calibrated measurement data for a linear oscillation is shown in Fig. 6, two of the acceleration axes are activated, this is due to the relationship between the principle angle of oscillation and the mounting angle of the sensor. In Fig. 7 the PCA for the data presented in Fig. 6 is shown.

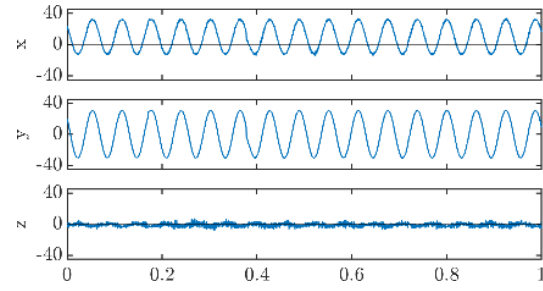


Fig. 6. This figure shows raw measurement obtained by the sensor module. An oscillation of the amplitude can be clearly seen in both axes x and y.

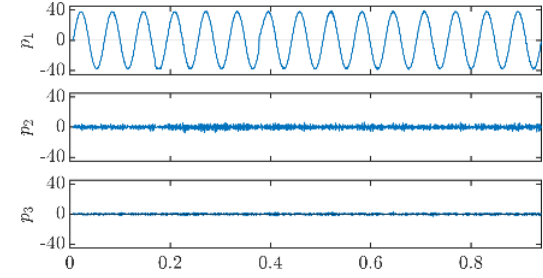


Fig. 7. The result of performing a PCA and periodicity trimming on the same data set as used in Fig. 6 reveals a linear vibration from two active sensors axes. Fig. 12(a) visualizes p_1 over p_2 of the same data set.

This reveals a linear oscillation since only one axis is active in the PCA. The normalized singular values for the presented data are: $s_1 = 1$, $s_2 = 0.0781$ and for $s_3 = 0.031$

6.2. Determine Orientation R_0 : Orientation of Orbits wrt. Gravity

The efficiency of the sieve depends on the orientation ϕ_g of the oscillation axes wrt. the gravitational field of the earth.

In order to obtain ϕ_g the mean-free and trimmed data $x_{mf} = x - \bar{x}$ and $y_{mf} = y - \bar{y}$ of the measurements M have to be considered in the following calculation. They are concatenated to form the matrix $M_o = [x_{mf} \ y_{mf}]$. A PCA is performed on M_o . The resulting components are given as follows:

$$M_o = U_o S_o V_o^T \quad (11)$$

The orientation angle ϕ_g can now be extracted from the 2D rotation matrix V_o .

$$V_o = \begin{bmatrix} \cos\phi_g & -\sin\phi_g \\ \sin\phi_g & \cos\phi_g \end{bmatrix} \quad (12)$$

In order to obtain the direction ϕ_g from M_o , the four-quadrant inverse tangent is applied.

7. Orbit Analysis

To monitor the machine efficiency the orbit must be examined for their movement behavior. In this section we introduce a new statistical approach to determine vibrational orbits and compare the results with the method of elliptical Fourier descriptors (EFD).

7.1. Spectral Analysis of Vibrational Data

A spectral analysis is performed to yield the maximum frequency $f_{\max}$ of oscillation and the respective magnitude $a_{\max}$ of both considered principal components $\mathbf{p}_1$ and $\mathbf{p}_2$. The spectrum $H_{p_1}(f)$ of $\mathbf{p}_1$ is yield by the Fourier transform $\mathcal{F}$ of $\mathbf{p}_1$:

$$H_{p_1}(f) = \mathcal{F}(\mathbf{p}_1), \quad (13)$$

whereby the maximum magnitude $a_{\max, p_1}$ and the corresponding frequency $f_{\max, p_1}$ is given by

$$a_{\max, p_1} = \max(H_{p_1}(f)) \quad (14)$$

The maximum magnitude and corresponding frequency of $\mathbf{p}_2$ can be yield in the same manner. In addition the corresponding velocity $v_{\max, p_1}$ and displacement $s_{\max, p_1}$ can be computed by

$$v_{\max, p_1} = \frac{a_{\max, p_1}}{2\pi f_{\max, p_1}}, \quad (15)$$

$$s_{\max, p_1} = \frac{v_{\max, p_1}}{\pi f_{\max, p_1}} \quad (16)$$

The so obtained orbit properties of $\mathbf{p}_1$ and $\mathbf{p}_2$ can be used for approximating a geometric model to the acceleration orbit the machine runs through during operation. Fig. 12(b) presents an elliptical measurement and its approximated orbit with elliptical Fourier descriptors.

7.2. Statistical Analysis of Vibrational Data

An alternative to spectral analysis was investigated to yield a geometric approximation for the orbit. Since the acquired data of an orbit is periodic and perturbed with Gaussian noise the probability function of the principal component $\mathbf{p}_1$ and $\mathbf{p}_2$ are expected to be a composition of the probability density functions (PDF) of a sine and Gauss [9]. Fig. 8 visualizes both PDFs and their composition. The PDF of a sinusoidal is given by

$$f(x) = \frac{1}{\pi \sqrt{1 - \left(\frac{x}{R}\right)^2}}, \quad (17)$$

where x is the equally spaced distribution of the amplitude R of the sine.

The well-known PDF of a Gaussian distribution is defined as

$$g(x) = \frac{1}{\sqrt{2\pi}\sigma^2} \exp\left(-\frac{(x-\mu)^2}{\sigma^2}\right), \quad (18)$$

where σ is the standard deviation and μ the mean.

The expected model of the vibrational data is a mixture of both distributions. It can be formulated by the convolution of both PDFs $f(x)$ and $g(x)$, analytically by

$$P_{f(x)*g(x)} = (f * g)(x) := \int_a^b f(\tau)g(\tau - x)d\tau \quad (19)$$

In this application the convolution is done numerically based on (17) and (18).

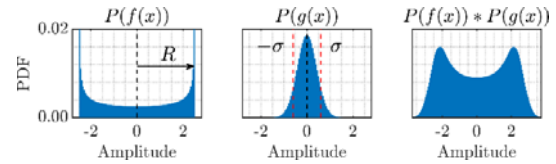


Fig. 8. Symbolic PDFs: left of a sine; middle of a Gauss; right: their convolution.

7.3. Statistical Analysis with Deconvolution of the Noise and Sinusoidal Components

A non-linear continuous approximation for the PDF $P_{f(x)*g(x)}(x)$ is performed to yield a adequate approximation of the model for the discrete PDF $P_{PC}(\hat{x})$ of the observations $\hat{x}$ of each principal component respectively. The cost between both PDFs is therefore minimized wrt. to the parameters R , σ and μ .

$$\min \varepsilon(\sigma, R) = \min \|P_{f(x)*g(x)}(x) - P_{PC}(\hat{x})\|_1 \quad (20)$$

7.4. Parameter Optimization

To find the parameters of the non-linear problem, a simplex algorithm for non-linear functions was used. The following iterative steps are performed for the optimization for each principal component respectively:

- 1) Compute the deconvoluted sine- and Gauss-PDF, $P(f(x))$ and $P(g(x))$ with the corresponding parameters;
- 2) Convolute both PDFs to yield the mixed distribution $P_{f(x)*g(x)}$;
- 3) Compute the 1-norm of the residual between $P_{f(x)*g(x)}$ and P_p .

The iteration is performed until certain termination criteria are fulfilled. Those criteria are e.g. when a certain number of iterations is reached or both, the function value and the parameter value change is smaller than a defined value.

8. Results

8.1. Module Calibration – Results

For the computation of the calibration parameters, 40000 samples in each sensor axis direction were recorded. The initial values for the non-linear parameter were set to $\phi_i = 0$. The linear parameter for the main axis were set to $g_i = 9.81$ and for the cross-axis $g_{i,j} = 0$. The initial offsets were set to $c_i = 0$. Table 1 shows the result of the module calibration. The gain g_i is shown close to be 1 in each axis. No correlated cross gains $g_{i,j}$ appeared. The offsets lie in the tolerances of $\pm 60 \text{ mg}$ for the x- and y-axis and $\pm 80 \text{ mg}$ for the z-axis, as specified in the sensor data-sheet. The assembly tolerances of the sensor chip orientation were found to be between approximately -1.2° in y and 3.3° in x.

Table 1. Results of the calibration parameters. All cross gains were calculated to be 0.

Axis	Gain g_i	Offset c_i	Angle ϕ_i
	[-]	[mg]	[°]
x	0.9971	26.8196	3.2975
y	0.9993	27.4924	-1.1998
z	0.9835	21.1213	0.4977

8.2. Statistical and Fourier – Results

The results of the statistical approach for a linear (see Fig. 9), elliptical (Fig. 10) and a circular (Fig. 11) are presented in histograms in the corresponding figures. However, each histogram shows the difference between the actual principle components $P(\hat{x})$ (= red bars) and the calculated model $P(x)$ (= green bars), where $P(x)$ is the result from the non-linear approximation of the PDFs. The optimized PDFs reveal a good estimate for the amplitude R of each principal component and the sensor noise σ . The different shapes of the distributions of the histograms are clearly visible; The circular PDF in Fig. 11 shows a quite steep gradient at the sides for both principal components. The PDF of the linear data set in Fig. 9 directly reveals the characteristically histogram for a motion in p_1 . The histogram of p_2 in Fig. 9 reveals the residual noise of the second principle component. It can be seen that the result of the model matches the

data from $p_2(\hat{x})$, which follows a Gaussian distribution.

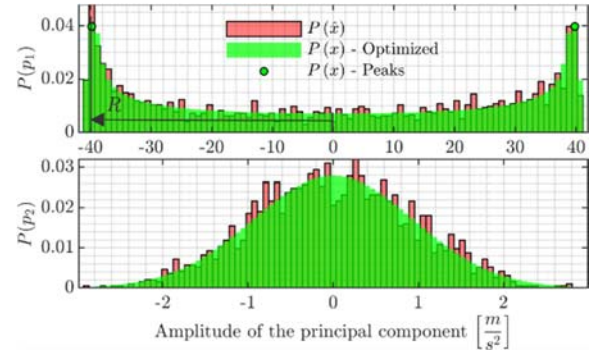


Fig. 9. Linear orbit: PDF (red) and optimized PDF (green) of the 2D principal components corresponding to the linear orbit dataset in Fig. 5(a). The optimization took place with 1-Norm residual. Note that the amplitude scale of p_2 is different to p_1 and there is only one Peak in the PDF of p_2 at an amplitude close to 0. This is what we expect since there is no oscillation in one dimension.

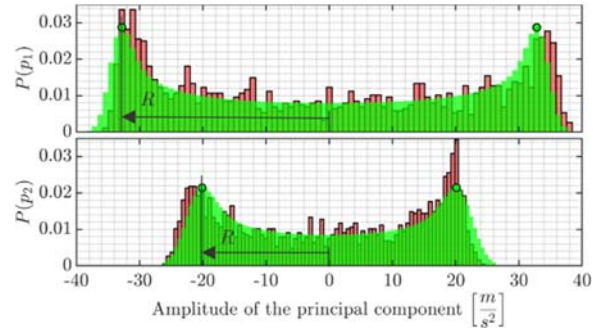


Fig. 10. Elliptical orbit: PDF (red) and optimized PDF (green) of the 2D principal components corresponding to the elliptical orbit dataset in Fig. 5(b). The optimization took place with 1-Norm residual.

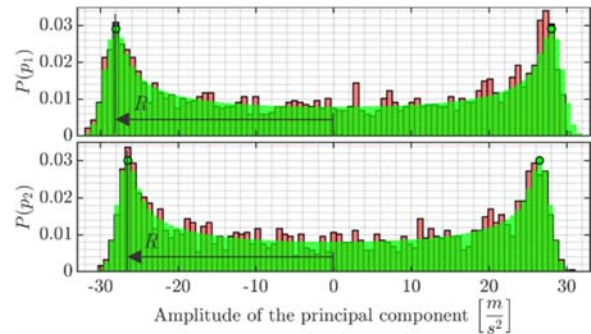


Fig. 11. Circular orbit: PDF (red) and optimized PDF (green) of the 2D principal components corresponding to the circular orbit dataset in Fig. 5c. The optimization took place with 1-Norm residual.

Table 2 summarizes the results yield by the non-linear optimization for the amplitude R and the standard deviation σ for each principal component. It also shows the results of the orbit approximation with elliptical Fourier descriptors with $n = 1$ harmonics.

Table 2. Results of the statistical characterization and of the elliptical Fourier descriptors.

Orbit	p_i	PDF Parameters		EFD Parameters	
		R	σ	R	σ
		[m/s ²]		[m/s ²]	
Linear	1	39.982	1.254	38.264	1.713
	2	0.132	1.322	1.2022	0.894
Elliptical	1	33.889	2.218	34.13	1.525
	2	21.434	3.001	21.979	1.573
Circular	1	28.977	1.633	28.775	1.284
	2	27.421	1.5	27.459	1.381

The amplitudes R between the PDF and EFD results differ for the elliptical and circular measurement by approximately 5 %. The difference between the amplitudes of the linear measurement differ significant for the principal component p_2 . The uncertainty in the amplitudes are clearly better for the EFD results.

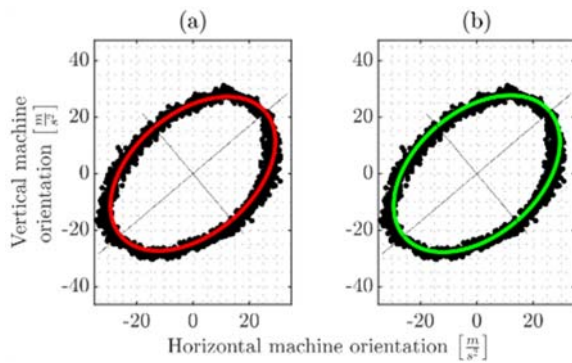


Fig. 12. (a) The orbital model yield by the ellipse computation with the optimized parameters. (b) The orbital model yield by the synthesized EFD results with $n = 1$ harmonics.

9. Conclusion

This paper introduces the combination of a MEMS-sensor together with computational power and

implemented algorithms in a compact sensor module. That allows the sensor module to provide calibrated acceleration measurements and the analysis of measurements. The challenge of retrofitting machines is facilitated since the necessity of an external computation device is no longer required. The advantage of the MEMS-sensor to measure acceleration down to DC enables the identification of the orientations of the vibrational orbit axes wrt. to the gravitational field; this parameter is decisive for the effectiveness of vibration sieves. In addition machine alignment registration in the field can be achieved. Through comparison, it can be concluded, that the presented statistical deconvolution based method is an alternative to approximate elliptical orbits, which avoids the difficulties associated with Gibbs errors.

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(028)

Reference Energy Measuring System for On-board Calibration of EMS Installed in Locomotives

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Summary: The voltages and currents supplied to locomotives are often subject to distortions with significant harmonic content (odd harmonics, sub-harmonics and inter-harmonics) that comes from train acceleration and braking. The energy measurement is a complex matter due to harmonics content. No reference systems and traceability had been developed before. This research has been developed in the framework of the European project 16ENG04 MyRails Metrology [1] for smart energy management in electric railway systems.

Keywords: Energy measurement, Energy measuring system, Calibration set-up, On-board calibration, Railway system, Harmonic, Distorted regime.

1. Introduction

A Reference Energy Measuring System has been designed for laboratory and on-board calibration of on-board Energy Measuring Systems (EMS) installed in locomotives used for AC railways 25 kV/50 Hz, 15 kV/16.7 Hz, with a maximum current of 500 A. This element is a part of the metrological framework for calibration to enable high accuracy energy and power quality (PQ) measurements under highly dynamic electrical conditions approaching the uncertainty limits stated in the EN 50463-2 [2], with a frequency range from a few hertz up to 5 kHz for AC systems. The uncertainty targets for laboratory or on-board calibration calibrations are 0.5 % for AC. Complementary requirements has been stated in the new procedure developed in the EU project [3].

2. Reference Energy Measuring System

The Reference Energy Measuring System is composed by a fluxgate current sensor with an appropriate shunt, a shielded resistive-capacitive voltage divider to avoid interference from external electrical and magnetic fields and two digital sampling multimeters controlled by a PC. Fig. 1 shows the schematic connection between sensors, multimeters and PC.

The current is measured by a sensor based on the fluxgate technology with a ratio error less than 0.1 % for frequencies up to 5 kHz. High voltage is measured by a mix resistive-capacitive divider designed by LCOE with flat-frequency response for frequencies up to 5 kHz. The characterization of this sensor reveals an accuracy class 0.2 for a range of frequencies from 20 Hz to 5 kHz.

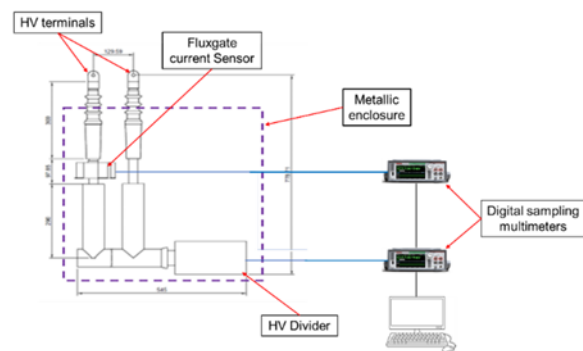


Fig. 1. Schematic layout of the Reference Energy Measuring System.

Fig. 2 shows the characterizations carried out to check the performances of the sensors.

The reference system is used to calibrate EMS installed on locomotives using the main voltage and current components with superimposed odd harmonics, sub-harmonics and inter-harmonics given by phase-fired waveform and the burst-fire waveform [2].

A data processing software has been also implemented to achieve a reliable measurement under extremely dynamic conditions ranging in frequencies up to 5 kHz. An ad hoc measuring software has been developed to process in real time the acquired digital data and determine power and energy components. The model functions for active, reactive energy, no active energy and apparent energy based on IEEE 1459-2010 [4], has been used for determining the energy of the reference measuring system for non-sinusoidal waveforms. These formulas have been adapted to digital sampling in measuring time interval corresponding to n periods after the starting time.

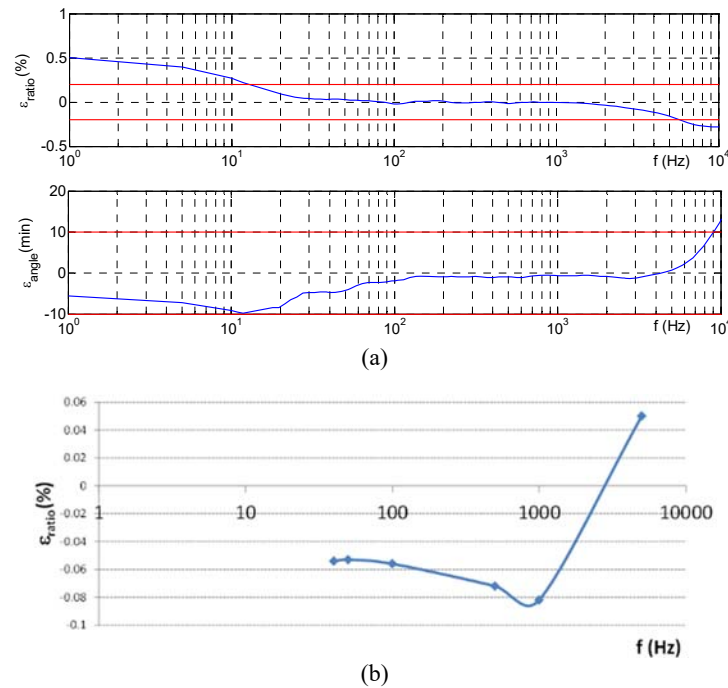


Fig. 2. Frequency response characterization: (a) HV divider; (b) Fluxgate sensor.

A characterization of the developed reference measuring system has been performed in order to achieve the traceability and a calibration to an EMS with its the uncertainty calculations has been performed.

3. Conclusions

The new Reference Energy Measuring System developed in this project will contribute to the improvement of the NMIs metrological capabilities for the calibration of AC voltage and current transducers and power and energy meters used under harsh electrical conditions, improving their reliability.

The Reference Energy Measuring System has been developed for industrial services. A measuring system working in distorted regimes is needed in trains equipped with an EMF, whose energy measurement accuracy has to be assessed and periodically re-verified, as required by EN 50463-2 [2].

The Reference Energy Measuring System has been designed for laboratory and also for on-board calibrations that means a clear advantage for companies and manufactures which demand this service.

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(030)

Intrinsically Non-overfitting, Nonlinear Multivariate Calibration: Application to Thermocouples

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Summary: We present an original and empirically validated approach for nonlinear multivariate data correction and sensor calibration. Although original, our intrinsically non-overfitting method is based on well-established nonlinear algebra. One major application of this approach is temperature compensation of analogue sensors, transducers, voltage references and piezo oscillators. This is particularly relevant for metrological applications outside laboratorial controlled environments (e.g., strain gauge networks for outdoors structural health monitoring). We will show a practical application for thermocouples, where we simultaneously compensate for the adverse influence of the cold junction potential and the temperature-induced drift of a voltage reference. We will present a simple and practical regression method for calculating the transfer function parameter matrix and show its implementation. We will also present a practical methodology for acquiring a representative calibration dataset.

Keywords: Sensor calibration, Multivariate, Nonlinear, Regression, Overfitting.

1. Introduction

Ideally, in order to attain the greatest measurement precision and accuracy, all the variables that account for the degradation of a signal should be measured and corrected for. Naturally, this is only possible for deterministic (i.e., non-random) influences. A prime example of a common adverse influence is temperature: virtually all analogue sensors and transducers are susceptible to this variable, particularly relevant in applications outside the controlled environment of a laboratory.

The use of multivariate calibration methods is relatively common, nowadays, in analytical chemistry [1] (thanks, in great part, to recent advances in machine learning methods). But it is still seldom applied in engineering metrology. This is certainly due to practical complications, among which we can mention the difficulty of identifying and analytically describing (or quantifying) the influence of relevant variables on the signal, and the considerable labor of acquiring calibration data sets for all the relevant combinations of conditions.

Machine learning methods directly address these difficulties. However, these have the tendency to *overfit* the data (i.e., excessive *granularity* caused by noisy data). By contrast, the herein proposed method is virtually immune to this undesirable behavior.

Thermocouples are a prime candidate for the application of our approach. We will demonstrate how it can be used to compensate for the influences of the cold junction, while simultaneously accounting for nonlinear response of these sensing devices (without costly piecewise calibrations) and for the temperature-induced drift of a voltage reference.

1.1. Multivariate Dependence: The General Case

If the major sources of error that deterministically affect a given signal are known and can be precisely measured, it is possible to define a transfer function that returns a good estimation of the true value:

$$y \approx \hat{y} := f(x_1, x_2, \dots, x_n) \equiv \hat{y}(\mathbf{x}), \quad (1)$$

where y is the value we want to measure, $\hat{y}$ its dependent estimator and $\mathbf{x}$ is the vector (or set) of x_i independent explanatory/predictor measured signals (i.e., covariates) [1-3].

Topologically, the continuous scalar $\hat{y}(\mathbf{x})$ function is a $(n+1)$ -dimensional surface defined in a n -dimensional vector space. For the current purposes, we will define $\hat{y}(\mathbf{x})$ as a polynomial surface of degree n^2 , which can be written in the generalized form of a scalar product of two matrices:

$$\hat{y}(\mathbf{x}) = \mathbf{A} \cdot \mathbf{X} \quad (2)$$

where $\mathbf{A}$ is the matrix of constant real coefficients a_{ij} that parametrically defines the calibration surface, and $\mathbf{X}$ is a multivariate polynomial matrix of $(M-1) \times (N-1) = n^2$ degree, where M and N are the matrix dimensions.

Crucially for the current application, a polynomial $\hat{y}(\mathbf{x})$ has injective properties, i.e., it returns a unique value for every unique vector $\mathbf{x}$:

$$\forall \mathbf{x}, \exists! \hat{y}(\mathbf{x}) \quad (3)$$

The underlying polynomial nature of (2) is revealed by writing it in summation notation:

$$\hat{y}(\mathbf{x}) = \sum_{j=0}^N \sum_{i=0}^M a_{ij} x_1^i x_2^j \quad (4)$$

1.2. Simplified Two-dimensional Special Case

Although the proposed approach could be, virtually, generalized to an arbitrary number of covariates or explanatory variables, it will henceforth be limited to only two, for the sake of presentation simplicity. However simplistic, this bivariate approach can be surprisingly sufficient in a great number of practical cases. For instance, in the context of engineering metrology, temperature is typically one of the main causes of measurement errors, affecting the response of analog sensors and transducers, and influencing voltage references and oscillator/clock frequencies. Whenever impractical or unfeasible to conduct the measurements in a temperature-controlled environment, the quality of measurement can be severely impacted. In these circumstances, accurately measuring the temperature and controlling for it, by applying an adequate bivariate correction, can make a decisive difference between invalid and valid results.

Additionally, we will assume that both the sensor response curve and the influence of its single covariate have a simple curvature in the working domain – i.e., no inflexions. Therefore, they can be adequately approximated by a 2nd-degree polynomial. In our experience, this simplification is realistically adequate for a great number of practical applications and, more specifically, for thermocouples (whenever this simplification is not acceptable – i.e., curvature with inflexion – a piecewise solution could be adopted). This assumption will further limit the exponent degree of the approximating bivariate transfer function to 4.

Limiting the analysis to a two-dimensional covariate space, allows us to clarify the meaning of $\mathbf{A}$ and $\mathbf{X}$ in (2), by explicitly expanding them:

$$\mathbf{A}^{(2)} = \begin{bmatrix} a_{00} & a_{01} & a_{02} \\ a_{10} & a_{11} & a_{12} \\ a_{20} & a_{21} & a_{22} \end{bmatrix}, \quad (5)$$

$$\mathbf{X}^{(2)} = \begin{bmatrix} 1 & x_2 & x_2^2 \\ x_1 & x_1 x_2 & x_1 x_2^2 \\ x_1^2 & x_1^2 x_2 & x_1^2 x_2^2 \end{bmatrix} \quad (6)$$

For notational simplicity reasons (which will become evident further on), we will, henceforth, adopt the following variable nomenclature for the two degrees of freedom of our model:

$$x_1 \equiv s \quad x_2 \equiv t$$

Let s and t be, respectively, the *raw* error-containing signal and the explanatory covariate (e.g., temperature). With this naming convention, the previous expressions immediately become less cluttered (albeit less general):

$$\mathbf{X} = \begin{bmatrix} 1 & t & t^2 \\ s & st & st^2 \\ s^2 & s^2 t & s^2 t^2 \end{bmatrix}, \quad (7)$$

$$\hat{y}(s, t) = \sum_{j=0}^2 \sum_{i=0}^2 a_{ij} s^i t^j \quad (8)$$

2. A Lagrangian Solution

We will now describe a practical and efficient procedure for calculating the set of a_{ij} coefficients of matrix $\mathbf{A}$, inspired by a classical noniterative polynomial regression technique.

2.1. The Conventional Approach

The *least-squares method* is the most usual conventional approach for obtaining polynomial regressions. It finds the *best* statistical fit of a polynomial function to a particular dataset, by minimizing the sum S of the squares of the residual errors ε_i , which are the difference between the actual value y_i and the value returned by the approximating function $\hat{y}(\mathbf{x}_i)$ [4-7]:

$$S = \sum_{i=1}^N \varepsilon_i^2 \equiv \sum_{i=1}^N [y_i - \hat{y}(\mathbf{x}_i)]^2 \quad (9)$$

This minimization is usually achieved by applying iterative algorithms like, e.g., the *Gauss-Newton* and *Gauss-Seidel*. These methods can effectively be used to obtain the aforementioned a_{ij} coefficients, directly from the multidimensional problem space.

Implementations of these iterative algorithms are readily and broadly available, since they are included in numerous software platforms (e.g., *Matlab*®, *Mathematica*®, *Mathcad*®, *Maple*®, etc.). In fact, these were used for the initial validation of the herein proposed multivariate calibration technique.

However, despite the formidable technological advances in computing power, these algorithms can be computationally taxing and somewhat cumbersome to implement in simple microcontroller-based solutions, which are still relevant for some applications where cost, power consumption and autonomy are key aspects (e.g., compact battery-operated dataloggers).

There are no alternatives to iterative error minimization algorithms for finding the best approximating function to *overdetermined* datasets (i.e., when there are more data points than the minimum needed to define the approximating curve/function). However, we will demonstrate that it is possible to obtain a less computationally demanding noniterative fitting solution if the number of input points is kept exactly sufficient for defining the approximating function (e.g., three points for a 2nd-degree polynomial).

2.2. Lagrange Interpolating Functions

Before delving into the deeper details of the proposed method, we will introduce the fundamental concepts of Lagrange's interpolation technique that will be applied.

There exists a unique N^{th} degree polynomial that passes exactly through a given set of $N+1$ $[x_i, y_i]$ nodes:

$$y(x) = \sum_{i=0}^N a_i x^i, \quad (10)$$

which is equivalent to say that there exists only one set of $\{a_0, a_1, \dots, a_N\}$ coefficients that satisfies:

$$\begin{bmatrix} 1 & x_0 & x_0^2 & \cdots & x_0^N \\ 1 & x_1 & x_1^2 & \cdots & x_1^N \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 1 & x_N & x_N^2 & \cdots & x_N^N \end{bmatrix} \begin{bmatrix} a_0 \\ a_1 \\ \vdots \\ a_N \end{bmatrix} = \begin{bmatrix} y_0 \\ y_1 \\ \vdots \\ y_N \end{bmatrix} \quad (11)$$

Solving this system of equations – by, e.g., Gauss elimination – will give us the unknown a_i coefficients. However, this is computationally costly. Instead, we will define a series of Lagrange basis L_i functions, such that:

$$y(x) = \sum_{i=0}^N y_i L_i(x), \quad (12)$$

where $L_i(x)$ are polynomials of degree N that weigh the known y_i values, in a way that satisfies the fundamental condition:

$$L_i(x_j) \equiv \begin{cases} 1 & \text{if } i = j \\ 0 & \text{if } i \neq j \end{cases} \quad (13)$$

We then define the L_i functions as:

$$L_i(x) := \prod_{\substack{j=0 \\ j \neq i}}^N \frac{x - x_j}{x_i - x_j} \quad (14)$$

The numerator $(x - x_j)$ guarantees that $L_i(x) = 0$ when x is equal to any of the x_j values, except when $x = x_i$, in which case the numerator becomes equal to the denominator, assuring the fundamental $L_i(x_i) = 1$ condition.

Putting it all together, we get the final expression of the Lagrange interpolating polynomial:

$$y(x) = \sum_{i=0}^N y_i \prod_{\substack{j=0 \\ j \neq i}}^N \frac{x - x_j}{x_i - x_j} \quad (15)$$

Extracting the set of $\{a_0, a_1, \dots, a_N\}$ polynomial coefficients from (15) – i.e., expressing (15) as a power series, in the guise of expression (10) – for the particular case of $N = 2$ yields:

$$a_0 = w_0 x_1 x_2 + w_1 x_0 x_2 + w_2 x_0 x_1, \quad (16)$$

$$a_1 = -w_0(x_1 + x_2) - w_1(x_0 + x_2) - w_2(x_0 + x_1), \quad (17)$$

$$a_2 = w_0 + w_1 + w_2, \quad (18)$$

where the weights w_0, w_1 and w_2 are given by:

$$w_0 = \frac{y_0}{(x_0 - x_1)(x_0 - x_2)}, \quad (19)$$

$$w_1 = \frac{y_1}{(x_1 - x_0)(x_1 - x_2)}, \quad (20)$$

$$w_2 = \frac{y_2}{(x_2 - x_0)(x_2 - x_1)} \quad (21)$$

Rewriting (16) to (21) in a more compact notation:

$$a_0 = \sum_{i=0}^2 w_i \prod_{\substack{j=0 \\ j \neq i}}^2 x_j, \quad (22)$$

$$a_1 = -\sum_{i=0}^2 w_i \sum_{\substack{j=0 \\ j \neq i}}^2 x_j, \quad (23)$$

$$a_2 = \sum_{i=0}^2 w_i, \quad (24)$$

$$w_i = y_i \div \prod_{\substack{j=0 \\ j \neq i}}^2 (x_i - x_j) \quad (25)$$

2.3. Bivariate Lagrange Regression

Lagrange interpolation is, intrinsically, a univariate method which does not lend itself to be generalized to higher dimensions. However, in the conditions of our particular case, it is possible to decompose a bivariate regression problem into two separate univariate ones, as will be demonstrated.

Let us expand and rearrange equation (8) in the following manner:

$$\hat{y}(s, t) = (a_{00} + a_{01}t + a_{02}t^2) + (a_{10} + a_{11}t + a_{12}t^2)s + (a_{20} + a_{21}t + a_{22}t^2)s^2 \quad (26)$$

Looking at (26), we can interpret the bivariate interpolation function $\hat{y}(s, t)$ as a univariate polynomial in s , whose coefficients are given, in turn, by three P_i univariate polynomials in t :

$$\hat{y}(s, t) = P_0(t) + P_1(t)s + P_2(t)s^2 \quad (27)$$

The procedure to find the values of the a_{ij} coefficients of the bivariate interpolator $\hat{y}(s, t)$ starts by acquiring an averaged signal $s(y_i, t_j)$ for each of the nine combinations of three discrete calibration standards $\{y_0, y_1, y_2\}$ (e.g., three reference temperatures) and three averaged discrete values for the controlled covariate $\{t_0, t_1, t_2\}$ (e.g. three *cold-junction*

temperatures), thus obtaining the following calibration matrix:

$$\mathbf{S} = \begin{bmatrix} \bar{s}(y_0, \bar{t}_0) & \bar{s}(y_0, \bar{t}_1) & \bar{s}(y_0, \bar{t}_2) \\ \bar{s}(y_1, \bar{t}_0) & \bar{s}(y_1, \bar{t}_1) & \bar{s}(y_1, \bar{t}_2) \\ \bar{s}(y_2, \bar{t}_0) & \bar{s}(y_2, \bar{t}_1) & \bar{s}(y_2, \bar{t}_2) \end{bmatrix}, \quad (28)$$

which, for notational simplicity, we will write as:

$$\mathbf{S} \equiv \begin{bmatrix} s_{00} & s_{01} & s_{02} \\ s_{10} & s_{11} & s_{12} \\ s_{20} & s_{21} & s_{22} \end{bmatrix} \quad (29)$$

Next, using Lagrange interpolation, we calculate three intermediate 2nd degree polynomials that interpolate the correct(ed) value y as a function of the measured average signal s , for each of the discrete controlled covariate values $\{t_0, t_1, t_2\}$:

$$\hat{y}(s) = \begin{cases} b_{00} + b_{01}s + b_{02}s^2 & \text{if } t = \bar{t}_0 \\ b_{10} + b_{11}s + b_{12}s^2 & \text{if } t = \bar{t}_1 \\ b_{20} + b_{21}s + b_{22}s^2 & \text{if } t = \bar{t}_2 \end{cases} \quad (30)$$

where the nine b_{ij} intermediate coefficients are obtained by the successive application of equations (22) to (25), with the previously acquired (s_{ij}, y_i) set of nine calibration points:

$$b_{i0} = \sum_{j=0}^2 w_{ij} \prod_{k=0, k \neq j}^2 s_{ki}, \quad (31)$$

$$b_{i1} = - \sum_{j=0}^2 w_{ij} \sum_{k=0, k \neq j}^2 s_{ki}, \quad (32)$$

$$b_{i2} = \sum_{j=0}^2 w_{ij}, \quad (33)$$

$$w_{ij} = y_j \div \prod_{k=0, k \neq j}^2 (s_{jk} - s_{ki}) \quad (34)$$

Lastly, the nine a_{ij} coefficients of the final bivariate polynomial interpolator $\hat{y}(s, t)$ are obtained by finding the three $P_i(t)$ univariate sub-polynomials of equation (27), which can be done by interpolating how the previously calculated b_{ij} intermediate coefficients vary in the continuous subspace of the t covariate. This is accomplished by using, once again, the previously derived equations (22) to (25), this time using the nine pre-calculated (t_i, b_{ij}) pairs:

$$a_{i0} = \sum_{j=0}^2 v_{ij} \prod_{k=0, k \neq j}^2 t_k, \quad (35)$$

$$a_{i1} = - \sum_{j=0}^2 v_{ij} \sum_{k=0, k \neq j}^2 t_k, \quad (36)$$

$$a_{i2} = \sum_{j=0}^2 v_{ij}, \quad (37)$$

$$v_{ij} = b_{ji} \div \prod_{k=0, k \neq j}^2 (t_j - t_k) \quad (38)$$

This completes the derivation of the method. In conclusion, equations (31) to (38) define the transformation (or mapping) of a 3×3 matrix of 3D points (or vectors) – in the current context, a set of nine averaged raw values returned by a sensor $(s_{ij}, i, j = 0, 1, 2)$ for all the combinations of three calibration standards $(y_i, i = 0, 1, 2)$ and three discrete controlled values of an error-inducing covariate $(t_j, j = 0, 1, 2)$ – to a 3×3 matrix of scalar values $(a_{ij}, i, j = 0, 1, 2)$, which are the coefficients of a bivariate function that defines a 4th degree polynomial surface that contains the nine 3D calibration points (without residuals). For the present purposes, the key characteristic of this polynomial surface is that the third dimension is injectively dependent on the other two (independent) degrees of freedom. Concretely, the calibrated value $\hat{y}$ is injectively dependent on the raw sensor signal s and the controlled error-inducing covariate t :

$$(\bar{s}_{ij}, \bar{t}_j, y_i) \mapsto a_{ij} : \hat{y}(s, t) = \sum_{j=0}^2 \sum_{i=0}^2 a_{ij} s^i t^j \quad (39)$$

3. A Practical Application: Calibration of Thermocouple Temperature Sensors

Although applicable to the generality of analogue sensors and transducers, the proposed method is particularly indicated to thermocouple temperature sensors. Here are some of the main reasons [2]:

- i. Although considered as *fairly* linear, accurate temperature measurements cannot assume that these sensors have a linear response, especially if the operating temperature range is relatively wide; the common solution consists in using piecewise linear interpolations; however, this method requires a relatively large set of calibration points and can be computationally wasteful if taken to the limit;
- ii. The nonlinear response curve is relatively smooth and has, at most, one major inflexion point; in fact, in many typical applications, there are no relevant inflexions inside the operational temperature range (meaning that the curve can be successfully approximated by a low degree polynomial); but, if these do occur, two alternative solutions can be employed: (a) increasing the polynomial degree and/or (b) definition of distinct piecewise temperature domains (*splining*);
- iii. The major external influence on the signal is the electric potential occurring at the *cold junction*(s) (i.e., the unavoidable bimetal couplings at the contact points with the signal acquisition electronics); this effect can be effectively compensated with the help of an additional temperature sensor;
- iv. Since the physical phenomenon giving origin to the *cold junction* potential is the same as the one

occurring at the sensing junction, it shares the same properties, i.e., smooth response with simple curvature (i.e., few or no inflexions);

- v. As the *cold junction* temperature sensor is usually placed close to the acquisition equipment electronics – and, particularly, the voltage reference used by the analogue-to-digital (ADC) converter – the signal from this sensor can be used to compensate for the combination of both the *cold junction* potential and the eventual temperature-dependent reference voltage drift (i.e., effectively combining two covariates into just one).

3.1. A Real Concrete Example

For the current demonstration, we will use the dataset represented in Fig. 1, acquired from a single thermocouple. Table 1 shows the calibration matrix S , containing the nine calibration points s_{ij} , obtained by averaging the recorded values for three different temperature standards T_{std} and three values of the controlled cold-junction temperature t . The calibration was targeted at measuring negative temperatures in the range $[-25\text{ }^{\circ}\text{C}; 0\text{ }^{\circ}\text{C}]$ with *cold-junction* temperatures in the interval $[18\text{ }^{\circ}\text{C}; 42\text{ }^{\circ}\text{C}]$.

The values shown are the uncalibrated and uncorrected values returned by the 10-bit analog-to-digital converter (ADC). The correction procedure will, inherently, provide an output in proper temperature units (converting the ADC values into temperatures units beforehand would add an unnecessary truncation error to the final results).

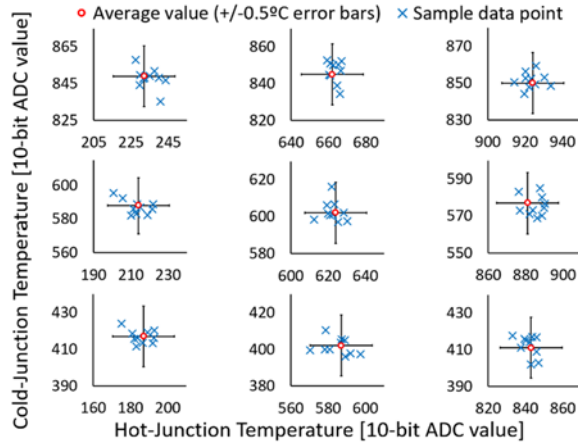


Fig. 1. Example dataset used to demonstrate the calibration method.

Table 1. Calibration matrix.

(s and t values in ADC arbitrary units)			
T_{std} [$^{\circ}\text{C}$]	$t \approx 407$	$t \approx 589$	$t \approx 843$
0.0	843	881	924
-10.6	587	624	662
-21.1	187	214	232

The nine representative points were obtained by averaging at least 10 measurements, with an interval of about 6s between consecutive acquisitions, which totals about 1 minute for each temperature combination (hot and cold junction values). Clear outliers were rejected. The standard deviation of all points was about $\sim 0.17\text{ }^{\circ}\text{C}$. The rejection threshold was heuristically defined as $\pm 0.5\text{ }^{\circ}\text{C}$ (i.e., approximately three times the standard deviation), which is about half the typical accuracy of a *type-T* thermocouple.

The precise temperature standards (used both for calibration and verification) were *ice baths* of *eutectic* salt-water solutions: pure water (0°C) and two eutectic saline solutions, one using potassium chloride (KCl) and the other sodium chloride (NaCl), with solid-liquid transitions at $-10.6\text{ }^{\circ}\text{C}$ [8] and $-21.1\text{ }^{\circ}\text{C}$ [9], respectively. The precise concentrations of these solutions are, respectively, 19.7 % w/w KCl [8] and 23.3 % w/w NaCl [9]. The three discrete cold-junction temperatures used were $\sim 18.0\text{ }^{\circ}\text{C}$, $\sim 30.0\text{ }^{\circ}\text{C}$ and $\sim 42.0\text{ }^{\circ}\text{C}$. This range is adequate for the environmental conditions of our non-climatized laboratory. The cold-junction temperatures were measured with a medium-precision thermistor-based active sensor.

The practical procedure for this calibration consisted in maintaining a fairly constant cold junction temperature, by containing both the electronic circuit and the cold-junction contacts in a thermally controlled enclosure (i.e., an adiabatic box with an internal fan and a heating element), while “dipping” the thermocouples under calibration in the three different ice-baths.

Table 2 shows the resultant matrix A , containing the nine a_{ij} calibration coefficients, calculated from the input values shown in Table 1, using the equations (31) to (38).

Fig. 2 shows the calibration surface generated with equation (8), using the a_{ij} coefficients. The surface includes the nine calibration points – represented as red circles – with zero residual errors.

Table 2. Matrix of a_{ij} calibration coefficients.

-2.306×10^1	-1.798×10^{-3}	1.929×10^{-6}
1.608×10^{-2}	-2.293×10^{-5}	9.778×10^{-9}
2.076×10^{-5}	8.453×10^{-9}	-6.626×10^{-12}

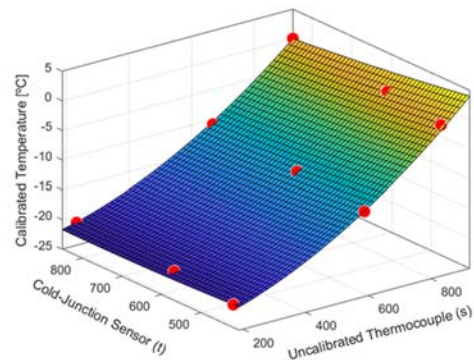


Fig. 2. Interpolated surface containing the 9 calibration points, represented as red circles.

4. Conclusions

The present method has the potential to augment both the accuracy and precision of, virtually, any electronic measurement system based on analog sensors and/or transducers. Being nonlinear, the method allows for a better modelling of the typically nonlinear response of analog electronic devices. Being bivariate, it allows compensation for an external or environmental detrimental influence (typically temperature).

Our practical experience has consistently confirmed the validity, practicability, and usefulness of the proposed method for improving the measurement quality of – specifically – thermocouples. These are inexpensive and fast-responding sensors, capable of covering very broad ranges of temperatures. However, they are notoriously lacking in precision. Besides the electrical noise that plagues their low-level signal, they have significantly nonlinear responses. Also contributing to their lack of precision is the absolute necessity to compensate for the cold-junction voltage, which needs to be measured with precision for best results. The fact that this voltage also varies nonlinearly with temperature adds additional complication. Addressing these drawbacks, the proposed method allows for the compensation of this nonlinear detrimental influence on the signal.

The accurate measurement of temperature by electronic means has always presented a challenge. Accurate temperature reference standards are not abundant and/or accessible, which makes it difficult to calibrate nonlinear sensors. The wider the calibration range, the more difficult it becomes. A nonlinear calibration method is, in these cases, practically mandatory, as it becomes unfeasibly onerous to conduct a piecewise linear approach.

Despite their relatively unwieldy appearance, the herein presented system of equations – which performs the essential transformation of the calibration dataset into the signal-correcting function parameters – is extremely straightforward to implement. It is also very much less computationally taxing than the more conventional iterative alternatives. These assertions are supported by our experience. For instance, we have successfully implemented this method in

microcontroller code, within the development of a data logging device.

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(031)

Sensitive Planar Microwave Diodes on the Base of Ternary $\text{Al}_x\text{Ga}_{1-x}\text{As}$ Semiconductor Compounds

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Summary: Experimental studies of direct current and high frequency electrical characteristics of planar microwave diodes on the base of $n\text{-Al}_x\text{Ga}_{1-x}\text{As}$ layer epitaxially grown on semiinsulating GaAs substrate are carried out. The results reveal that detection properties of the microwave diode strongly depend on AlAs mole fraction x in the semiconductor compound: the voltage responsivity is about several V/W for the diodes with $x = 0$ while it reaches hundreds of V/W when $x = 0.15$ and 0.3 .

Keywords: Planar diode, Microwave detector, Voltage responsivity, Aluminum gallium arsenide structure, Probe station.

1. Introduction

Electromagnetic radiation in microwave and millimeter wavelength range attracts attention of scientists and engineers due to various possible applications in modern fields of technology such as imaging of concealed objects, material homogeneity inspection, adaptive cruise control for automotive vehicles, as well as for future broadband cellular communication networks [1-2]. Successive development of these technologies requires microwave detectors that can be used as sensors of electromagnetic power. For commercial application of these technologies, microwave detectors must not only be sensitive and fast but also reliable and inexpensive. Earlier we have proposed a simple design of inexpensive planar dual microwave diode fabricated on the base of low resistivity gallium arsenide (GaAs) substrate [3]. The diodes were suitable to detect microwave continuous wave signals as well as to measure pulsed microwave power in nanosecond time scale [4]. However, the fast ones had low voltage responsivity, while more sensitive diodes demonstrated high electrical resistance in the continuous wave regime. Therefore, the main issue of this report is to increase voltage responsivity of the planar dual microwave diode and to reduce the spread of its electrical parameters by means of its fabrication on the base of epitaxially grown aluminum gallium arsenide $n\text{-Al}_x\text{Ga}_{1-x}\text{As}$ layer on semiinsulating GaAs substrate.

2. Samples and Experimental

The $n\text{-Al}_x\text{Ga}_{1-x}\text{As}$ layers of submicrometer thickness ($0.4 \div 0.8 \mu\text{m}$) with different aluminum arsenide (AlAs) mole fraction ($x = 0; 0.15; 0.3$) were grown onto a semiinsulating GaAs substrate using liquid phase epitaxy technique. The layers were

non-intentionally doped, and the charge carrier density varied from layer to layer within the range of $8 \times 10^{15} \div 2 \times 10^{17} \text{ cm}^{-3}$.

Rectangular mesas of the microwave diodes were created using optical photolithography technique and wet chemical etching of the epitaxial AlGaAs layer down to the semiinsulating GaAs substrate. The second photolithography step was used to form metallic contacts of the diodes. Thermally evaporated germanium-nickel-gold (Ge/Ni/Au) contacts were annealed in hydrogen atmosphere for 2 minutes. Electrical resistance and voltage responsivity of the diodes were controlled by means of varying the small contact's area on the diode's mesa (Fig. 1).

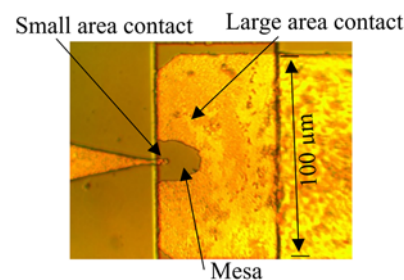


Fig. 1. Photograph of the planar microwave diode.

Measurements of direct current-voltage (I-V) characteristics and high frequency electrical parameters of separate MW diodes were carried out directly on the uncut substrate with the diode array using appropriate probe stations.

3. Experimental Results and Discussion

The I-V characteristics of the dual microwave diodes are presented in Fig. 2. Asymmetry of the characteristics depends on the AlAs mole fraction of the $\text{Al}_x\text{Ga}_{1-x}\text{As}$ epitaxial layer as well as on the area of the smaller contact of the diode. The diodes with

higher mole fraction show stronger I-V asymmetry, what should lead to higher voltage responsivity values due to more effective rectification of microwave currents. The resistances of the measured diodes with $x = 0, 0.15$, and 0.3 are $1.0 \text{ k}\Omega$, $1.5 \text{ k}\Omega$, and $13.0 \text{ k}\Omega$, respectively.

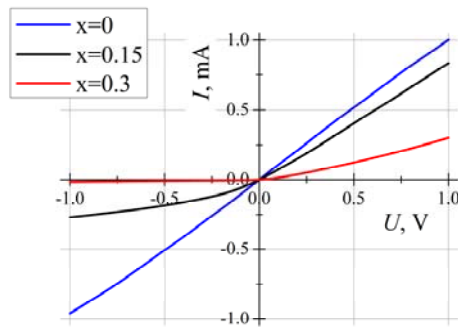


Fig. 2. I-V characteristics of planar dual microwave diodes with different AlAs mole fraction of the epitaxial layer, $x = 0, 0.15, 0.3$.

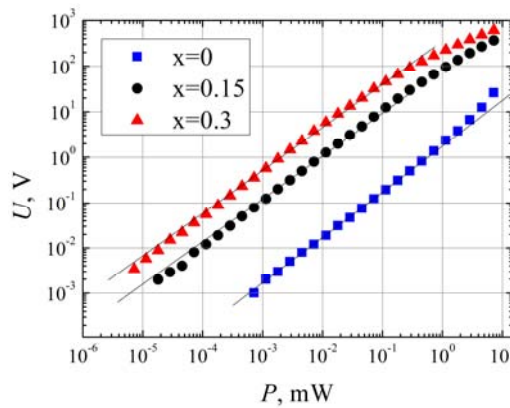


Fig. 3. Voltage-power characteristics of planar dual microwave diodes with AlAs mole fraction $x = 0; 0.15; 0.3$ of the epitaxial layer (lines provide guides to the eye).

Investigation of detection properties of the dual microwave diodes was performed at 30 GHz frequency radiation. The detected voltage of the diodes had polarity typical of that of a metal-semiconductor Schottky junction with negative potential of the voltage induced on the smaller contact of the diode. The voltage-power characteristics of the dual microwave diodes are presented in Fig. 3. The voltage responsivity also strongly depended on the AlAs mole fraction. The lowest values of the responsivity ranging over several volts per watt demonstrated diodes fabricated on the base of GaAs epitaxial layer ($x = 0$). The diodes with $x 0.15$ revealed higher responsivity, from tens up to one hundred V/W, depending on the

area of the smaller contact. The highest values of the voltage responsivity were demonstrated by the dual microwave diodes fabricated on the base of $\text{Al}_{0.3}\text{Ga}_{0.7}\text{As}$ layer: it exceeded 500 V/W in the case of diodes with smaller contact area as low as several square micrometers.

4. Conclusions

The asymmetry of the I-V characteristics and the voltage responsivity of the dual microwave diodes can be raised by increasing the AlAs mole fraction x in $\text{Al}_x\text{Ga}_{1-x}\text{As}$ semiconductor compound. The highest value of the voltage responsivity of the planar dual microwave diodes was achieved when the diode was fabricated on the base of $\text{Al}_{0.3}\text{Ga}_{0.7}\text{As}$ epitaxial layer. Our experimental findings let us conclude that the electromagnetic sensing properties of the planar dual microwave diode can be tailored by proper choice of both AlAs mole fraction in the semiconductor active layer and the area of the smaller contact.

Acknowledgements

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(032)

Automated Setup for Van der Pauw Resistivity and Hall Measurements

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Summary: This contribution reports on an automated setup for measuring the Hall coefficient and resistivity of planar samples by Van der Pauw method. The setup allows the characterization of thin-film samples without modifications of their shape or patterning in a wide range of resistivity values. Hall effect measurements can be performed in magnetic fields up to 1.2 T using a compact permanent magnet of original design. The measurement steps are fully automated by the use of the developed dedicated software. The methodology of measurements and detailed description of the setup are presented.

Keywords: Hall effect, Resistivity, Hall measurements, Van der Pauw technique, Measurement system.

1. Introduction

The complete electrical characterization of conductive thin-films implies the measurement of several parameters, being resistivity and Hall coefficient amongst the most important ones. These parameters play an important role in any field of research concerned with the electronic transport properties of conductive materials, namely semiconductors and metals.

By knowing the Hall coefficient and the material resistivity, one can calculate the charge carrier concentration and its mobility. Important information about charge scattering mechanisms can also be inferred from the temperature dependence and magnitude of the carrier mobility, when Hall coefficient and resistivity are measured simultaneously [1].

Traditionally, the simultaneous measurement of the resistivity and Hall coefficient implies the use of a bar geometry with six contacts, being necessary to know the exact dimensions of the sample and the distance between the contacts. These difficulties can be overcome if the technique suggested by L. J. van der Pauw [2] (VDP) is used. In this technique the number of contacts is reduced to only four, and additionally, one needs only to know the sample thickness to calculate the resistivity.

Generally the commercial systems that implement the VDP technique for combined resistivity and Hall coefficient measurements are costly and not sufficiently flexible to be used in research activities. To overcome these problems, in this work we present an automated compact Hall measurement system that can be built with a fraction of the cost of a commercial system, nevertheless presenting similar technical characteristics.

2. The Van der Pauw Method

Van der Pauw showed, through conformal mapping mathematics [3], that the resistivity of any isotropic,

homogeneous thin-film sample could be determined by using only four probes located arbitrarily around the edge of the sample. In this method, the specific probe location is less critical to achieve reliable measurements, but the knowledge of the sample thickness remains important. Instead of being calculated by the distance between the contacts, the geometric factor is determined by switching the probes used for current application and voltage measurement.

For measuring the resistivity using the Van der Pauw technique, an electrical current is applied between two adjacent probes on one sample edge, inducing a voltage difference between the two additional probes on the opposite edge of the sample. This is shown in Fig. 1(a), the resulting resistance is simply given by Ohm's law: $R_{AC,BD} = V_{BD}/I_{AC}$.

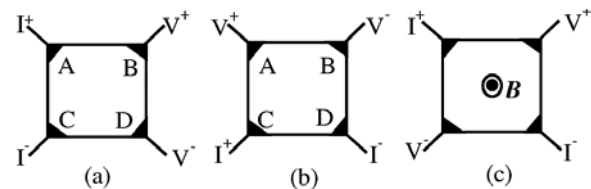


Fig. 1. Contact arrangement for resistance measurement (a) (b) and Hall coefficient measurement (c).

Due to symmetry, reversing the polarity of both the current source and voltmeter or interchanging the voltmeter and current source results in the same resistance being measured. Measuring all four possible combinations of source and voltmeter positions and averaging will reduce the effect of non-ideal contact position/geometry and inhomogeneity of the sample.

In the same manner $R_{CD,AB} = V_{AB}/I_{CD}$ is measured in the configuration indicated in Fig. 1(b) and generally gives a different value than the preceding. These resistances will obey the VDP relation:

$$e^{-\frac{\pi R_{AC,BD}d}{\rho}} + e^{-\frac{\pi R_{CD,AB}d}{\rho}} = 1, \quad (1)$$

where ρ represents the bulk resistivity and d is the sample thickness. The Hall coefficient can be obtained using the same four contacts used in the resistivity measurement. But now the current is passed diagonally across the sample, which is in a magnetic field perpendicular to the sample plane Fig. 1(c).

3. Setup for Hall Measurements

The setup includes a permanent magnet with a stage for positioning a sample holder, and an electronic unit performing measurement operation under the control of a personal computer. The magnet assembly comprises two 70-mm-diameter neodymium (NdFeB) disc magnets and iron housing with a movable cylindrical part for air-gap adjustment. The magnetic

field strength is 1 Tesla at an air-gap of about 9 mm. The sample holder is suitable for samples with dimensions up to 25 mm $\times$ 25 mm.

A block diagram of the electronic unit is shown in Fig. 2. It is based on a simple 8-bit microcontroller (Arduino Nano) which is linked to the PC through an USB port. The microcontroller is also coupled to a current source, a switch matrix sub-unit, a programmable gain amplifier (PGA), and a 16-bit ADC through an I²C bus. The biasing current can be varied from 250 nA up to 100 mA, and the use of an external current source is also an option. The amplifier gain can be set to 1, 10, and 100. All system parameters such as a biasing current, amplifier gain, configuration of the switching matrix, as well as measuring values can be locally monitored through a liquid-crystal display (LCD) on the system.

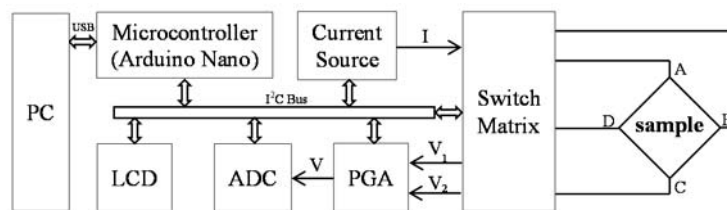


Fig. 2. Block diagram of the measurement system.

The PC software was developed targeting the following tasks: the programmed control of the microcontroller unit; setting and monitoring of all system variables and measurement parameters; data readout, processing, and recording for post-analysis and documentation. The program was written applying the object-oriented programming concept and using VB.NET for the MS Windows platform. The user interface here is intuitive and self-explanative, with the functional blocks for each block, in a way that a rapid look is sufficient to get information about system status.

The setup was successfully used for electrical characterization of transparent conducting oxide thin-films exhibiting a charge carrier density up to 10^{21} cm^{-3} and mobility $\geq 15 \text{ cm}^2/\text{V}\cdot\text{s}$ [4].

4. Conclusions

We have developed an automated setup for measuring the Hall coefficient and resistivity of planar samples. Van der Pauw measurement method is implemented by using common digital and analog electronic circuits and user-friendly software. This setup can be used in scientific and industrial

applications to study the electronic transport properties of metals and semiconductors.

Acknowledgements

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eSPR: New Tool for Screening Cancer

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Summary: In this work, we combined electrochemical techniques with SPR (eSPR) for the label-free detection of cancer biomarker CA 15-3. Detection was performed in a simple two-step assay. In the first step, the gold sensor surface, previously functionalized with CA 15-3 antibody, is incubated with the sample containing the target protein biomarker. In this step, the interaction of CA 15-3 antigen with immobilized antibody was monitored in real-time by SPR. In the second step, electrochemical measurements were performed to improve the sensitivity of the immunoassay, thereby allowing the detection of the cancer biomarker at clinically relevant levels (LOD=0.0998 U mL⁻¹), without sample derivatization or post-hybridization treatment for signal amplification. Good linearity was achieved (R²=0.999) over the concentration range from 0.1 to 500 U mL⁻¹. Furthermore, the developed eSPR biosensor showed high selectivity against other circulating cancer biomarkers.

Keywords: Surface Plasmon Resonance (SPR), Electrochemistry, eSPR, Biosensing, Cancer, Biomarkers.

1. Introduction

Cancer is one of the major public health problems worldwide representing the leading cause of morbidity and mortality in industrialized countries [1].

Despite the impressive advances in tumor biology research that contributed to enlarge the number of potential new cancer biomarkers, such as the microRNAs [2, 3], only few serological protein tumor markers are currently used in routine clinical practice on oncologic patients for the early cancer diagnosis, prognosis and response to therapy [4].

Conventional techniques typically used for screening biomarkers on clinical samples rely on immunoassays [5], requiring labelling or additional reagents to monitor the binding event, making detection rather laborious, time-consuming, expensive and having limited automation of procedures. Thus, the development of new sensing solutions is currently needed in the diagnosis field towards the label-free, fast, simple, cost effective and automated screening of cancer biomarkers, without compromising detection sensitivity, specificity and reproducibility.

This work aims to develop a new fully automated biosensing methodology for the detection of cancer biomarkers in biological fluids, by merging the features of surface plasmon resonance (SPR) spectroscopy and electrochemical techniques (eSPR). SPR affinity biosensors are capable of monitoring interactions in real time between a recognition element immobilized on the surface of a metal film (such as gold) and a ligand (in solution) [6, 7]. Meanwhile, electrochemical (bio)sensors emerged as an alternative to optical approaches, offering fast, robust, low cost and high detection sensitivity [8-10].

2. Methodology

In eSPR experiments, the thin metal film on substrates has the dual function of excite surface plasmons and act as working electrode for electrochemical readout. The combination of SPR and electrochemical detection in the same instrument is technically very simple, since only requires the integration of an external potentiostat to control the potential applied to the cuvette. This set-up offers the unique advantages of rapid, easy, fully automated and reproducible in situ (i) preparation of sensor platforms and (ii) detection procedures, due to reproducible hydrodynamic conditions [11] provided by the SPR robotic auto-sampler (using pre-programmed injection sequences).

The automated detection procedure is performed in two simple steps (see Fig. 1). In the first detection step, direct SPR monitoring of interactions between the protein biomarker (in solution) with surface immobilized bioreceptor is performed until adsorption reaches dynamic equilibrium. In the second detection step, electrochemical measurements are performed in the presence of a biocompatible redox probe. The electrochemical response to the binding event consists on typical amperometric readout employed for detection of non-redox-active biomolecules [8-10].

The described eSPR methodology is label-free and operates without complicated sample treatment procedures or additional complex signal-amplification steps to enhance the sensitivity. As proof of concept, commonly used protein tumor marker CA 15-3 was chosen for the studies. Classical biological antibodies are used for detection with high affinity and results are compared with those obtained using novel (bio)sensing materials assembled at sensor platforms, by means of imprinting (MIP) technology. These

(bio)recognition elements are designed to display a great affinity for each biomarker while cost will be significantly reduced by using more stable artificial sensory materials.

The analytical performance of the eSPR methodology was evaluated by assessing its accuracy, precision, reproducibility, robustness, linearity of reportable concentration range, interfering substances

(selectivity), analytical sensitivity and limit of detection [5]. Initially, the analysis of the biomarkers is performed in synthetic samples and, after optimization of experimental conditions, real clinical samples from oncologic patients are tested. Furthermore, validation of the analytical data is made by comparison to conventional methods (such as ELISA) and making use of suitable statistical tests.

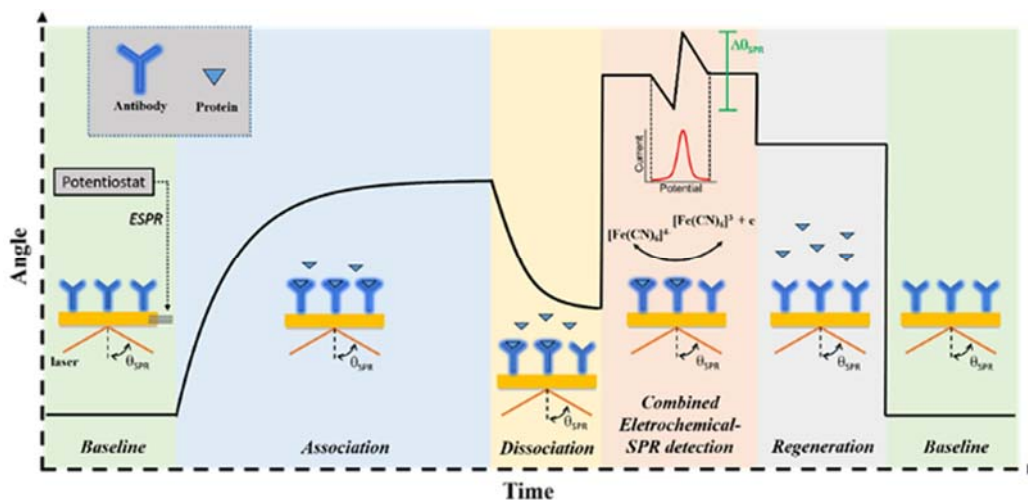


Fig. 1. Schematic representation of the combined electrochemical-SPR detection approach for detection of cancer biomarker CA 15-3, using a biological antibody as bioreceptor.

3. Results

In this work, CA 15-3 antibody immobilization on the gold SPR substrates was achieved by covalent coupling of amine residual groups of the antibody to the COOH groups of a pre-formed SAM on electrode surface, using EDC/NHS activation mechanism [12], resulting in random orientation of antibody binding sites over the sensor surface [13]. After antibody immobilization, unreacted activated ester groups were deactivated via reaction with ethanolamine at high concentration levels avoiding the binding of unwanted biomolecule on the sensing surface [13]. The use of ethanolamine as blocking agent instead of BSA was reported to improve detection sensitivity in electrochemical immunoassays [14].

For effective CA 15-3 antibody immobilization, experimental conditions (association buffer composition, pH, antibody concentration, reaction times, etc.) were carefully optimized. The overall immobilization process of CA 15-3 antibody on the sensor surface, previously modified with a mercaptosuccinic acid (MSA) SAM, is shown in Fig. 2.

Experimental conditions of the immunoassay, such as association buffer pH, incubation time, regeneration conditions, etc., were carefully optimized to achieve improved performance. The label-free and automated two-step immunoassay was carried out by incubation of the sensor surface with various standards with increasing CA 15-3 concentration (from 0.1 to 500 U mL⁻¹). In the first step, SPR technique was used for real-time monitoring of the specific interaction

between CA 15-3 antigen and surface immobilized anti-CA 15-3 under the established optimal conditions, as can be seen in Fig. 3A. Firstly, running buffer solution (10 mM PBS, pH 7.4) was injected into the SPR cuvette and the system baseline was collected for 60 s. Then, the sensor surface was incubated for 15 min with CA 15-3 standard solutions followed by surface wash with background solution to removed unbound molecules. Moreover, this step allowed establishing the interaction time need for adsorption to reach dynamic equilibrium (steady-state conditions) before electrochemical measurements, thus, optimizing incubation times for quicker analysis.

The SPR angle change resulting from the interaction of the antigen with the immobilized antibody on the sensing surface for the various CA 15-3 concentrations tested was used to build the calibration curve, shown in Fig. 3B, after triplicate measurements for each antigen solution. As predicted from SPR equilibrium analysis theory [15], a non-linear dependence between the SPR intensity response with the amount of captured antigen was observed (see inset of Fig. 3B). Thus, logarithmic concentration range (in U mL⁻¹) was employed [15] and two linear antigen concentration logarithmic-SPR response ranges were obtained, for concentrations from 0.10 to 10 U mL⁻¹ ($R^2 = 0.981$) and from 30 to 500 U mL⁻¹ ($R^2 = 0.987$). From the plot, the detection limit (LOD) was estimated to be 21 U mL⁻¹. The LOD obtained was estimated according to Recommendations of IUPAC for ion-selective electrodes [16].

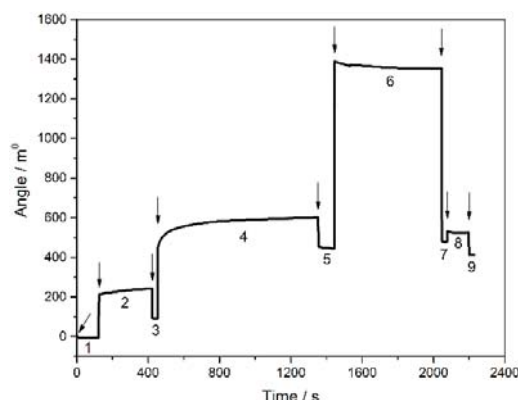


Fig. 2. Representative sensorgram showing the different steps involved in the immobilization of CA 15-3 antibody on MSA-modified SPR gold disk. Line 1: baseline collected in 10 mM PBS, pH 7.4, for 120 s; Line 2: activation of carboxylic-acid-terminated SAM with a mixture of EDC/NHS (50 mM, in water) for 300 s; Line 3: wash with 10 mM PBS, pH 7.4, for 30 s (return to baseline); Line 4: immobilization of 20 $\mu\text{g mL}^{-1}$ CA 15-3 antibody, dissolved in 10 mM acetate buffer solution, pH 4.0, for 900 s; Line 5: wash with 10 mM PBS, pH 7.4, for 90 s (return to baseline); Line 6: deactivation of remaining reactive esters with 1 M ethanolamine (pH 8.0) for 600 s; Line 7: wash with 10 mM PBS, pH 7.4, for 30 s (return to baseline); Line 8: wash with 10 mM PBS, pH 7.4, containing 0.1 % SDS, for 120 s; Line 9: wash with 10 mM PBS, pH 7.4, for 30 s (return to baseline). The arrows indicate the injection of the different solutions into the measuring system.

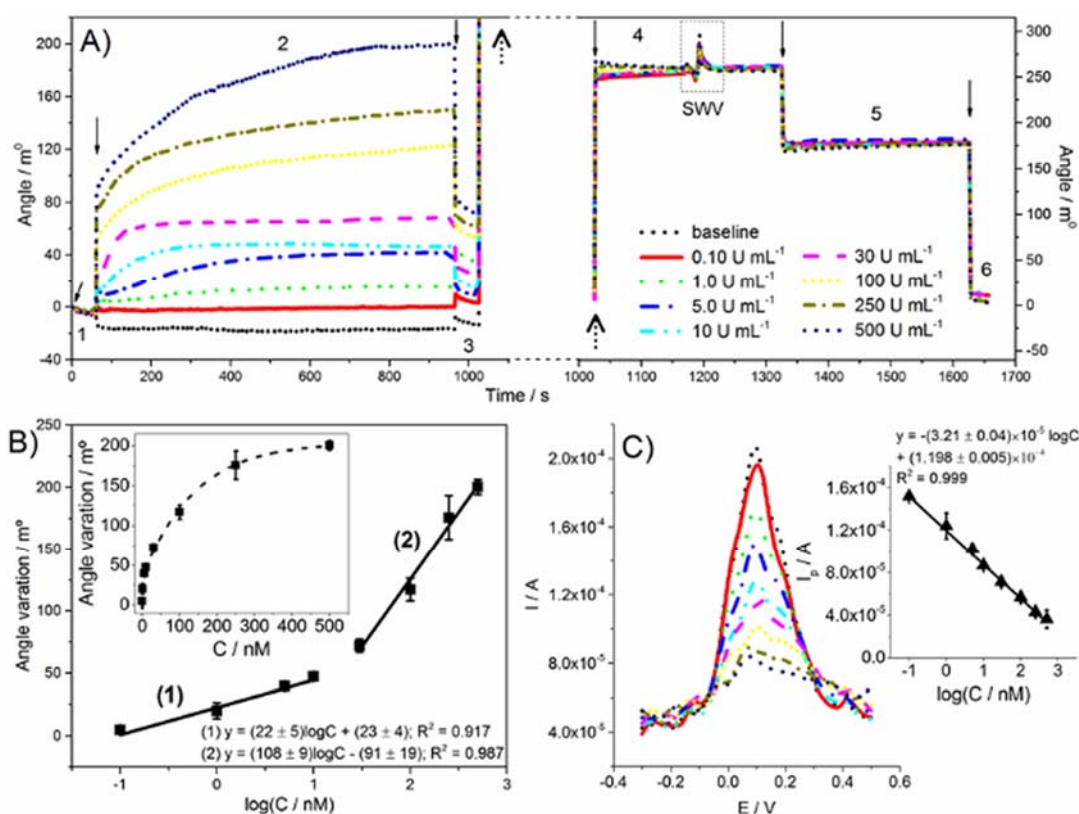


Fig. 3. (A) Real-time monitoring of the interaction between immobilized antibody on SPR gold substrates and CA 15-3 by SPR (left), followed by electrochemical control (SWV)-SPR measurement in the presence of a redox probe (right). The concentrations of CA 15-3 tested ranged from 0.1 to 500 U mL⁻¹. Line 1: baseline collected in 10 mM PBS, pH 7.4, for 60 s; Line 2: real-time monitoring of the interaction between immobilized antibody with CA 15-3, dissolved in 10 mM acetate buffer solution, pH 4.0, for 15 min; Line 3: surface wash with 10 mM PBS, pH 7.4, for 60 s (return to baseline); For comparison, the response obtained after injection of association buffer (without CA 15-3) into the measuring channel (dashed line) is also shown in the figure; Line 4: combined electrochemical control and SPR detection after injection of 5 mM ferro/ferricyanide redox probe into the measuring channel. After 150 s equilibration, a SWV measurement between -0.3 – 0.5 V (vs. Ag/AgCl) was performed; Line 5: surface regeneration after wash with glycine-HCl buffer solution (pH 2.0) for 300 s; Line 6: wash with 10 mM PBS, pH 7.4, for 60 s (return to baseline). The arrows indicate the injection of the different solutions into the measuring system. (B) Graphical representation of the angle variation as a function of the CA 15-3 concentration (inset) and logarithm of the CA 15-3 concentration obtained by SPR monitoring. (C) SWV voltammograms collected in step 4. Inset: graphical representation of the redox probe peak current (I_p) as a function of the logarithm of the CA 15-3 concentration.

In the second detection step, electrochemical measurements in the presence of a biocompatible redox probe were performed after direct monitoring of the binding event. Square wave voltammetry (SWV) technique was selected for the amperometric readout since this technique has been widely used for the detection of biomarkers with high sensitivity and much faster measuring time (few seconds only) comparing to other electrochemical techniques [17]. To do so, a 5 mM ferro/ferricyanide solution was injected into the measuring channel in first place. After 150 s equilibrium, where no potential was applied in the cell, the current generated by the oxidation of the redox probe was recorded in potential range from -0.3 to 0.5 V (vs. Ag/AgCl). The voltammograms were collected under hydrodynamic conditions since the Autolab ESPRIT integrated syringe is continuously aspirating and dispensing the redox probe solution into the cuvette, which increases enormously the mass transport at the surface. Finally, the redox probe was washed-out from the cell and regeneration of the sensing surface was achieved by incubation with a 0.1 M Gly-HCl solution (pH 2.0) for 5 min, followed by surface wash with PBS.

The SWV voltammograms collected are displayed in Fig. 3C and the resulting redox probe peak currents were plotted as function of the increment of CA 15-3 logarithm concentration (see inset of Fig. 3C). SWV measurements showed that peak currents decreased with increasing of CA 15-3 concentration due to electron transfer inhibition after the formation of antibody-antigen immunocomplexes at the electrode surface. Good linearity of calibration curve ($R^2 = 0.999$) was achieved over the concentration range from 0.10 to 500 U mL⁻¹, resulting in better accuracy when compared with detection based on SPR response. The LOD obtained was found to be 0.0998 U mL⁻¹. Thus, the sensitivity of the electrochemical approach for screening CA 15-3 antigen is much superior to the obtained by direct SPR sensing, emphasizing the important contribution of electrochemical detection to plasmonic biosensors towards the quantification of low levels of cancer biomarker. These results are consistent with other eSPR immunosensors reported in literature where lower LOD was achieved by electrochemical readout relatively to SPR detection [15].

Overall, in our opinion, our simple and label-free eSPR methodology can provide an extremely attractive way of sensing CA 15-3 from the sensitivity, time of analysis and operational point of views.

4. Conclusions

The proposed eSPR methodology for screening cancer disease can accurately and rapidly report severe health hazards, in a simple and cost-competitive manner. Taking into account the rigorous validation process of a candidate cancer biomarker test [5], we expected that, at the end of our work, this new detection approach offers the (i) analytical validity, (ii) clinical validity and (iii) clinical utility to progress

to routine use in clinical laboratories, contributing to improve public health and quality of life.

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Application of Low-cost Pulse Radar for Heart Rate Detection in Vehicle Interior

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Summary: Driver and passenger integrity are important topics for passive safety research and recent developments on vehicular technology applied to interior monitoring may open new opportunities for new assistance systems that provide important information about the passenger's health condition. This paper characterizes the usage of a low-cost and low energy consumption single radio-frequency pulse radar as a mean to assess driver's respiration and heart rate frequency inside a vehicle. Tests with real users were performed and the accuracy of the radar evaluated by comparison with a ground truth (ECG). Two signal processing techniques (FFT and EMD) were applied and compared in three distinct scenarios (normal breathing, holding breath and after exercising). For the heart rate measurement, results revealed that FFT yielded outputs with a better curve fitting than the EMD for both time windows. Furthermore, lower average error was encountered when the participants were breathing normally compared to the other two scenarios.

Keywords: Radar, Vital signal, Signal processing, Physiological supervision, Vehicular technology.

1. Introduction

Radar is a robust technology that has found different applications, such as in automotive technology, industrial surveillance and biomedical assessments [1].

Radar is commonly used for outside surveillance in the object detection or tracking. Due to its ability to scan the environment and collect information on multiple channels, the use of radar has been considered very promising in studies focusing autonomous driving and driving assistance systems [2]. New fields for radar application keep rising, as in the vehicle interior, focusing on enabling passenger detection and tracking, for recognition of passenger's activities, or even infer its physiological state.

Up to now, for this purpose, continuous wave (CW) or impulse radars were used [3], however the use of radar type is many times limited due high energy demand and its elevated costs. For this reason, the present work analyzed the use of a low-cost pulse radar in the assessment of drivers' respiration and heart rate frequency. The sensor used in this work is approximately 3× more economical than the second comparable sensor and provides open development kit for acquisition and processing of data.

This paper is structured in the following form: first a general review on the topic and premises assumed, presented in this introduction section, followed by a review of the radar systems in Section 2. Section 3 presents the methodology for the experiments and Section 4 about the processing techniques for the signal. Finally, Sections 5 and 6 bring the results and discussion and the conclusion, respectively.

2. Radar

As stated in the literature, the radar system transmits a wave and detects the scattered echoed signal. The receiving antenna detects the energy of the reflected wave coming from the target, and with this information it is possible to measure the distance and relative velocity of the target. The energy received by the radar depends on a series of properties from the radar itself and the target. One important information regarding the target is the Radar Cross Section (RCS), which can be described as the portion of the target that reflects the wave [4].

1.1. Pulse Radar

The pulse radar has been mainly used for long distances, the detection of targets with small precision movements was not, until now, a main goal for its application [5]. It emits electromagnetic waves (EM) of very short time duration (pulse width), being the receiver isolated from the transmitter's high-power electromagnetic waves. After the transmission, the receiver connects to the antenna and is able to receive the signal. The sum of the transmitting and receiving time is the pulse radar time cycle or Pulse Repetition Interval (PRI). The Pulse Repetition Frequency (PRF) is the number of cycles completed per second and measured in Hz and it is the inverse of the PRI [6].

The pulse radar is a time-of-flight sensor, measuring the distance based on the time the echoed wave takes to return to the system. The distance to the object can then be calculated by multiplying the time- of-flight with the speed of the radio wave (the

speed of light), and then dividing by two. The travel distance of the wave equals two times the distance to the object [7]. Due to the range being calculated from this delay time, problems can occur if the target is at a distance such that the time for the wave round-trip is greater than the Interpulse Period (IPP). In this case, the returned EM, will not arrive before the next pulse is transmitted, resulting in something called a range ambiguity. This means that, the reflection can be either from the recently transmitted pulse due to close target, or from the previously transmitted and reflected by the further target [6].

2.2. Health Concerns

As high doses of radiation can be harmful for human beings, when using a radar inside the vehicle one has to be sure of the equipment's innocuity. Considering the type of radar used in the present study, the intensity field of the electromagnetic wave is measured in W/m^2 , needing $1000 W/m^2$ to produce harm. The Acconeer radar used in the study was tested against the normative US/FCC 47 CFR part 15. This normative regulates the requirements for radio spectrum usage and sets the maximum permissible exposure at 20 cm to be $0.1 W/m^2$. The used radar was also tested against the European Union normative (UE/CEPT) of The European Commission Radio Equipment and respect the limit values given by this regulation. More information on the test results can be found at the equipment datasheet [7].

3. Related Works

The goal of implementing a technique to detect breathing and heart rate overlaps with other applications that aim to detect presence and motion of passengers. Ivashov et al (2004), [8], tested a 1.6 GHz radar to detect breathing and heartbeats behind a wall, application particularly useful for search and rescue of humans after a catastrophe or accident. In the same scenario, an ultra-wideband (UWB) radar was used for life detection in the bands of 4 to 7 GHz by Li and Lin (2008), [9]. For more direct measurements, such as patient supervision, higher frequencies were already used by Obeit et al. (2011), [10], the output of a 16 GHz CW radar was compared to the measurements performed with an ECG. Also making use of a continuous wave, Zhang (2017), [3], analyzed a 24 GHz radar to find both breathing and heart rate of targets in several positions. With the advances on research and experience using radars for such purposes, lower frequencies have been avoided due to the narrow band interference, null detection points and high interference from body movement [3]. Following this tendency, the present research was done with a higher frequency radar (60.5 GHz). Furthermore, due to the fact that it is a pulsed radar (saves energy, which is an important aspect for the characteristics of measurements conducted in a vehicle interior), it will

allow feasibility and performance comparisons to the other CW radars used in previous studies.

3. Methodology

The aim of this research was to evaluate the feasibility of using a low-cost pulse coherent radar to measure the heart rate and breathing rate of drivers inside a vehicle. This work supports the efforts in finding reliable lower cost and contactless sensors to track the physiological parameters of drivers. For that, use cases were defined and 10 subjects were invited to take part in the study in the premises of the research and test center CARISSMA.

The equipment used in the experiments was an Acconeer™ XM112 radar module sensor and XB112 breakout board. This radar has a carrier frequency of GHz, with band- width from 57 to 64 GHz, with the range of the measurement between 0.2 and 0.8m and an update frequency of 20 Hz. The Acconeer radar is coherent, using a combination of low-energy consumption with the accuracy of a coherent system. For this reason, the IQ mode was used in this project, where both the phase and amplitude of the signal are measured. The returned signal value is a complex number, and allows the system to detect small changes in distance with μm accuracy or efficient background cancellation [11].

The measurements were made inside a real vehicle (Mitsubishi Carisma) parked in a Laboratory. The radar was fixed in the windshield of the vehicle by a suction cup and a PLA structure (designed in Catia V5™ and 3D printed), in such a fashion that it was possible to adequate the angle of the sensor accordingly to the subject. The distance between the radar and the subject was approximately 0.6 meters. As ground-truth, an ECG device (shimmer3) was used to perform accurate heart rate and respiration rate measurements.

A homogeneous group of subjects was recruited to take part in the experiments. The sample was then composed by 10 males with ages between 20 and 45 years old. Upon arrival, subjects were informed about the goals of the study and a detailed explanation of the procedure was given. The study procedure was composed by one unique session (lasting approximately 30 min.), where five measurements were made in three distinct scenarios. For all scenarios, the subject was seated on the driver seat (siting angle approximately 110° , facing the radar):

- Measurement 1: Subject breaths normally for 60 seconds (scenario 1);
- Measurement 2: Subject holds breath for as long as he can (scenario 2);
- Measurement 3: before performing the measurement, subjects steps out of the vehicle and performs 1'30" of cardio exercise, to increase heart frequency. Immediately after, subject returns to the vehicle and measurement is performed for approximately 1 minute (scenario 3);

- Measurements 4: repetition of scenario 1;
- Measurement 5: repetition of scenario 2.

The comparison between both sensors was made by calculating the root-mean-square-error (RMSE) of both dependent variables (heart rate and respiration frequency) between both sensors (radar and ECG) using the mean frequency for the same 15 seconds window of the signal. The window size with which the signal was processed is dependent on the time and acquisition frequency. To perform this task, both signals were synchronized in a way that both radar and shimmer work along the same timestamp. Further on, the whole signal will be evaluated comparing with the radar point to point.

4.1. Signal Processing

All algorithms to process the radar data were built in Matlab. Python was used for applications related with the ECG data processing and visualization of to specific libraries, such as Biosppy [12].

4.1.1. Data Structure

The radar data structure is dependent on the distance of measurement, where each bin has 0.05 mm. As Fig. 1 illustrates, the distance is represented in the N axis, with every range bin covering a space based on the selected wavelength. The time duration, on the M axis, increases the number of pulses, thus, yielding an M×N matrix. Fig. 1 also shows two targets, one in the same resolution bin (target 1) and one moving in different resolution (target 2). Fast time means the velocity of acquisition of each range bin inside a pulse and slow time is the time between pulses.

4.1.2. Short Time FFT algorithm

This method was used to process the raw data obtained from the radar. It relies on finding the highest peaks of the spectrum after applying the Discrete Fourier Transform (DFT) across the time direction to find breath and heartbeat frequencies. The DFT was applied using a Fast Fourier Transform (FFT) in a windowed signal, to clear the peaks in the frequency spectrum. Different windows were applied and the best fit was verified. The Blackman was the one showing the best results.

Firstly, a presence detection step to find in which range bin the target is located was performed, evaluating the peak position in one pulse. This is assumed to be the position of the driver, and also (due to the methodology used for data collection) to be the same throughout the measurement. Ergo, the pre-processing of the signal and search for the breath frequency is performed, applying an IIR Butterworth band-pass filter between 0.12 and 2.5 Hz, removing the DC component, windowing and applying the FFT to estimate the frequencies present. After the FFT, the frequency with the highest magnitude in the spectrum was taken as the breathing rate.

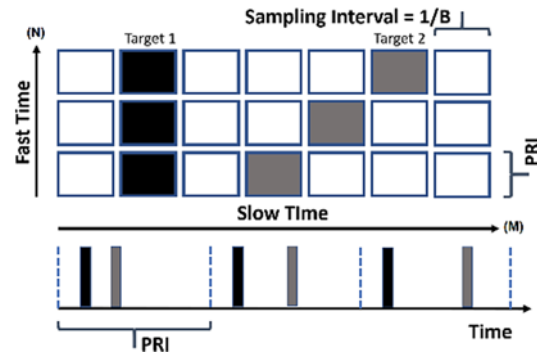


Fig. 1. Radar matrix structure (being B: bandwidth and PRI: Pulse Repetition Interval).

For the heart rate, first a notch filter was applied to remove the estimated breathing rate signal and its harmonics from the original signal. This was done due to the ratio of magnitude of the breathing signal to the heart signal being very high. Then a band-pass IIR butterworth filter with cut-off frequency between 1 and 4 Hz followed by a window were applied and from the FFT. In the selected range bin across the pulses, the heart frequency was then estimated.

4.1.3. EMD with Euclidean Distance

EMD stands for Empirical Mode Decomposition and it is a method that allows to separate the observed signal into its Intrinsic Mode Functions (IMF) and a residual function. Each IMF has one instantaneous frequency and it represents one vibration mode. The Euclidian distance was used in the present work to compute the distance between oscillatory mode and the original radar signal.

Again, the position was estimated and then the pre-processing applied. Firstly, the respiration signal was processed, the EMD was applied to split the signal into its IMFs and a window was applied to each IMF. An FFT followed, and the maximum peak held the estimated rate of each IMF. The Euclidian distance was used to select which IMF is closer to the original signal.

Consequently, the notch filter was applied to remove the breathing frequency and its harmonics from the signal. A new EMD was applied on the signal to find new IMFs, window and FFT were applied to the signal to find the frequencies of the IMF. The Euclidian was used to select which IMF is correct.

5. Results

In this section, results of the analysis regarding signal processing are presented and discussed. The analysis performed took in consideration 10 subjects completing the experiment. The outcome of the FFT and EMD methodology applied in the radar signal were compared against the ECG, which serves as golden standard for the measurement. Results reveal the average frequency present in the window of signal, both for the ECG and radar measurements.

As the window time affects the warm-up time of the sensor (the time required to provide the first measure), both time windows were considered to evaluate their suitability in evaluating the signal accuracy. Biosppy library [12] for python, was used to select the most adequate lead ECG signal, i.e. which electrode superposes the heart waves in the best manner, leading to the best QRS complex match.

5.1. Analysis Procedure for One Subject

To illustrate the procedure, for a 10 seconds window, the graphical results of one subject for scenario 1 are displayed.

In what regards the evaluation of the heart rate signal, the first step was the ECG signal processing for the selected heart electrode. The raw signal was initially processed with a Blackmann window followed

by a FFT to find frequencies components. The raw and windowed FFT signal is shown in Fig. 2. The peak value of the frequency spectrum points for a frequency of 1.0996 Hz, which corresponds to a heart rate of 63.996 BPM. This value was taken as the real mean value of the heart rate in the given time window. An example is shown in the table below.

On Fig. 3, it is shown the processing using the FFT method for the radar on the same time window. On the left side, it is possible to see how the signal behaves across the selected range bin for the target position in time. On the right side, the signal in the frequency domain, which highest peak shows a HR frequency of 1.100 Hz that corresponds to a heart rate of 66.00 BPM. This measurement results in an error of 0.0334 per second, which corresponds to 2.004 BPM/s and 3.13 %, related to the ECG.

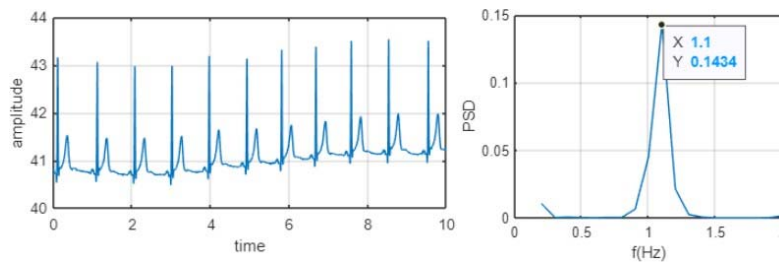


Fig. 2. ECG heart signal filtered (left) and on frequency domain (right).

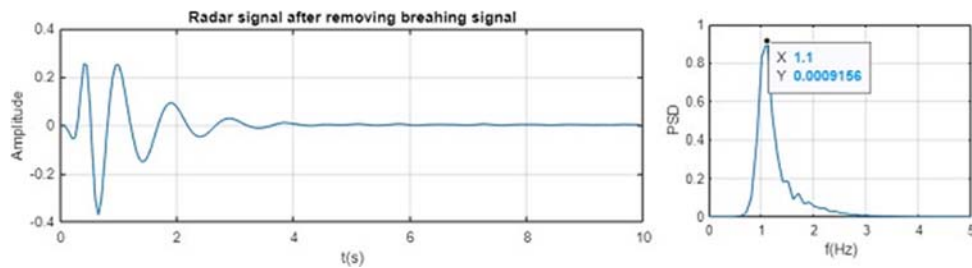


Fig. 3. Radar signal raw (left) and on frequency domain (right).

For the EMD radar signal analysis, the frequencies of each vibration mode found are 1.100, 1.000 and 0.4000, Fig. 4. By the Euclidian distance, IMF 1 represents the signal more effectively, Fig. 5, yielding a 1.100 Hz, which represents 66 BPM and, by comparison, an error of 0.0334 per second or 2.004 BPM/s (3.13 %).

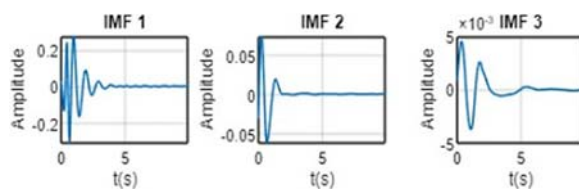


Fig. 4. Intrinsic mode function of the radar signal.

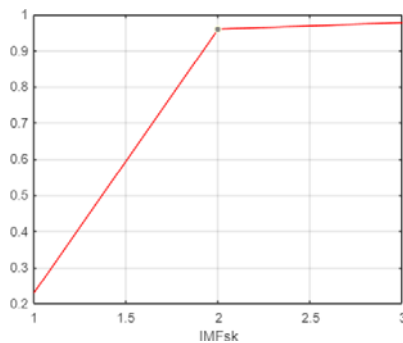


Fig. 5. Euclidian distance of IMFs.

Using the same procedure as before, the analysis of the respiratory rate was performed for the subject yielding a frequency of 0.2999 Hz, which corresponds to 17.994 respiration per minute (RPM).

The FFT analysis of the radar signal showed that the frequency spectrum peaks at 0.2600 Hz signaling 7.2 RPM. The comparison with the ground truth evidenced an error of 0.0399 per second (13.30 %).

Considering the EMD, the IMF frequency values for each vibration mode are 0.200, 0.200 and 0.200 Hz. According to the Euclidian distance, IMF 1, showed the smallest distance to the original signal, and was therefore taken as the breathing signal, with values of 0.2 Hz, 12.0 RPM. Referring to the ground truth, it presents an error of 1×10^{-4} per second (0.05 %).

5.2. Mean Heart Rate (HR) and Respiration Rate (RR)

The procedure explained in the previous section was performed to all 10 subjects taking into account all scenarios: breathing normally (scenario 1), holding breath (scenario 2) and after performing exercise (scenario 3). The analysis procedure was conducted to both 10 and 15 seconds time window. A summary of the average error resulting from the comparison between radar signal and ECG is presented in Table 1. Considering the heart rate (HR) mean error for the 10 second window, it is possible to verify that the FFT method was the one delivering lowest mean errors for all three scenarios. A comparison of the three scenarios evidences that scenario 1 (breathing normally) in the FFT analysis was where the deviation from the ground truth was lower. On the other hand, scenario 3 (measurement conducted after performing exercise) had the highest HR mean error.

Table 1. Mean error values.

Window	10 seconds				
Scenario	1	2	3		
Measurement	HR	RR	HR	HR	RR
FFT error(%)	7.09	19.31	11.56	15.23	13.47
Std FFT (%)	5.57	7.88	5.72	10.46	12.39
EMD error (%)	13.81	17.14	13.47	16.98	15.35
Std error (%)	9.93	16.57	5.91	5.55	12.2
Window	15 seconds				
Scenario	1	2	3		
Measurement	HR	RR	HR	HR	RR
FFT error(%)	10.65	26.28	12.9	15.23	13.47
Std FFT (%)	4.54	8.91	6.04	7.98	9.26
EMD error (%)	12.12	19.20	16.74	14.77	16.95
Std error (%)	5.89	11.47	8.73	9.99	13.54

An analysis of the HR for the 15 second time window reveals, comparatively to the 10 s analysis, higher error values. Here the FFT method yielded the best performance. Again, scenario 1 was the one showing the lowest mean error value, however scenario 2 was the one revealing the highest heart rate measurement differences.

Considering the respiration rate (RR) results, for the 10 s window analysis, it is possible to note that in scenario 3, the FFT approach presents a superior performance than EMD but a higher standard deviation. However, for scenario 1, EMD presents a smaller average error but superior dispersion of measurements.

Likewise, examining measurements for the 15 s window, the preferred method for the first scenario was the EMD, with a smaller average error but greater standard deviation. For Scenario 3, EMD again, outperformed, with a lower average error and, again, higher standard deviation.

Comparatively to the 10 seconds analysis, the 15 s window presented the higher average error, for both procedures. For the heart rate the FFT approach yielded better results in both windows, having its better performance in 10s and for respiration, EMD outperformed in the 10s window as well. The mean error values for all subjects and scenarios are shown in Table 1.

5.3. HR Comparison Across Time

In order to track how the heart rate measurements changed across time a comparison was made, point to point, using both windows of 10 and 15 seconds, the three measured scenarios and both analysis methods. Table 2 shows the values for the absolute mean error, the relative error, the MSE and standard deviation of the measurements.

Table 2. Mean error values – complete signal.

FFT					
Scenario	Window(s)	Abs	Rel(%)	RMSE	Std
1	10	11.48	22.46	32.50	11.73
	15	9.46	13.75	22.27	9.76
2	10	11.02	14.76	23.44	10.367
	15	10.05	14.29	16.73	8.37
3	10	13.00	17.16	40.03	12.97
	15	9.78	19.95	23.10	9.64
EMD					
1	10	11.35	16.75	36.28	13.20
	15	10.95	16.20	26.34	11.71
2	10	12.00	16.87	29.93	12.56
	15	11.69	16.73	21.28	9.86
3	10	12.43	17.92	40.69	14.31
	15	9.11	15.46	26.20	11.71

According to the Table 2, the 15 second window presented a smaller value when compared to the 10 seconds in all metrics that is expected since the amount of information in the signal is bigger. Small difference was perceived between the scenarios in absolute error values, but the MSE value shows that with an increased heart rate the both methods perform poorly for 10 seconds window, but remains stable for 15.

6. Conclusions

The present study was conducted using real conditions and real radar detection data from different subjects under distinct conditions as a mean to investigate feasibility of using a low-cost pulse radar in vehicle interior for recognition of driver's vital parameters.

In this paper algorithms were applied to support the contactless estimation of breathing and heart rate measurements. Two main signal processing pipelines were developed using FFT and EMD methods. The given results corroborate the idea of radar as a potentially useful health assessment tool and shows that it is possible to lower the costs and power usage for future implementation in automotive context. To do so, further research needs to be developed, as well as improvement of the current used methods.

Results revealed that although the FFT and EMD yielded similar absolute accuracy, the FFT presented smaller error values, which indicates better data fitting of the prediction. For the breathing scenario, the respiratory rate had higher mean error values for both windows, with a substantial standard deviation from the mean. There was also no clear tendency on which method yielded better accuracy as this was highly influenced by the scenario. Such outcomes points out to a very volatile measure.

Additionally, in scenario 1 (breathing), it is possible to notice that, for the heart rate, a higher accuracy was obtained with a bigger window in both approaches. That can be extended also for not breathing and exercises scenario, where all measurements presented a smaller error for 15 seconds. The scenario with the highest accuracy is also the one with the most stable behavior of the target. The acquired signal has the regular variability expected from the signal, but points to the importance of tracking the peaks of the signal and identification of what might be other body movements.

Between the 2 models used to find the information in the radar signal, for the 10 seconds window and heart rate, FFT presented a smaller error value or similar to EMD, overall. For 15 seconds, and heart rate, EMD had a better performance and even at the exercise scenario it pointed to a considerable improve of results. However, for the complete signal, FFT has proven to be more consistent with its results, especially for the 15 second windows. Yet, the results in the best case still points to a 14 % error regarding the ground-truth. Nonetheless, this result is consistent with previous works using CW radars, pointing to the ability to perform the measurement under certain conditions, although not yet satisfactorily as other devices.

Other techniques can be applied to further increase accuracy of measurement. However, for the time being, this work demonstrates that low cost radars have the potential of being used for physiological state assessment system if improvements are performed, especially in atypical conditions (breathing frequencies significantly different than normal).

Besides the presented techniques in this work, another method of selection of interest frequency peaks should be attempted, in order to improve results in cases of a frequency spectrum with several peaks. The presence of any movements in the same frequency dilutes the information and can drive the measure to errors. One alternative could be the distance of the chest and the tracking its movement and phase shift, as observed in the graphical interface of the radar kit [1]. With further work and improvement of techniques, as wavelet approach or a regression model, it would be interesting to investigate possibility to improve further on the results and have the radar working in less constrained conditions.

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Calibration of a Low Cost Sensor for PM_{2.5} using a Reference PM Monitoring Station

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Summary: Air pollution is related to the cause and aggravation of several diseases as well as to producing ecological damage to the environment, soil and water. One of the most important pollutants with greatest health concern is Particulate Matter (PM). A very popular and low-cost PM sensor is the Shinyei PPD42NS. It is based on the light scattering method. A reference PM monitoring equipment approved by U.S.A. EPA is the Ecotech HiVol 3000, which uses the gravimetric method for PM concentration measurements. In general, gravimetric-based monitoring equipment are large, require constant maintenance and a trained team to handle them and optical-based sensors like the PPD42NS are very smaller and can be distributed to facilitate the detection of PM concentration with spatial and temporal resolution. This paper describes a calibration procedure of the PPD42NS sensor using Ecotech HiVol 3000 as the reference. In addition, a performance evaluation of the PPD42NS is given related to its accuracy. The obtained main experimental result is the obtained linear correlation was greater than 0.9, indicating a strong association between them.

Keywords: Air pollution, Particulate matter, Calibration process, Low-cost optical sensors.

1. Introduction

Air pollution is a major environmental problem associated with several diseases in humans and ecological damage to the environment, animals, soil and water [1].

The air contains a complex combination of pollutants, which depends on atmospheric conditions and sources of emissions. Short and long-term exposure can mainly cause heart disease, lung disease and respiratory infections [2, 3].

The main sources of pollutants in the atmosphere are means of transport, such as cars and trucks, and industries, mainly chemical, metallurgical and paper industries [4]. The pollutants that cause greatest health concern are Particulate Matter (PM), Ozone (O₃), Nitrogen Dioxide (NO₂), Sulfur Dioxide (SO₂), Lead (Pb) and Carbon Monoxide (CO) [4].

Particulate Matter (PM) is formed by solid and liquid particles from different sources, such as combustion centered on carbon and secondary inorganics, and is related to several respiratory and cardiovascular health problems [5].

In addition to presenting health problems in humans, PM also harms the environment. When deposited in the vegetation, PM inhibits photosynthetic activity and modifies the properties of the plant. As it accumulates in the soil, it hinders the exchange of nutrients and decomposition [6].

PM is classified according to the aerodynamic diameter of the particles suspended in the atmosphere, namely: PM₁₀ inhalable particles, that represents particles with 10 µm and smaller diameters, and PM_{2.5} fine inhalable particles, that represents particles with 2.5 µm and smaller diameters. PM_{2.5} particles pose the greatest risks to human health, because they can pass through the lower airways and settle in the lungs.

The main objectives of the environmental policies under development are aimed at monitoring air pollution. Based on the monitoring, it is possible to identify places and times with higher levels of emissions and, then, adopt strategies to reduce the levels of pollutants. PM concentration measurements are necessary to observe the emission limits and air quality standards.

New alternatives for environmental monitoring arise with the development and advancement of embedded systems and low-cost sensors. There are several sensors of particulate matter, pollutant that will be analyzed in this work that use different techniques for its measurement, such as by scattering or absorbing light. However, the reading of such sensors must be validated to verify the quality of the data, according to a regulated monitoring station.

Generally, measuring instruments of PM concentration are based on gravimetric, optical or microbalance methods [7]. The gravimetric-based PM monitoring equipment are called the reference method and, for instance, the Ecotech HiVol 3000 is a

reference station approved by U.S. EPA (Environmental Protection Agency). A very popular low-cost small-size PM sensor is the Shinyei PPD42NS that is based on an optical method.

As new alternatives for environmental monitoring are welcome, mainly those that can be used to monitor large areas in a distributed way. In this paper, it is described a calibration procedure of the PPD42NS sensor using Ecotech HiVol 3000 as reference. Experimental results show the performance of the PPD42NS related to its accuracy.

2. Measurement of PM_{2.5} Using Reference PM Monitoring Station HiVol 3000

The U.S. EPA defines 24-hour reference period for the average concentration of PM_{2.5} to assess the risks to the population. Ecotech HiVol 3000 is a reference station, approved by U.S. EPA, and uses the gravimetric method for PM concentration measurements. The gravimetric method consists of sampling particles by filters, as shown in Fig. 1.

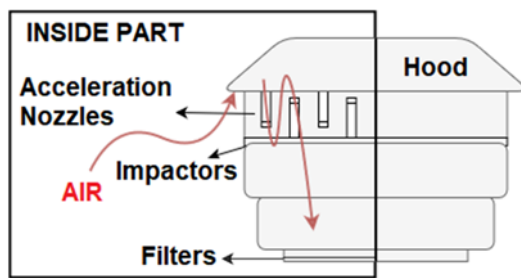


Fig. 1. Ecotech HiVol 3000.

In the first stage of the HiVol 3000 sampling process the airflow is kept constant from acceleration nozzles. Particles with a greater amount of mass gain greater momentum and are trapped in the impactor plate, which has a greased surface that helps the bigger particles to stick. The smallest particles are left and dragged to the second stage, which consists of a filter of porous material responsible for collect the particles [8].

After a predefined time period, the particles deposited on the filter are weighed to calculate the PM concentration in $\mu\text{g}/\text{m}^3$ using Eq. (1).

$$PM = \frac{(P_f - P_b) \cdot 10^6}{vol}, \quad (1)$$

where P_f and P_b , are the final and the initial mass values, respectively, and vol , the total volume of air sampled in m^3 .

This method is called manual, because even if some sampling stage is performed from an automated equipment, such as the acceleration nozzles, the filters must be changed and weighed manually daily.

3. Measurement of PM_{2.5} Using Shinyei PPD42NS Sensor

Several types of equipment are available on the market to detect PM concentration. Gravimetric instruments are large, require constant maintenance and a trained team to handle them. Sensors based on optical methods are generally small in size and can be distributed to facilitate the detection of PM concentration with spatial and temporal resolution. Most low-cost sensors use the principle of light scattering [9].

One of the main differences between the filters of the gravimetric method and the optical and microbalance instruments is that the last two allow the measurement in real time. Therefore, it is possible to analyze the concentration of particulate matter throughout the day, obtaining data on which hours are most harmful to health.

In this work, we used the Shinyei PPD42NS particulate sensor, which is based on the optical light scattering method for detecting particulate matter [10]. Light scattering method consists of two main elements: an infrared led and a phototransistor, as shown in Fig. 2.

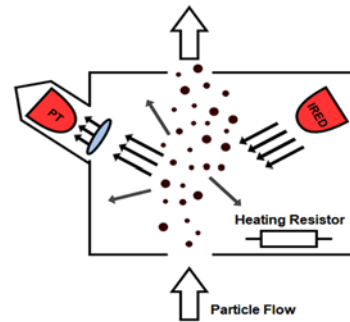


Fig. 2. Light scattering method.

The LED (IRED) emits infrared light with a wavelength equal to 940 nm and when it reaches the particles, they will refract, that is, "spread" part of the incident light, which will be captured by the phototransistor (PT), generating a certain level of voltage directly related to the concentration value of particles [11].

The air will enter the lower cavity carrying the particles and must flow to the upper cavity, therefore, the sensor must always be used vertically. To direct the airflow, the PPD42NS has a heating resistor which is responsible for the air convection, because the hot air is less dense and will go to the top, so the particles flow through the area detection.

4. Calibration Process and Evaluation Process of the PPD42NS Sensor

For the evaluation of sensors that are not reference from the reference equipment, the "collocation" method is used, where both must be positioned close

to each other, in a predetermined period, so that the measurements can be compared [12].

In this experiment, for the evaluation of PPD42NS sensor related to the reference equipment we used the collocation method. The sensor was positioned approximately 1 meter away from the HiVol 3000 equipment, as can be seen in Fig. 3.

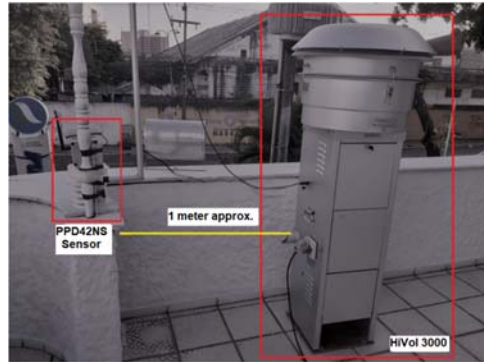


Fig. 3. Experimental setup.

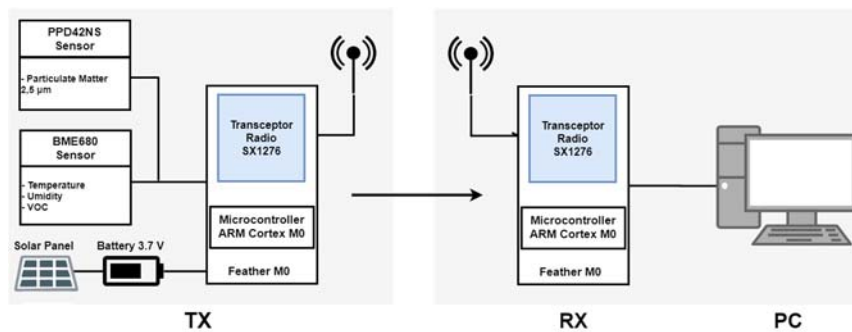


Fig. 4. Schematic of the experiment for measuring particulate material.

PM_{2.5} [$\mu\text{g}/\text{m}^3$] data were collected over 5 days of experiment, as shown in Fig. 5.

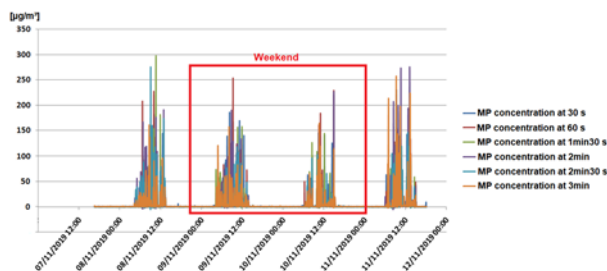


Fig. 5. Concentration data of PM_{2.5} over the days of the experiment.

After such evaluation period, it is possible to check the performance of the proposed sensor and the quality of the data. When analyzing the graph, it is possible to see that the concentration data on the weekend, highlighted in Fig. 5, have lower values than on other days. This drop in concentration values is even more notable on Sunday. This is due to the fact that the movement of people and cars downtown is reduced when compared to other days of the week.

As HiVol3000 provides only PM average concentration data on daily base, the same period of

The sensor node developed in this work consists of a BME680 sensor, which provides data on temperature, humidity, atmospheric pressure, VOC gases and altitude, and the PPD42NS, previously presented. The sensor is powered by battery and solar panel, it is portable and wireless. So it can be positioned in any location to assess air quality.

The schematic of the proposed system can be seen in Fig. 4. The data from the environmental sensors are transmitted via LoRa modulation to a receiver connected to a computer with internet access, where the data is stored. The treatment and calculation of PM concentration values were performed offline in the period considered after finishing the tests.

The data are sampled every 30 seconds, but are transmitted periodically every 3 minutes via LoRa modulation. The radio is put into sleep mode until it reaches 3 minutes, in order to minimize the current consumption in the transmission mode and increase the battery life.

time was used to calculate the PM average concentration using the PPD42NS sensor that is sampled every 30 seconds.

Daily averages of the PPD42NS sensor were calculated from the arithmetic mean, adding up the concentration values according to the number of packets received during 24 hours. The four daily average concentration values refer to the period in which the HiVol 3000 equipment was in operation in a building in downtown area. The results of the average daily PM concentration for the reference equipment and for the PPD42NS sensor are in Table 1.

Table 1. Values of PM Concentration in Reference and Low Cost Equipment.

Day	HiVol 3000 [$\mu\text{g}/\text{m}^3$]	PPD42NS Sensor [$\mu\text{g}/\text{m}^3$]
07 to 08 nov2020	10.345	7.558
08 to 09 nov2020	10.681	7.347
09 to 10 nov2020	7.838	5.057
10 to 11 nov2020	10.969	8.112
Average	9.958	6.483

To obtain the calibration curve from the points of the reference equipment on the ordinate axis and the mean values of the low cost sensor on the abscissa axis,

the function we used the Octave (Matlab similar program) function “polyfit”, which returns the coefficients of the polynomial that best represents the entry points provided by the user.

The resulting first-degree polynomial is expressed by Eq. (2), and the curve can be seen in Fig. 6.

$$y = 1,0539x + 2,5608 \quad (2)$$

The fit quality of a curve can be determined by the correlation coefficient R^2 . The correlation coefficient varies from -1 to 1, when equal to 1 indicates positive perfect linear correlation that means if one variable increases linearly, the other also increases; when equal to -1 indicates negative perfect linear correlation that

means if one variable increases linearly, the other decreases. When equal to 0, it means that the two variables have no association at all [13].

From Eq. (2), the R^2 value obtained was 0.9837, indicating a high linear correlation between the variables and the mean square error obtained was 0.2239.

With Eq. (2), the readings of the PPD42NS sensor will be the input values x and the output y will be the adjusted value according to the calibration implemented with the HiVol 3000 equipment as a reference, being able to provide reliable values, assisting in real time monitoring of particulate matter concentration.

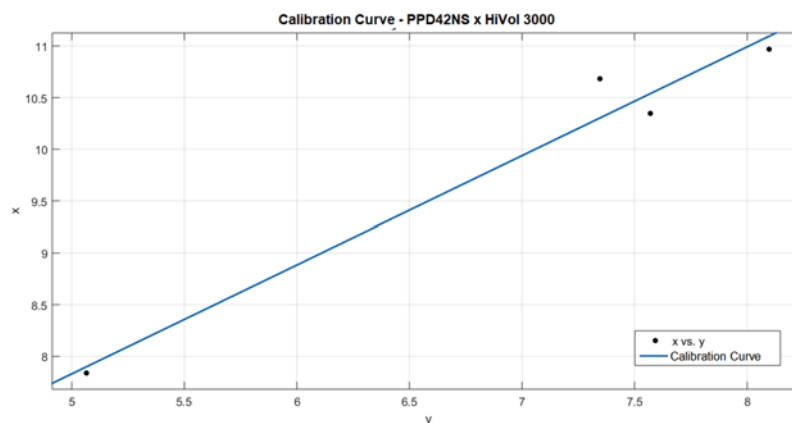


Fig. 6. Calibration curve of the PPD42NS sensor.

5. Conclusions

In this work, we evaluate the performance of the PPD42NS sensor having the reference PM monitoring equipment Ecotech HiVol 3000 as reference. To analyze the quality of the data, a calibration was performed and the linear correlation was greater than 0.9 indicating a high linear correlation between them. We concluded that PPD42NS offers small-size, low cost and dispensability of daily filter changes when compared to the gravimetric method, which suggests that the PPD42NS sensor can assist in the sampling of PM_{2.5} for monitoring the evolution of air quality.

Acknowledgements

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Automated Integration of Electronics in Smart Textiles Using Ultrasonic Soldering

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Summary: There is no suitable connection method to produce smart textile automatically at the moment. Ultrasonic soldering could be a good solution since it works with flux-free solder, which avoids embrittlement of the textile integrated wires. This article describes the detailed process of robot-assisted ultrasonic soldering of smart textiles to printed circuit boards (PCB). The aim is to understand the influencing factors affecting the connection and to determine the corresponding solder parameters. Various test methods are used to evaluate the samples, such as direct optical observation of the microstructure, a peeling test and contact resistance measurement.

Keywords: Smart textiles, Wearables, Ultrasonic, Soldering, Bonding, Integration of electronics.

1. Introduction

According to current studies, the market volume for intelligent textiles will grow significantly in the coming years. In addition to the current medical and military applications, the usage will expand into the fields of fashion and sports [1]. The challenge is to ensure the usual textile properties for the customer, such as textile feeling, washability, wear comfort and drapability [2]. However, we currently do not see a real market breakthrough. One of the reasons for the low market penetration of electronic textiles is the high proportion of manual activities during production [3]. Automated methods for contacting PCBs with insulated wires have not yet been sufficiently investigated in the literature. In this work, we have focused on textile insulated wires. These wires are located inside the textile enabling to connect components from both sides. The manufacturing process typically consists of three steps. First, the wires are automatically uninsulated with a laser. Then, the main contacting process starts, which is investigated in this article. To protect the electronics from environmental influences, the components are overmolded in an injection molding process [4].

2. Materials and Methods

2.1. Ultrasonic Soldering

During ultrasonic soldering, the heat input occurs primarily through the temperature-controlled solder tip. It melts the solder and the additional ultrasonic movement of the tip removes boundary layers on the surface, which ensures solderability without flux. The ultrasonic excitation leads to cavities in the liquid solder, which remove impurities (impurity layer) from the surface similar to the principle of ultrasonic

cleaning. Special solder alloy without flux must be used, since soldering materials with flux inhibit the transmission of ultrasound.

2.2. Experimental Setup

A warp-knitted textile tape made out of polyester with four integrated PTFE insulated AWG 32 strands (19 wires) was used. The tape consists of flexible (stretchable) areas and inflexible contact points. The insulation and textile tape are removed locally at the contact points from both sides with a CO₂-laser. A USS-4200 from MBR Electronics with different widths of tips (1, 3 and 5 mm) was used. The temperature was continuously adjusted between 0 °C and 500 °C, the US power was varied between 4 W and 12 W at 60 kHz. A four-axis industrial robot arm (Dobot M1) was programmed to automate and ensure reproducibility in the production of the test samples. In addition, a special holder for the textile tape and the PCB was manufactured. The ultrasonic soldering iron and the solder feed are carried by the robot. We used three different solder alloys: Sn95.5Ag3.8Cu0.7, Sn99.7Cu0.3 and Sn97Ag3 without flux. These alloys are common in the electronics industry, in which flux-free solders are used.

2.3. Test Procedures

Micrographs of the samples were first prepared and optically examined under a microscope. The measurement of the contact resistances is an indicator of the quality of the contact itself. For a reliable measurement and to eliminate line resistances, a four-wire measuring instrument was used. This is due to the reason that contact resistances of metal connections are in a range milliohms [5].

The peeling test was used to determine the mechanical strength of the joint for further processing. The force was applied perpendicular to the PCB. Therefore a load cell with a maximum force of 1000 N was used. The travel speed is 100 mm / min [6].

4. Results and Discussion

Micrographs of the cross-section were made of all solder materials (Sn95.5Ag3.8Cu0.7 (Fig. 2a, d); Sn99.7Cu0.3 (Fig. 2b, e); Sn97Ag3 (Fig. 2c, f)). Sn99.7Cu0.3 (Fig. 2e) shows significantly less pores compared to the other samples. Sn95.5Ag3.8Cu0.7 has a needle-shaped micro-structure, which indicates that the metal alloy is brittle. The microstructures of Sn99.7Cu0.3 and Sn97Ag3 show a much more homogeneous picture. This is also reflected in the results of the peeling test.

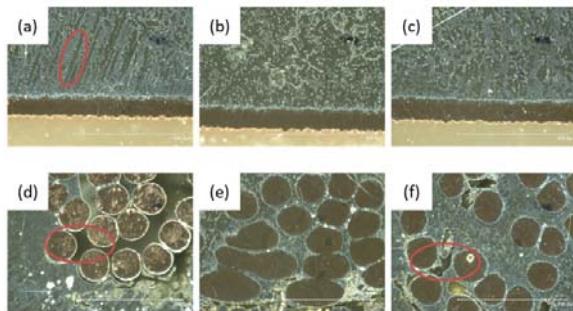


Fig. 2. Microsection of various solder materials, (a, d) Sn95.5Ag3.8Cu0.7, (b, e) Sn99.7Cu0.3, (c, f) Sn97Ag3.

In the peeling test, the textile tape is pulled off at a 90° angle from the PCB. The contacting points usually fail one after the other. There are mainly two types of failure. One is the break outside the joining zone, the other one is an adhesive breaking at the surface. In addition, other influencing variables and their effects were investigated. The used solder materials have effects on the contact resistance and the mechanical strength of the connection. The contact resistances ranged from 6 to 9 mΩ. Solder joints made of Sn97Ag3 had the lowest electrical resistance and low strength, Sn99.7Cu0.3 had the highest strength but also the highest electrical resistance. The amount of solder added has only a minor influence on the electrical and mechanical properties of the connection.

The temperature at the soldering tip far above the liquids temperature leads to lower strength. A larger diameter of the soldering tip increases the heat capacity and thus facilitates melting of the solder. Furthermore, a larger soldering tip can lead to an acoustic coupling of the soldering tip with the solder. An increased ultrasonic power ensures a more homogeneous micrograph at the contact point. However, measurements were made at higher ultrasonic power levels and longer processing times, where the copper

pad of the board partially diffused into the melt. If the power is too low, there is no permanent electrical contact, as the impurity layers cannot be removed. During the measurement of the contact resistances only a very small influence can be determined. The greatest influence can be seen in the silver content of the solder. The higher the silver content, the lower the resistance. The strength increases while reducing the temperature. We found the following optimal soldering parameters: Sn99.7Cu0.3 at 350 °C and an ultrasonic working time of 0.5 s with a power of 10 W.

5. Conclusion

This article focuses on the development of a contacting method by means of a partially automated ultrasonic soldering system to integrate PCB components on smart textiles.

Subsequently, the mechanical and electrical properties of the soldered connection were analyzed as a function of the soldering parameters. The investigations of the peeling test showed that connections with solder made of Sn99.7Cu0.3 yields higher strengths than connections with Sn95.5Ag3.8Cu0.7 or Sn97Ag3. The reason is the better homogeneous microstructure with only few defects. The solder temperature is also a very important influencing factor. The strength increases while reducing the temperature. In addition, a too high or too low working temperature damages the contact. The ultrasonic power and the ultrasonic process time have a great influence on the formation of the microstructure. To be precise, the shorter the ultrasonic treatment time during the soldering process, the better the properties of the solder connection.

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(038)

Insights to Electrochemical Polymerization Conditions Effect to Electrochromic Properties of Polypyrrole

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Summary: Conducting polymer polypyrrole (Ppy) electrochemically was deposited on the indium tin oxide coated glass (glass/ITO) electrode. The intended field of future application of the obtained glass/ITO/Ppy system was to use it as the electrochromic sensor. For that purpose, the adhesion of polymer on the glass/ITO electrode was tuned by modification of the electrode surface with triethoxymethylsilane (TEMS). Absorbance changes of glass/ITO/Ppy were followed in a Britton – Robinson buffer (BRB) at different pH and potential values.

Keywords: Conducting polymers, Polypyrrole, Electrochromic polymers, ITO, Electrodeposition.

1. Introduction

Conducting polymer polypyrrole is a very common object in the field of electrochemical studies. Ppy is easily synthesizable by chemical [1] or electrochemical methods [2]. The applications of Ppy include the development of molecularly imprinted polymers [3], glucose sensors [4], using it as the electrochromic material [5, 6], and many others.

The aim of this study was to draw some insights into electrochemical polymerization conditions effect to electrochromic properties of polypyrrole.

Ag/AgCl_(3M KCl) was used as a reference electrode and a platinum wire was used as a counter electrode. The electrochemical deposition was performed at room temperature by 10 potential cycles in the range from -0.2 V to +1.0 V vs Ag/AgCl_(3M KCl) at the sweep rate of 50 mV/s and step potential of 2.44 mV.

BRB was made of 0.01M of boric acid, 0.01 M of acetic acid, 0.01 M of phosphoric acid, and 0.1M of LiClO₄. The pH of BRB was adjusted with 0.1 M of NaOH. The optical measurements were performed with a spectrometer USB4000-FL equipped by 'SpectraSuite' software from Ocean Optics (Largo, USA).

2. Experimental

All the chemicals used experiments were of analytical grade and were used as obtained, except pyrrole, which was purchased from Sigma-Aldrich (Steinheim, Germany) and distilled before electrochemical deposition.

Glass/ITO electrode (15–25 Ω/cm²) was purchased from Sigma-Aldrich (Steinheim, Germany). Before the deposition of Ppy electrode surface was cleaned in the following order: 27 % NH₄OH and 30 % H₂O₂ (3:1, 50 °C) for 5 min, then cleaned in ultrasonic bath subsequently in water, acetone, and water for 15 min in each liquid.

The glass/ITO surface with silane was modified in a 4 % (v/v) solution of TEMS in acetone overnight. After that glass/ITO was washed with acetone and ultra-pure water.

Ppy was electrochemically deposited from 10 mM of pyrrole and 0.1 M of LiClO₄ solution using a computer-controlled potentiostat/galvanostat PGSTAT 128N with Nova 1.10 software from Eco-Chemie (Utrecht, The Netherlands). The

3. Results and Discussions

The first part of the study was attributed to evaluation of TEMS influence on the electrochemical deposition process. It was observed that adhesion of Ppy film on bare glass/ITO electrode surface was insufficient during the measurements and degraded by vigorous stirring of solution. Electrochemical deposition of pyrrole on the glass/ITO_(bare) (Fig. 1A) and on the glass/ITO(TEMS) (Fig. 1B) electrodes demonstrated some differences of polymerization process.

Changes of absorbance of glass/ITO(TEMS)/Ppy followed in a Britton – Robinson buffer (BRB) at several different values of pH and potential (Fig. 2). Absorption changes were followed at 750 nm wavelength.

The potential of glass/ITO_(TEMS)/Ppy electrode was changed from 0 V to +0.6 V vs Ag/AgCl_(3M KCl) electrode. It was observed that only in the acidic medium the absorbance exhibited significant absorbance differences during the alteration of the

electrode potential. When the pH of BRB solution was increased, the changes of Ppy layer absorbance at 750 nm became negligible.

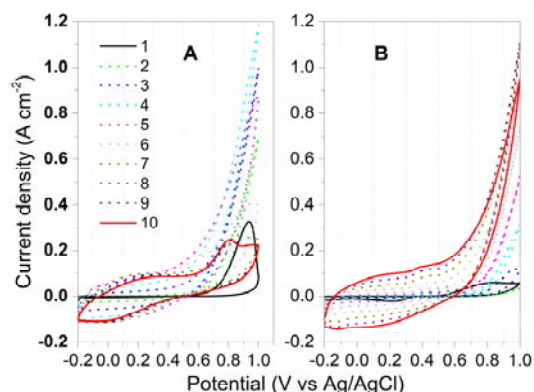


Fig. 1. Electrochemical deposition of pyrrole: (A) on the glass/ITO_(bare) and (B) on the glass/ITO(TEMS) electrodes in room temperature by 10 potential cycles in the range from -0.2 V to +1.0 V vs Ag/AgCl_(3M KCl) at the sweep rate of 50 mV/s and step potential of 2.44 mV.

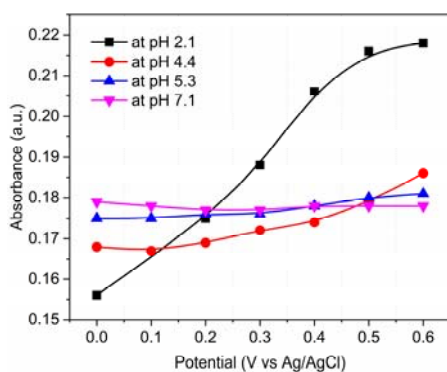


Fig. 2. Absorbance changes of glass/ITO(TEMS)/Ppy followed in a Britton – Robinson buffer (BRB) at different pH and potential values. Absorption changes were registered at 750 nm wavelength.

4. Conclusions

An electrochemical polymer deposition method was applied for modification of indium tin oxide (ITO) coated glass-based light-transparent electrode by polypyrrole. Adhesion of Ppy on the glass/ITO

electrode improved with modification by triethoxymethylsilane. Only in the acidic medium were observed significant differences of absorbance of glass/ITO(TEMS)/Ppy electrode when the potential was changed from 0 V to +0.6 V vs Ag/AgCl_(3M KCl).

Acknowledgements

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(042)

Development of Quartz Crystal Microbalance Based Sensor for Real-time Ozone Monitoring

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Summary: This paper presents the development of quartz crystal microbalance based sensor for the real-time monitoring of ozone (O₃) in workplaces in the range of 10-100 ppbv (part per billion per volume). Quartz crystal microbalances are interesting because of their capability to detect the adsorption of gaseous compounds on the crystal coating. Here, we focused on the polybutadiene performance as coating material in terms of stability, sensibility and reproducibility under real conditions. Chemical sensors were exposed to low concentration of ozone in a dedicated controlled test atmosphere device in order to determine the influence of several parameters: temperature, humidity, concentration and sampling flow rate. Results show that our detection device can detect 7 ppb in few minutes and does not undergo decrease of sensibility under moisture. In addition, the presence of nitrogen dioxide does not interfere with ozone detection.

Keywords: Ozone sensor, Microbalance, Polybutadiene.

1. Introduction

Ozone (O₃) is a powerful oxidizing gas and a strong disinfectant used in different processes, including water treatment, gas purification, textile bleaching, and food industries. Additionally, O₃ is emitted in workplaces by laser printers and photocopiers and from the processes of ultraviolet-radiation and electric arc welding [1]. In France, the occupational threshold limit value (TLV) for O₃ exposure is 100 ppbv and 200 ppbv for an exposure time of 8 hours and 15 minutes respectively. O₃ concentration TLV exceedance can cause lung damage and respiratory diseases such as asthma [2]. Occupational exposure assessment should be performed by individual sampler. Therefore, it is necessary to develop a personal miniaturized sensor that provides real-time measurements O₃ concentrations in workplaces.

Conventional chemical analysis techniques, based on individual sampling on solid adsorbents or filters, are not convenient for ozone real-time personal exposure assessment. They require the analysis in chemical laboratories which does not alert in real time in case of high exposure. In addition, the results obtained are the average concentrations over an extended period of time. Various instruments were developed for measurements of ambient O₃ based on strong light absorption at 254 nm and chemiluminescence techniques [3]. These instruments exhibit high sensitivity, selectivity, and reliability. However, the size, cost, power consumption and regular calibration needs, reduce their use for real-time measurements when portability of the equipment is required. Colorimetric O₃ sensors were also developed on the basis of dye fading colorimetry [4] such as indigo carmine [5]. However, these sensors show poor reproducibility, low sensitivity and selectivity. Takada et al. [6] show that semiconductors are sensitive to ozone but they also highlighted their strong

dependence to temperature variation. Therefore, the development of efficient, portable and low-cost ozone detection devices is of high interest concerning workers security. The quartz-crystal microbalances (QCM) are one of the most promising devices. Black et al. [7] previously shown that polybutadiene (PBU) coated quartz were performing as ozone sensing layer. Additional investigations are described in this work: PBU performances will be discussed, in particular at low O₃ concentrations, under humid conditions and in presence of other gases. Ageing of the PBU coated quartz will also be assessed.

2. Experimental Section

Quartz crystals were coated by spin-coating. A solution of PBU (Sigma-Aldrich pur.) at 9 mg/mL in cyclohexane is prepared. 7.5 µL of the solution are deposited on both sides of the quartz.

The controlled test atmosphere device used to perform O₃ exposure trials was fully described by Delcourt et al. [8] (Fig. 1). O₃ is generated by passing a flow of dry air through an O₃ generator (Thermo Scientific Model 49i UV Photometric O₃ Analyzer); the O₃ monitor is also used for assessment of O₃ concentrations in the exposure chamber. A portable humidity and temperature probe is placed close to the detection device. A variable flow pump (50-500 mL/min) is used to ensure continuous flow of O₃ inside the measuring cell. The quartz frequency variations are measured by a QCM (Novaetech S.r.l. Italy) integrated with a homemade microfluidic circuit (Fig. 2).

Air is sampled at the top of the fluidic cell (a), through a carbon fiber to filter out any particles. The pump is connected to the metal tube (b). Thus, the two faces of the quartz placed into the cell are properly exposed to the sampled air flow.

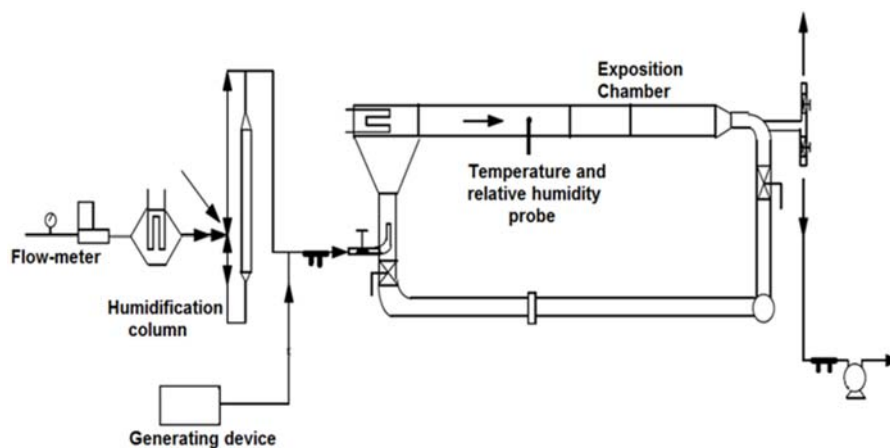


Fig. 1. Schematic diagram of the generation bench.

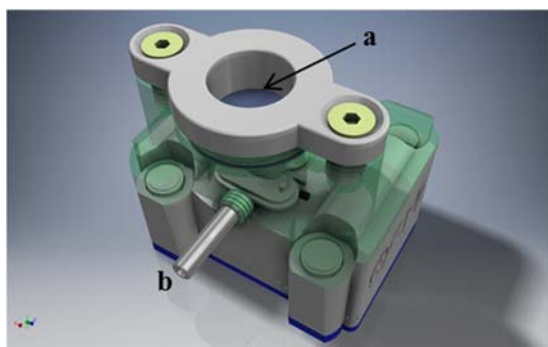


Fig. 2. 3D view of the QCM integrated with the homemade microfluidic cell.

3. Results and Discussion

Sensor sensitivity is tested in the concentration range between 10-100 ppbv. The sensor response as a function of ozone level was evaluated during 15 minutes, firstly under dry air at 200 mL/min, then at 400 mL/min, and finally under 50 % of relative

humidity (RH) at 400 mL/min. Each exposure test was duplicated.

Fig. 3 shows that sensor sensitivity increases as flow sampling increases.

Good repeatability is obtained at 10, 35 and 80 ppbv. In addition, the frequency variation measured at 50 % of RH is similar to that obtained under dry air.

Figs. 4 and 5 report QCM-1 and QCM-2 frequency shift per minute, as a function of the ozone concentration.

These results show that sensors response is linear in the range 10-100 ppbv. In addition, Fig. 4 clearly confirms that the sensor sensibility is much better at 400 mL/min than at 200 mL/min. Furthermore, results reported on these two figures show that humidity does not reduce the performance of the sensors at 10, 30 and 70 ppbv of ozone. Performances of both QCM-1 and QCM-2 are evaluated several days after the quartz coating. Fig. 4 shows that QCM-1 response after 26 days is still comparable to results obtained after 7 days. Tests performed with QCM-2 (Fig. 5) show that its performances are constant until 50 days after the coating.

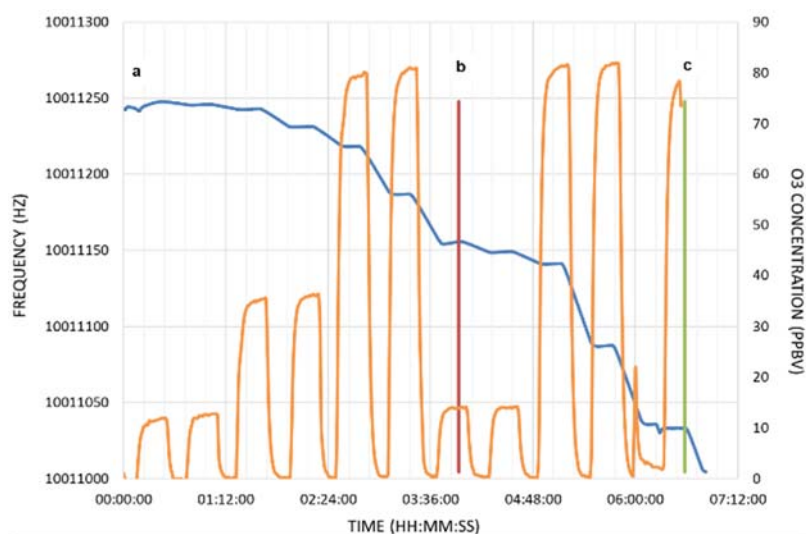


Fig. 3. QCM-1 frequency variation (—) and ozone concentration (—): (a) under dry air at 200 mL.min⁻¹ (b) under dry air at 400 mL.min⁻¹ (c) under 50 % of RH at 400 mL/min.

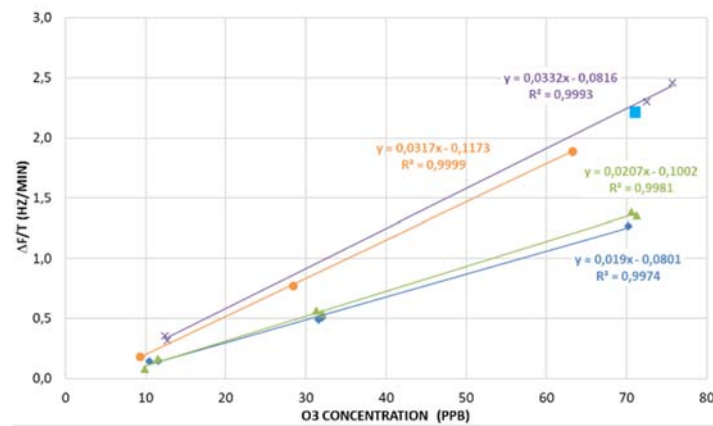


Fig. 4. QCM-1 response as a function of ozone concentration: (♦) day 5 exposure under dry air at 200 mL/min (▲) day 7 exposure under dry air at 200 mL/min (×) day 7 exposure under dry air at 400 mL/min (■) day 7 exposure at 400 mL/min and 50 % of RH (●) day 26 exposure under dry air at 400 mL/min.

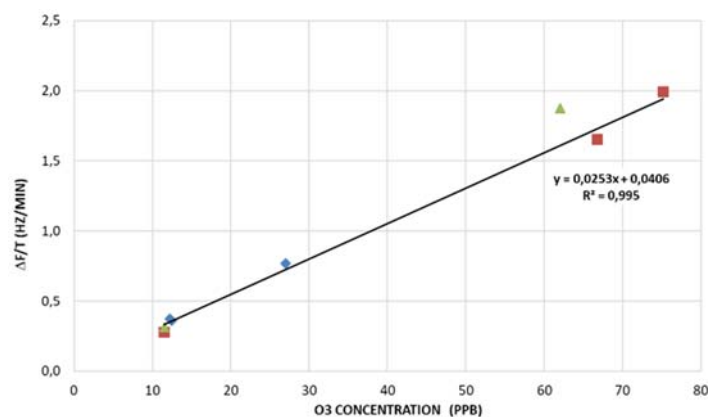


Fig. 5. QCM-2 response as a function of ozone concentration at 400 mL/min: (■) day 1 exposure under dry air (♦) day 1 exposure under 50 % of RH (▲) day 50 exposure under dry air.

The sensor selectivity was considered in presence of toluene and nitrogen dioxide (NO₂), another oxidizing gas. During those tests, the sensor was exposed to O₃, interferent, and both gases. Figs. 6 and 7 respectively report the influence of toluene and NO₂ on QCM-3 and QCM-4 responses.

No influence of both interferent is observed on QCM responses. Moreover, the frequency shift is the same under O₃ either with or without interferents.

Consequently, the nitrogen dioxide does not react with the sensing layer, which is a very positive result, as most of commercial ozone sensors are sensitive to NO₂.

Additionally, Fig. 6 shows that QCM-3 reacts under 5 ppb of ozone, in presence of 50 % RH. These results are very promising as no other sensor has shown such low sensitivity in humid air, combined with very good selectivity towards NO₂.

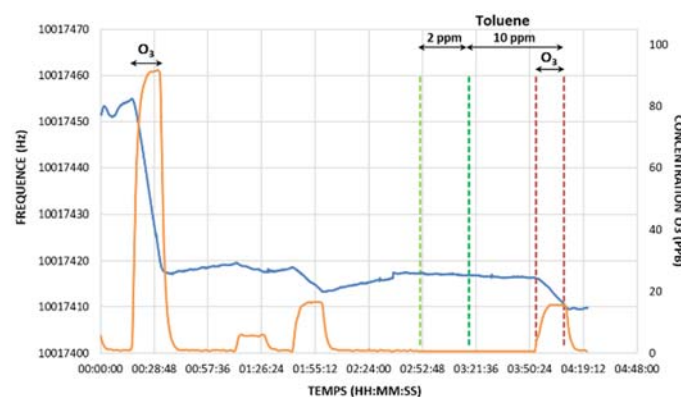


Fig. 6. QCM-3 response as a function of ozone concentration at 400 mL/min, under 50 % of RH, in presence of toluene at 2 and 10 ppm.

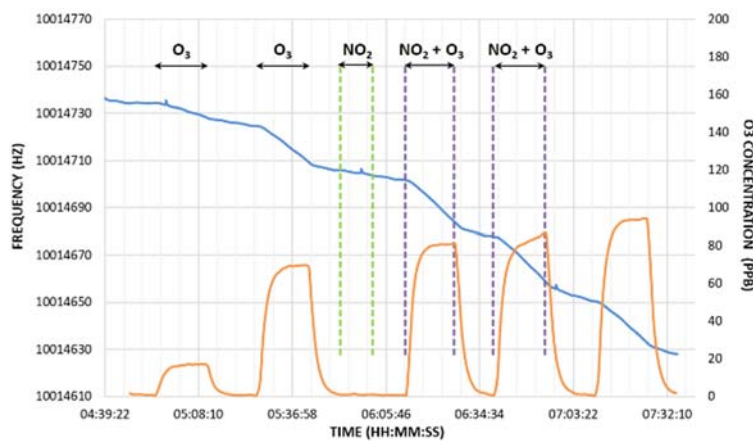


Fig. 7. QCM-3 response as a function of ozone concentration at 400 mL/min, under 50 % of RH, in presence of NO₂ at 500 ppb.

4. Conclusions

We developed a small and low cost real-time ozone sensor, suitable for individual sampling, able to detect ozone below 10 ppbv in a few minutes. Its sensibility is not affected by moisture and the presence of toluene or nitrogen dioxide and obtained results show that quartz can be used until 50 days after the coating.

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(043)

High Energy Photo-neutron Interrogation of Uranium with Tensioned Metastable Fluid Detectors

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Summary: Active special nuclear material (SNM) photoneutron interrogation research with Tensioned Metastable Fluid Detector (TMFD) sensor technology demonstrated the TMFD technology to readily and conclusively detect the presence of SNM while remaining insensitive to the intense 36,000 R h⁻¹ 9 MeV endpoint X-ray background. Results for the centrifugally and acoustically tensioned metastable fluid detectors show how SNM (uranium) fission neutron energy detection thresholds can be used to detect SNMs under intense radiation conditions wherein conventional state-of-art detectors get saturated.

Keywords: Active photon interrogation, Special nuclear material, Linear accelerator, Tension metastable fluid detectors.

1. Introduction

Active photon or neutron interrogation is widely considered the primary method for monitoring the presence of special nuclear material (SNM) for combating nuclear terrorism – a grand challenge for the 21st century [1, 2]. Active interrogation induces nuclear reactions in deeply buried targets, used then, to obtain information about the target, via an interrogating source. Sensors for the interdiction of SNMs have been developed. However, significant technology gaps remain for enabling a real-time, high efficiency, gamma blind neutron radiation monitor [2].

Purdue University's Metastable Fluids and Advanced Research Laboratory (MFARL) have developed and demonstrated gamma blind, real-time, high efficiency (80-90 %+) neutron-alpha-fission sensors for the rapid detection of SNMs and also Rn-progeny [3-8]. This article will discuss results of photoneutron active interrogation experiments, conducted at the University of Michigan (UM), that rapidly and conclusively determine the presence of SNM using TMFDs in a 10⁴ R h⁻¹ 9 MeV photon field.

1.1. TMFD Overview

TMFD sensors use centrifugal or acoustic force to create metastable, sub-vacuum, negative pressure (P_{neg}) states in fluids [4]. Intermolecular bonds in these states are stretched and broken via external energy deposition leading to cavitation and vapor formation. This process is dependent on the tailored negative pressure state and allows for select discrimination of neutron energies. Additionally, the insensitivity to beta and gamma radiation is inherent due to the low linear energy transfer for each type of radiation. Several experiments have been performed to verify the blindness to direct photon and beta detection [4]. This gamma/beta blindness is useful for active interrogation work where photon sources are present.

In Acoustic TMFDs (ATMFDs) P_{neg} states are induced via ultrasonic acoustic perturbations. In ATMFDs, a piezo-electric material is affixed to the base which converts electromagnetic energy to acoustic wave patterns, which, in resonance mode inducing large spatio-temporal varying P_{neg} states. When in a compressive state, the system remains blind to ionizing radiation; however, when in tensile, i.e., P_{neg} states, neutron signals can be monitored readily, while also offering directional information for incoming neutron radiation [3-4].

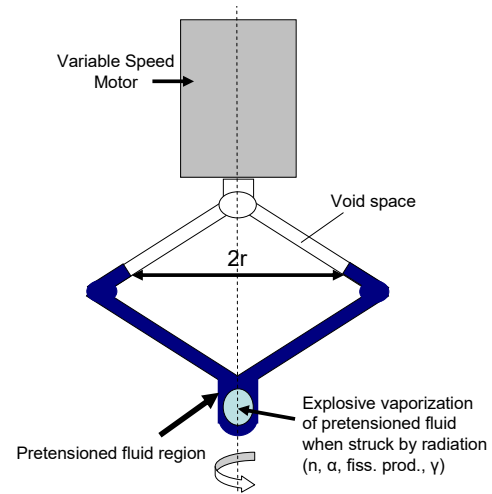


Fig. 1. Schematic of CTMFD sensor apparatus.

In the relatively simpler Centrifugal TMFDs (CTMFDs), shown schematically in Fig. 1, the tailoring of negative pressure can be expressed as [3],

$$P_{neg} = 2\pi^2 \rho r^2 f^2 - P_{amb}, \quad (1)$$

where, P_{neg} is the negative pressure at the centerline of the bulb, ρ is the density of the fluid, r is the distance from the centerline to the average of the fluid's inner

and outer meniscus, f is the rotational frequency, and P_{amb} is the ambient pressure.

Selective tailoring of the negative (sub-vacuum) pressure (P_{neg}) states determines neutron sensitivity and therefore can be used as a neutron energy threshold detection mechanism. The higher the P_{neg} state, the lower is the elastically scattered detectable energy of the neutron. The neutron detection threshold can then be set below the neutron background energies, enabling insensitivity to neutron background from the interrogating source. This neutron background insensitivity combined with inherent photon blindness, implies that any measured neutron signals below these thresholds must originate from photon induced neutron-fission of the targeted SNM materials.

2. Methodology

Experimentation was performed at the University of Michigan to demonstrate for the ability to rapidly detect a depleted uranium (DU) target while distinguishing it from both a photo-neutron background and a 600 R/min 9 MeV endpoint photon source. A 3.2 kg DU (SNM surrogate) target was placed at a 10m distance from the collimator, where the beam width is ~ 0.3 m in diameter. At this location, the CTMFD and ATMFD were positioned ~ 1.85 m off axis at 90 degrees. At 5m further down the beam line, a beam stop consisting of 2.5cm of polyethylene followed by 10cm of lead, and then finally 1m of concrete is used to terminate the photon beam. One 20 min. trial was run with the 3.2 kg DU target, and two trials (20 min. & 5min.) without DU.

3. Results

Fig. 2 shows the results of tests with the ATMFD which show that for acoustic drive powers of < 7.75 W, the ATMFD remained blind (i.e., zero counts) to the intense (36,000 R/h) 9MeV X-Ray photon background. However, only when the 3.2 kg DU is included, do we convincingly detect it via active interrogation sensing of the fast energy neutrons emanating from photo-fission (from within the target).

Fig. 3 depicts similar, simultaneous results with the CTMFD for the trials with and without the DU target. Interestingly, while the beam is operating with no target, a sudden decrease in detection rate occurs at $P_{neg} \leq \sim 3$ bar. For 5 minutes at 2.75 bar there were zero detections while the beam was operating with no target. Therefore, it was determined that the P_{neg} threshold for detection of the photo-neutrons being generated was 2.75 bar. Comparing this P_{neg} of 2.75 bar with our previously (yet unpublished) data [8] indicates that the likely source of this background is due to $^2\text{H}(\gamma, n)$ reactions occurring in the hydrogenous beam stop generating ~ 3.4 MeV photo-neutrons.

With the DU target in place, for $P_{neg} \geq 2.75$ bar, the CTMFD count rate jumps from zero, to around 2+ CPM and beyond, demonstrating the successful

detection by TMFDs (at P_{neg} states < 2.75 bar) of the photon induced fission spectrum neutrons (> 3.4 MeV) emanating from the DU target - while successfully remaining blind to the intense background.

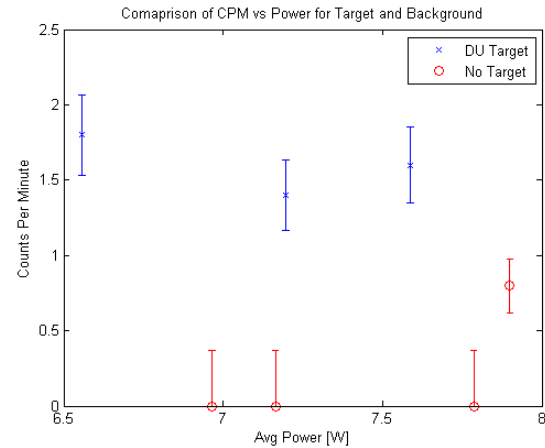


Fig. 2. Results of 9 MeV (endpoint) X-Ray Interrogation with and w/o 3.5 kg DU with ATMFD sensor technology. Measurement times for each drive power are 5 minutes. Error bars are 1σ .

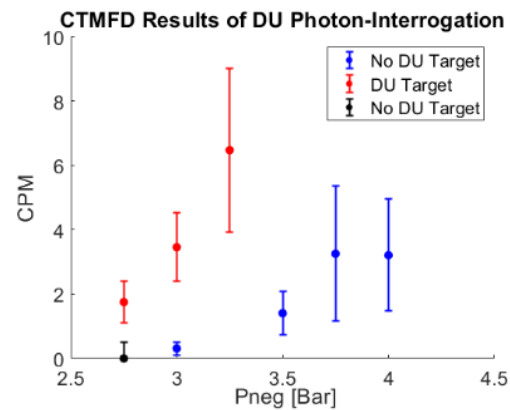


Fig. 3. Results of 9 MeV (endpoint) X-Ray Interrogation with and w/o 3.5 kg DU with CTMFD sensor technology.

4. Conclusions

An active interrogation experiment using a 36,000 R/h, 9 MV LINAC at the University of Michigan was performed with DU (as a surrogate to HEU) using TMFD sensor technology. The results show that the TMFDs can readily detect the presence of the U target, while discriminating neutrons from background photo-neutron production reactions. Overall, the experiment shows the capability for TMFD sensors to perform in extreme (10^4 R/h) intensity gamma/neutron fields, while remaining sensitive to fission neutrons.

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Broadband Multimode Single-channel Optical Rotary Connector

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Summary: Optical rotary connectors (ORC) belong to a new generation of newly created optical devices capable of providing continuous optical communication between objects mounted on a common axis of rotation. Due to the wide bandwidth, contactlessness and insensitivity to electromagnetic interference, ORC allow you to reliably transmit various digital and analog signals from rotating objects to stationary ones on broadband fiber-optic communication lines, as well as to take contactless information in optical form from fiber-optic sensors of physical quantities mounted on rotating objects. This brings the telemetry of rotating objects to a qualitatively new much higher level. In this work, we developed a broadband multimode single-channel ORC, built on the basis of ball optical lenses made of optical sapphire. The comprehensive theoretical and experimental studies of this ORC have been conducted. Its design is characterized by high manufacturability, ease of assembly and many times less requirements for the accuracy of size and shape of individual structural components, assemblies and assemblage of such ORC compared to those built on the basis of gradient rod lenses.

Keywords: Fiber optic rotary joint, Fiber optic collimator, Graded-index (GRIN) lenses, Ball lenses.

1. Introduction

In modern control and measurement devices and communication technology between the moving (rotor) and body (stator) parts of rotating objects, there is a problem of transmission of broadband signals from one part of the object to another. The traditional design and circuit solution of this problem due to the use of current-carrying contact rings is limited by a number of significant disadvantages. They include: the insufficient (for transmission of broadband signals) frequency band of electrical signal transmission channels; high level of electrical noise of contacts; insufficient protection against electromagnetic interference; low reliability; short service life, etc.

In addition, with the introduction of fiber-optic sensors of physical quantities and fiber-optic communication lines on rotating objects, additional difficulties arose due to the need to couple physically incompatible electrical interfaces of slip rings with optical interfaces of fiber lines.

In recent years, to solve these problems, fundamentally new devices have been created – the so-called optical rotary connectors (ORC). These devices provide contactless transmission of digital and analog signals from rotating objects in optical form. They adequately meet the task and are compatible with fiber lines. Compared to contact connectors, the contactless ORC have tens of thousands of times wider bandwidth, can operate under the influence of powerful electromagnetic interference (being insensitive to electromagnetic fields), are much more reliable, have a long service life, small size and weight.

Today, dozens of different designs of multi-channel and single-channel facilities, operating

on different physical principles, have been created and commercially produced [1-2]. Optical collimator devices based on graded-index rod lenses (GRIN lenses) and C-lenses are widely used in many of these designs.

The main advantages of GRIN lenses are low chromatic aberration, simple cylindrical shape of lenses with flat end surfaces, convenience for installation in optical devices (easy to assemble due to the plane surfaces). At the same time, the use of GRIN lenses in ORC requires extremely high precision from the structural elements of ORC and its assembly.

For example, the angular beats of the optical axes of the lenses relative to the axis of rotation of 0.1° lead to rotational oscillations of optical losses in the ORC at the level of 1.6 dB [3].

The following work is devoted to the development of broadband multimode single-channel ORC, built on the basis of ball optical lenses. This allowed significantly reduce (compared to devices that use GRIN lenses) the requirements for the accuracy of basic elements of the ORC design and significantly simplify its assembly technology.

2. The Design of ORC, Its Characteristics and Theoretical Analysis

The principle of operation of the device is explained by a simplified optical scheme of the ORC in Fig. 1.

Optical signals are transmitted from one optical fiber to another by means of optical lenses through a small air space between the lenses. The lenses with optical fiber segments are mounted on a common axis

of rotation at a certain distance L from each other. The ends of the transmitting and receiving optical fibers are installed at the focal points of the respective lenses at distances F from the edge of the lens.

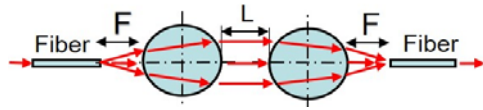


Fig. 1. A simplified optical scheme of the ORC.

The collective drawing of ORC is given in Fig. 1. The basic units of ORC are cases 1 and 2; inserts 3 and 4 with a lens; lenses 5 and 6; double row ball bearing 7; standard threaded FC-type connectors 8 and 9.

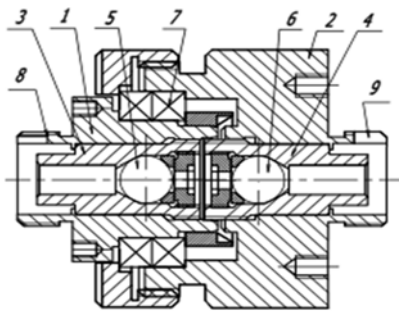


Fig. 2. Collective drawing of ORC with ball lenses.

Most commercial products use fiber-optic pigtails glued to the ORC body on both sides to connect the ORC to the fiber. In the developed design of ORC (Fig. 2) pigtails are not used. Fiber optic cables are connected to the bodies 1 and 2 via standard FC-type threaded connectors.

Such a circuit solution allows the use of ORC with different types of fiber cables and gives the product certain advantages in the possibility of wider application. In addition, this approach greatly simplifies and reduces the cost of design. To ensure the ORC maximum manufacturability, its theoretical analysis was performed in terms of accuracy.

According to the analysis, the necessary requirements for the accuracy of manufacturing of individual structural elements and their assembly in housings were determined. This allowed to optimally form the requirements for tolerances and landings of individual parts and assembly units, which were the basis of design documentation. To conduct a theoretical analysis of ORC with spherical lenses, a physical and mathematical model of ORC in the framework of geometric optics was built and a computer program for computer simulation of ORC in the software environment of the Matlab package was created.

One of the results of the theoretical analysis of the device (the calculated coupling loss against rotational angle) is shown in Fig. 3. These results take into account the optical Fresnel losses on the optical surfaces of the lenses, which are about 26 %.

Calculations were performed for cases of signal transmission on multimode fiber-optic cables with a

core diameter $50\ \mu\text{m}$ and numerical aperture $NA = 0.16$, lenses diameters $D1 = D2 = 4\ \text{mm}$, lens refractive index $n = 1.75$. The displacement of the axes of the optical fibers relative to the optical axis of the lens is indicated in the figures by the parameters $\Delta Xf1$ and $\Delta Xf2$, the displacement of the centers of the lenses relative to the axis of rotation are denoted as $\delta Xlens1$ and $\delta Xlens2$.

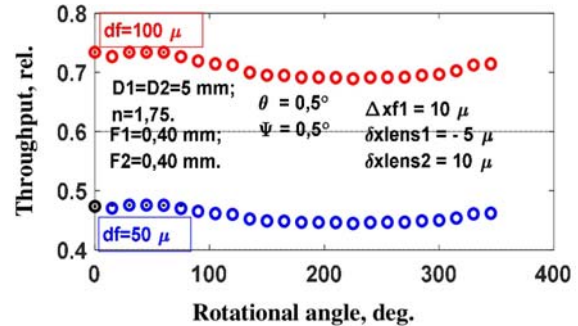


Fig. 3. The variation of optical throughput with rotational position.

It should be noted that the results of calculating optical losses in the optical fiber are strongly influenced by the radiation source radiation pattern and the numerical aperture of the receiving fiber. The rotational characteristic in Fig. 3 was obtained for a laser radiation source with a limiting radiation angle of 7° and for a receiving fiber with a numerical aperture of $NA = 0.18$.

3. Conclusions

Optical rotating connectors built on the basis of ball lenses meet the requirements of modern multimode fiber-optic communication lines, both in terms of optical losses and the amount of rotational oscillations of the optical signal. The device can be used with bidirectional data links and at wide wave range (lengths of the 850-1550 nm).

In our experimental samples of ORC with uncoated lenses, optical losses were 3 dB when operating ORC in the composition of standard fiber-optic cables with a core diameter of $50\ \mu\text{m}$. Optical losses will not exceed 1.5 dB if you use lenses with anti-reflective coatings. The variation of insertion loss caused by device rotation was less than 0.25 dB.

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Data Slicing Model Proposals for Low-availability Smart Metering Equipment

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Summary: Following the analysis of load profiles and other patterns of energy use using novel algorithms, augmentations are sent back to the smart meters in different formats. This paper proposes a data slicing model for large data files which have to travel securely and reliably throughout the smart grid. Through using data slicing with our proposed model, data transfers were fully reliable without any information mislaying. Furthermore, devices spread across the grid have seen the effective delivery of broad profiling data frames, complex energy tariff schemes or active software amendments.

Keywords: Data slicing, Smart grid, Low-availability networks, Smart metering profiling, Meter firmware update, Narrow-bands.

1. Introduction

In order to maximize resource utilization in an efficient smart grid, power supply management is handled by smart metering equipment. This equipment can gather large amounts of information that need to be sent for analysis along the power lines back to an energy supplier. After the analysis of load profiles and other energy usage patterns by using novel algorithms, augmentations are sent in different formats back to the smart meters. The information at the end device can come as sets of parameters or as entire new software packages that fit the user profile at the targeted household.

In order to handle transition to a rapidly expanding Smart Grid, it is important to be able to update software remotely, such as the one seen in meters, without changing the equipment or doing upgrades manually in the field. Remote image uploading functionality enables those meter to get updated characteristics on an as-needed basis [1].

While the deployment of smart meters in smart grids and smart cities aim to provide more energy options and energy-saving choices for customers, the challenges of obtaining information from and to end users across the smart grid have been highlighted frequently [2, 3].

Smart meters can talk to their main energy providers through power lines, radio or mobile communications. It is often the case that some nodes in the smart energy grid have a low network availability and sending a few data frames is the best that can be achieved at isolated ends of the grid or in very noisy environments.

1.1. Current State of the Smart-grid

For the electricity market in Romania, the chosen deployment model has a middleware layer with data

concentrators connected by wire to the smart meters. Communication of data is made through Power Lines Communication (PLC) wiring from the meters to the concentrators and through various communication ways from concentrators to the central application. The national rollout plan aimed for 80 percent coverage in the country by 2020 and complete coverage by 2022. As for the chosen communication protocols so far most of the market uses PLC, M-Bus, Wi-Fi, RS-485, cellular network. PLC systems have a divided market between the G3 and Prime Narrow-Band PLC protocols [4]. The latter technologies, while they have great advantages of not requiring new wiring during deployment, suffer from interference generated by consumer equipment's or from large distances between households. The chosen equipment requirements by ERBD and the Romanian Authority for Energy Regulation are robust and reliable in high density areas but the limitations on wiring network and low throughputs of the chosen protocols become challenging when we need to deal with large amount of data that needs to be moved across the grid. While the deployment of the smart meters has not finished yet, the requirements are already at the upper limit of the grid.

The DLMS/COSEM (IEC 62056, EN13757-1) global standard for smart energy metering already requests certain data structures such as tariffs that can change dynamically every other minute or support for multiple energy suppliers and most importantly load profiling of all supported tariffs for all supported contract at a resolution of 1 minute for all types of energies (active, reactive, imported/exported and by quadrant). Such complex structures usually overwhelm the chosen protocols and can easily reach sizes of several megabytes. On the other hand, performance evaluations of the two narrowband PLC systems (PRIME and G3) show that this transmission can be achieved at a greater speed on PRIME but it behaves worse with noise interference [5, 6]. This can

be improved by using the correct data slicing and hybrid communication distribution across multiple protocols as proposed below.

1.2. Proposed Solution

This paper proposes a data slicing model for large data files that need to travel throughout the smart grid in a secure and reliable manner. To reach this goal, our solution needs to avoid retransmissions, to have a high tolerance on erroneous transmitted packets and to be capable of assembling back the initial file whenever we detect a complete transmission. Slices might be delivered on separate communication channels where those are available on the smart-meter itself or on neighboring peers.

For this application we have used the chosen meter technologies required by the regulation authorities: PLC communication using PRIME protocols applied on ST smart-metering evaluation hardware kit EVLKSTCOMET10-1 based on a Cortex-M4 core. The default configuration is fit for PRIME (ITU G.9904) CENELEC A-band protocol standards and the PLC line coupling interface is designed to allow the device to transmit and receive on the AC mains line using any narrow-band PLC modulation (single carrier or OFDM) up to 500 kHz. [7] The end points are connected in series and the network is controlled by a third board acting as a PRIME concentrator. All devices are fit with a W25Q16JV external flash for storage of 16 Mbits of load profiles in DLMS formats.

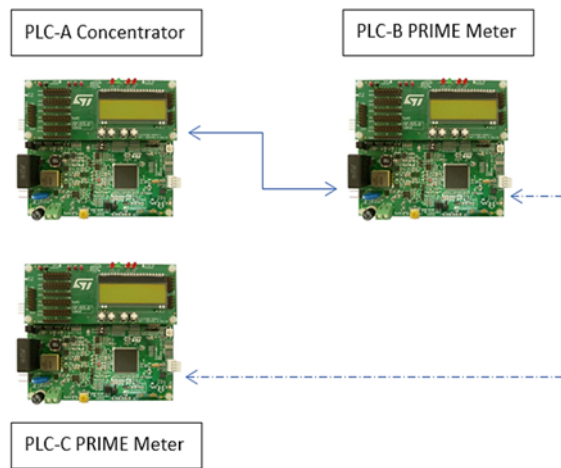


Fig. 1. PRIME data slicing test network topology.

2. Implementation

Each type of network has a maximal amount of data that can be moved across nodes at any given time. In computer networking, the maximum transmission unit (MTU) is the size of the largest protocol data unit (PDU) that can be communicated in a single network layer transaction. Depending on the chosen method of transport, the data slicing approach needs to account

for the transport type constraints. In a typical Ethernet v2 connection, an MTU of 1500 bytes is expected [8].

For devices operating over 802.11 Wi-Fi networks, an MSDU (media access control service data unit) of 2304 bytes is used instead. Although larger packets mean fewer transmissions and less traffic used for overhead transportation, there are certain trade-offs that come with bigger chunks of data being transmitted at once. If we are using a link with a slow speed of processing a large packet, will keep the network busy for a longer time, reducing the availability of those nodes till transmission is complete. Another known issue of larger packets is that even slight errors detected in large packets will cause the entire frame to be retransmitted. In environments that are prone to data corruption, smaller data slices will be used instead.

Our proposed solution consists of slicing large data packets into chunks that are equal to the smallest MTU in the communication chain.

2.1. Packet Structure

Each block of data being sent over the network is preceded by an overhead called data slice header. A packet has two sections for each layer: the header and the body. The header provides protocol details specific to the network, while the body comprises data for the node, mostly made up of an entire packet from the next network. The details derived from the layer above each layer is viewed as data and its own heading refers to the above data. This cycle of data protection is regarded as encapsulation when adding a new header.

In our protocol, the body contains the updated binary information in a form or another and the header provides sufficient information for the receiver to be able to process, validate and control the reconstruction of the end file. This is commonly achieved by appending a sequence ID and a total number of slices to each individual data slice.

PRIME v1.3.6 used in currently deployed smart meters uses differential modulation with one of three possible constellations: DBPSK, DQPSK or D8PSK. If the channel environment is good and no $\frac{1}{2}$ rate convolutional coding is required, it can support speeds up to 47 kbps, 94 kbps and 141 kbps respectively. Table 1 pictures the raw data speed expected in a common PRIME network [9].

Given that those throughputs are ideal and the amount of overhead is significant, moving the large data blocks around the network will cause fragmentation throughout the entire OSI layers. The current smart-grid implementation looks as follows in Fig. 2.

D8PSK modulation with no convolutional code has been chosen to achieve the highest PRIME throughput with an MSDU = 2268 bytes. To avoid having any segmentation and reassembly on PRIME layer, we need to look at its Common Part Convergence Sublayer, CPCS. The CPCS functionality is responsible for splitting outgoing data into CIMTU constant segments of 256 bytes each. PRIME 1.3.6

supports up to 64 segments of CMTUSize. This leads to a maximum transmission length of 16384 bytes after which upper layers need to handle further segmentation and reassembly.

Table 1. PHY data rate and packet size parameters, for various modulation and coding schemes.

Modulation and coding	DBPSK	DBPSK	DQPSK	DQPSK	D8PSK	D8PSK
Convolutional Code (1/2)	On	Off	On	Off	On	Off
Raw data rate (kbps approx)	21.4	42.9	42.9	85.7	64.3	128.6
Maximum MSDU length with 63 symbols (in bits)	3016	6048	6040	12096	9064	18144
Maximum MSDU length with 63 symbols (in bytes)"	377	756	755	1512	1133	2268

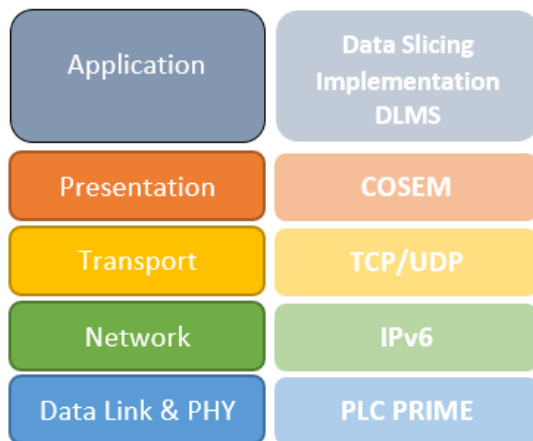


Fig. 2. PRIME smart energy meters OSI layers.

Segmentation at application level will follow the same overhead structure as the PRIME Phy Common Part Convergence Sublayer as described in the R1.3.6 standard in Table 2.

Table 2. Segmentation header fields.

Name	Length	Description
Type	2 bits	Type of segment. 0b00: first segment; 0b01: intermediate segment; 0b10: last segment; 0b11: reserved
NSegs	N bits	'Number of Segments' - 1.
SEQ	N bits	Sequence number of segment.

In order to define a fit N number of supported segments, we will look at one of the most common failed transmission in poor smart PLC networks which consists of retrieving the energy profile of the previous day with a resolution of 5 minutes:

$$\begin{aligned} \text{LoadProfileDaySize} &= \text{NumTypesOfEnergyRegisters} \\ &\quad * \text{RegisterSize} \\ &\quad * \text{NumOfSupportedTariffs} \\ &\quad * \text{NumOfSupportedContracts} \\ &\quad * \text{MinTimeResolution} \end{aligned}$$

$$\text{LoadProfileDaySize} = 6 * 8 \text{ bytes} * 16 * 6 * 288$$

$$\text{LoadProfileDaySize} = 1327104 \text{ bytes}$$

$$\text{NSegs}_{\min} = \text{LoadProfileDaySize} / \text{CMTUSize}_s$$

$$\text{NSegs}_{\min} = 1327104 \text{ bytes} / 256 \text{ bytes}$$

$$\text{NSegs}_{\min} = 5184$$

Considering the above a number N = 16 bits will suffice, leading to an overhead of 34 bits for segmentation logic.

To achieve the highest throughput, all tests are run using D8PSK modulation and coding with no convolutional code.

3. Results

We have run the proposed solution in a residential building, using an evaluation data concentrator connected through power lines to DLMS PLC PRIME boards as in Fig. 1. The chosen modulations and codings could reach a theoretical speed of 128.6 kbps but the size of the daily load profile and the noise on the network made the data transfer impractical.

By using data slicing with our proposed pattern, data transfers became reliable with complete load profiles being transferred from PLC-C PRIME meter board back and forth to PLC-A PRIME concentrator through PLC-B PRIME meter. By choosing slices close to MTUs of Fig. 2 layers, the following tests have been carried out:

Test 1. Initial test throughput with an emphasis on the rate at which slices needed retransmission during a

normal transmission of a Daily Load profile reading that has a size LoadProfileDaySize = 1327104 bytes;
Test 2. Sliced test with the transmission of a Daily Load profile reading configured for a data slice size = 256 bytes. This will split the formatted data into 5184 slices.

Figs. 3 and 4 show the improvements in retransmission rates from an average of ~13 % of slices needing a retransmission down to 0 % retransmission on CIMTUSize sized slices. On the other hand, Figs. 5 and 6 show the cost of having a more paced transfer of slices, the throughput went from ~90 Kbps down to ~6 Kbps, allowing a full day profiling to be transmitted in ~20 minutes in the most exhaustive format.

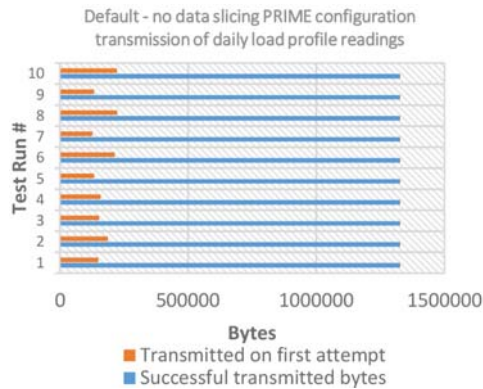


Fig. 3. Initial results on daily load profile reading over PLC-PRIME from PLC-A to PLC-C.

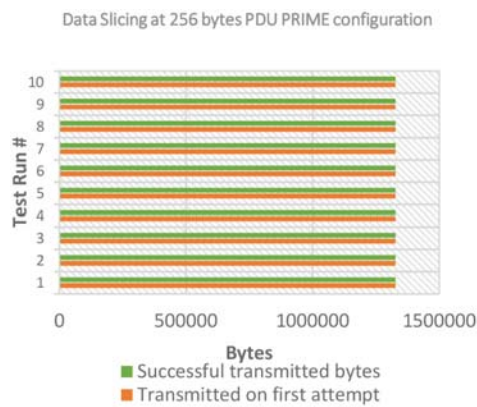


Fig. 4. Transmission of daily load profile readings with confirmation on receiving with no retransmissions required.

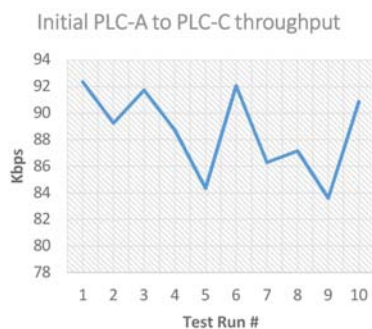


Fig. 5. Thresholds observed on unreliable PLC communication channel with no data slicing at application level.



Fig. 6. Thresholds observed on PLC-A to PLC-C communication channel with data slicing at application level.

4. Conclusions

Results from pilot trials on smart meters running the above-described model for data slicing in smart grid environments with low-availability have proven the approach to be working correctly. By using data slicing with our proposed pattern, data transfers became more reliable with no need for retransmission attempts all at a cost of throughput. In addition, devices scattered around an unreliable network have seen a noteworthy increase in successful transmission of large profiling data frames.

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Optimizing a Non-contacting High-sensitivity GMR-based Current Sensor Design for Low Field Applications

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Summary: This paper presents a practical method of improving the sensitivity and accuracy of the response of a GMR-based current sensor. The novelty of our approach consists in utilizing a multiple trace current detection setup in a double differential system on a custom prototype PCB (printed circuit board); In essence this would constitute a planar coil that will increase the usable field detected by the GMR sensor. In support of these claims, an analytical model was developed for calculating the magnetic field present in GMR sensing area. Also a custom GMR-based current sensor setup was developed which showed an 8.3 to 13.2 times increase in sensitivity compared to a single biased sensor. The system can monitor currents for prolonged periods of time from 500 mA currents below 5 mA, however, for a linear response without EMF shielding, a current of a minimum of 20 mA is necessary. In terms of prospects, these methods of improving the sensitivity of GMR sensors can be useful for a number of applications requiring low field detection. For example, these improvements can be applied to a high sensitivity detection setup of magnetic nanoparticles (MNPs) used to label biomolecules in lab-on-a-chip (LOC) applications.

Keywords: Current sensors, GMR effect, spin-valve sensor, Magnetoresistive sensors, Bias magnetic field.

1. Introduction

Often, for modern electronics, the issue of detecting low current values accurately and reliable with low cost solutions can prove critical. This paper is a natural continuation and expanding of the authored work presented in [1].

Non-contacting current measurement methods (Current transformers, Rogowski coils [2], (micro)fluxgate [2], MR sensors (magnetoresistive sensors: anisotropic magnetoresistance AMR [3], giant magnetoresistance GMR [4], TMR [5]) present a number of advantages from which we can note galvanic isolation, high resistance to external magnetic fields, thermal stability. Compared with AMR sensors, GMR sensors offer more extensive high sensitivity capabilities and are not as susceptible to overcurrent or magnetic field spikes.

The most notable drawbacks of GMR sensors are non-linearity, hysteresis, offset, and a temperature dependent output. Regarding methods to improve the GMR sensor response, several methods have proven to be effective. By using a bias field parallel to the sensitive axis can shift the operating point of the sensor to the linear region, thus reducing the hysteresis behavior and creating a bipolar signal. This magnetic field can be created either by using a permanent magnet or a coil system with DC, AC, or short pulse currents which can have either open or closed-loop control [4]. Also, when being used as current sensors, some noteworthy methods to improve the GMR sensors response are: closed loop operation [6], hysteresis modelling compensation or low frequency capture.

2. Materials and Methods

2.1. Principle of Operation

The novel advancements of the proposed setup consist in the design rules to develop a GMR-based current sensor setup to improve sensitivity up to 10 times compared with a single GMR sensor. The current measurement setup is an indirect one (the GMR sensors act as a magnetometer). Thus, if a current, I , passes through a wire, the magnetic field B will produce a change of the output voltage on the GMR sensor. The working principle of the setup can be seen in Fig. 1.

As shown in [1], the low currents sensing capabilities can be improved with a narrower trace. Thus, the method shown will focus on creating a planar coil setup (Fig. 2) - the multiple traces shown are in fact the same trace in a spiral pattern.

The current, I , from the conductive trace (denoted as "Current trace") generates a magnetic field, whose component, B_x , will be detected by the GMR sensor. To estimate B_x , we derived an analytical method based on [1], which assumes that the sensor is centered above the trace at distance h (Fig. 2).

Thus, assuming a long conductive trace, the elementary field produced by the current dI is expressed, using the Biot-Savart law, by eq. 1,2 (where $\mu_0 = 4\pi \times 10^{-7}$ H/m is the vacuum magnetic permeability, D is the trace width, t is the trace thickness, h is the distance between the sensing element and the trace, and θ is the angle from Fig. 2).

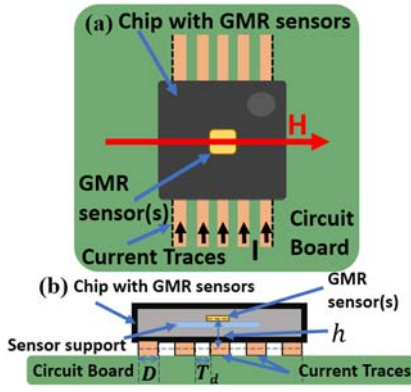


Fig. 1. Non-contacting current measurement setup using multiple traces and a GMR based sensor chip: (a) plane view; (b) cross section (adapted from [1]).

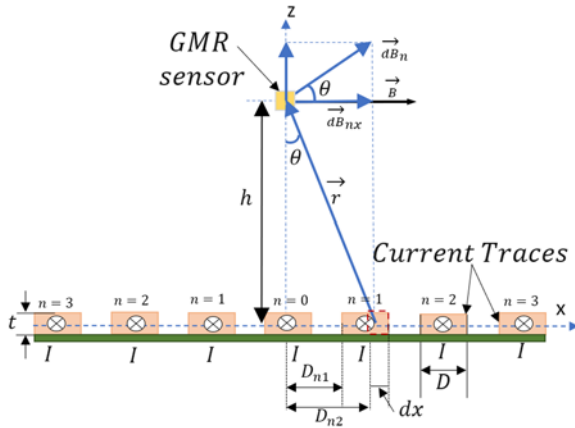


Fig. 2. Cross-section and parameters for the analytical model for field calculations on multiple traces.

$$dB_n = \mu_0 \frac{Idx}{2\pi r} = \mu_0 \frac{Idx}{D} \cdot \frac{1}{2\pi\sqrt{h^2+x^2}}, \quad (1)$$

$$dB_{nx} = dB_n \cdot \cos \theta = \mu_0 \frac{Idx}{2\pi D} \cdot \frac{1}{\sqrt{h^2+x^2}} \cdot \frac{h}{\sqrt{h^2+x^2}} \quad (2)$$

By assuming a linear current density, integrating equation (2) and some basic calculations, the result can be extrapolated for an "n" number of traces. We obtain:

$$B_n = \frac{\mu_0 I}{2\pi D} \left[\tan^{-1} \left(\frac{D_{n2}}{h} \right) - \tan^{-1} \left(\frac{D_{n1}}{h} \right) \right] \quad (3)$$

From Eq. 3, the various n fields can be added to calculate the total magnetic field in the sensor area. Note that for only one trace ($n = 0$), the result is detailed in [1]. An analysis was performed with this method for two different trace structures and number of traces (Fig. 3). From Fig. 3 we can note that the magnetic field in the sensor area is increased significantly from the case with just one trace (the results were obtained with a 0.5 A current passing through each trace). The best balance between the obtained field and practicality of implementation was

found to be at the 5 and 7 traces structure as there is an asymptotic dependency. Thus, a prototype PCB for current sensing with the specifications from case II (for higher magnetic field output) will be developed. From [1], we can denote that other factors also play a major role in the sensitivity of the sensors (such as biasing of the sensors, differential measurement setup, amplifying the output etc.). Thus, we can expect much better results with an optimized designed of the setup and appropriate biasing of the GMR sensors. From the results obtained with the analytical algorithm, we estimate that compared with a single biased NVE AA003-02 GMR sensor (supplied with 1 mA of current, sensitivity S around 13.75 mV/Oe), by using multiple traces on an optimized double differential design as in [1], we can expect a sensitivity S increase for a 500 mA current from $S = 0.0275$ mV/mA (single sensor) to $S = 0.207$ mV/mA (5 traces, case I) or $S = 0.285$ mV/mA (7 traces, case II) (notice Fig. 3) which is a projected 7 to 10 times increase in sensitivity.

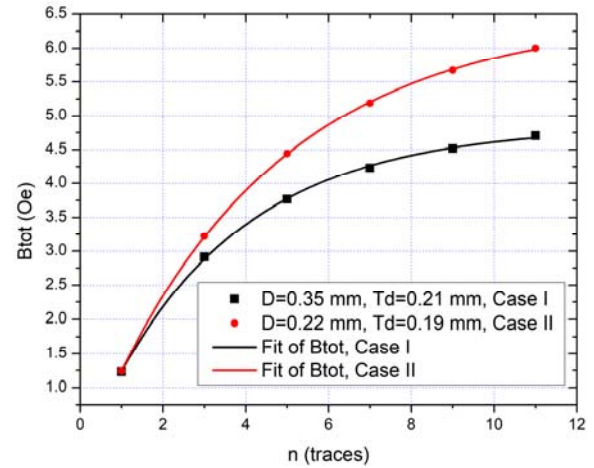


Fig. 3. Total magnetic field in the GMR sensor area (for $I = 0.5$ A) dependency on the number of current traces. The asymptotic tendency of B_{tot} for the case where $D = 0.22$ mm is 6.485 Oe (case I) while for $D = 0.35$ mm is 4.849 Oe (case II). Also, the field multiplying effect is clearly visible.

2.2. Characterization of the GMR Sensor

Commonly, GMR sensors are made from multilayered structures of the type AFM (antiferromagnetic layer)/PL(pinned magnetic layer)/NM (non-magnetic space layer)/FL(free layer). The free layer is the sensing layer, as the magnetization can rotate upon an applied magnetic field).

The rate of change in the resistance of a GMR sensor element is expressed by:

$$GMR = \frac{R_{ap} - R_p}{R_{ap}} \cdot 100 [\%], \quad (4)$$

where, R_p and R_{ap} are the resistance values for the parallel and antiparallel configurations of the magnetic

layers. The GMR ratio for such structures is around 5-15 % [4]. For the NVE AA003-02 GMR sensor, which was used in our setup, a GMR ratio between 13-16 % is reported [6]. In order to characterize the AA003-02 sensor (Fig. 4), a magnetic field was applied by two round-shaped coils in a quasi Helmholtz like configuration (note Fig. 5).

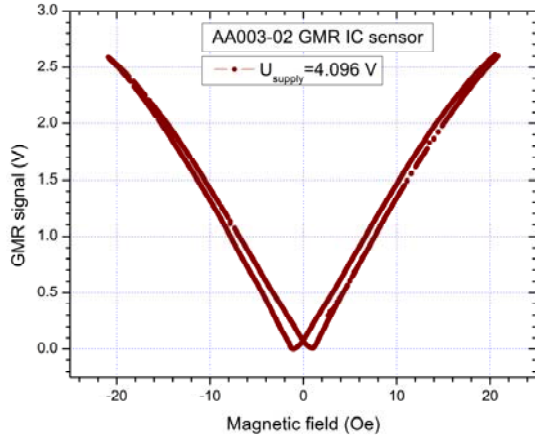


Fig. 4. Typical measured field dependencies for the AA003-02 GMR sensor. In this case, the sensor was supplied with a constant 4.096 V supply.

2.3. Design Rules and Measurement Setup

Based on the results from the analytical method, a GMR-based current measurement was developed (the PCB setup can be seen in Fig. 5). In terms of design rules for the current traces, a single trace in a U shaped spiral pattern (7 traces) passes through the sensors. Also, in order to amplify the useful magnetic field, the characteristics are those from Case II (note Fig. 3). Due to the thickness of the trace, the setup is designed to operate with currents up to around 0.5 A for prolonged periods of time. The GMR sensors, from an operational point of view, operate in a similar differential setup as the one detailed in [1].

The output voltage from the sensors are amplified by two (one for each sensor) INA118 instrumentation amplifiers integrated on the board. The real gain for each amplifier was set to 10 and the results were offset corrected. The data acquisition system is done with an NI 6281 USB data acquisition board. The implementation allows measurement of both AC and DC currents with the ability to bypass the included filtering system for the amplifier output. An LM135AZ temperature sensor was also integrated in the sensor area. It should be noted that the supply lines for the GMR and temperature sensor are orientated perpendicularly compared with the current trace (any produced magnetic field lower 20 Oe (Fig. 4) is not detected by the GMR sensors) as to eliminate any additional magnetic field.

The biasing coils are placed in a symmetrical quasi Helmholtz like configuration because of the requirement for a minimum distance was ensured between each GMR sensor (such as the traces beneath

each sensor to not influence the output of the other sensor). More studies can be done to find out the minimum possible distance between the GMR sensors to ensure proper results. This effect could have been mitigated by using larger biasing coils but that would have contributed to the costs and power consumption of the system. As a prepolarization field is necessary for linearizing the GMR sensors output, the biasing for the sensors was set to 8 Oe (this corresponds to a 58 mA current passing through the coils). In Fig. 6, the functional block diagram for the current measurement system as well as the amplifier setup is shown.

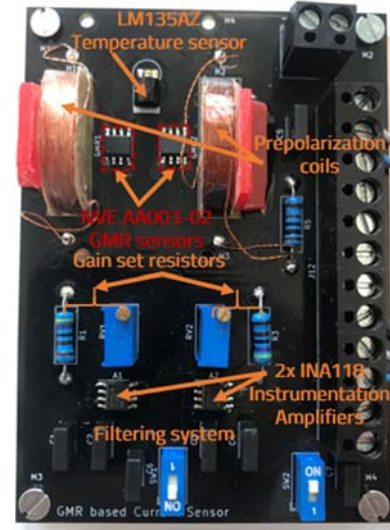


Fig. 5. Custom PCB optimized for low fields detection (current measurement) using GMR sensors.

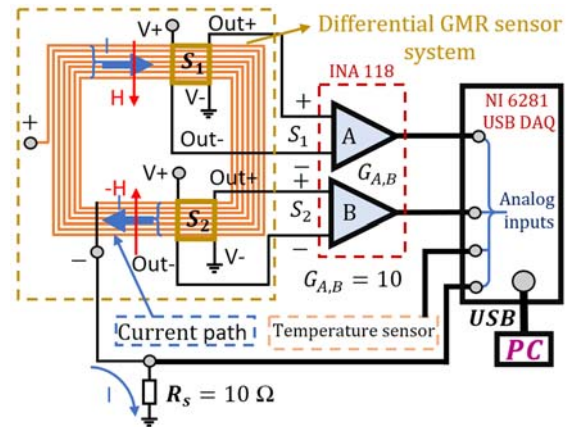


Fig. 6. Functional block diagram for the multi-trace current measurement differential system optimized for low currents sensing.

Furthermore, if we take into account the thermal influence for the sensors response, by considering that the system is thermally balanced and the same type of sensors are used, the total output voltage of the sensors from the differential system can be expressed with:

$$\Delta U = (K_{S1} + K_{S2}) \cdot H_I = (K_{S1} + K_{S2}) \cdot C \times I = S \cdot I, \quad (5)$$

where K_{S1} and K_{S2} are the sensitivities of each sensor, S (mV/mA) is the sensitivity of the differential measurement system, $H_I = C \cdot I$ (C is a constant) is the magnetic field created by the current passing through the trace.

3. Results and Discussion

The results presented in this section are a summary and focus on demonstrating the low currents sensing capabilities of the proposed setup (Fig. 5). Note that for all results, the real sensor sensitivity is shown (without amplification). The biasing field for both sensors is set at 8 Oe. In Fig. 7, the response obtained with the setup by measuring a variable DC current between ± 150 mA is shown. The sensitivity of the measured differential

output is $S_{measured} = 0.2352$ mV/mA. The calculated sensitivity, using the analytical method is $S_{calculated} = 0.285$ mV/mA which shows a good correlation between the theoretical and experimental results. Compared with the results presented in [1], for the NVE AG003-01E sensor evaluation kit (which utilizes the same model of sensors), measured on a single trace with a similar width 0.254 mm, the obtained sensitivity is $= 0.0179$ mV/mA while for the differential system in [1] with a trace thickness of 4 mm, for the same 150 mA test, the obtained sensitivity $S_D = 4 \text{ mm} = 0.02845$ mV/mA. Thus, in this test, with the multi-trace setup an increase in sensitivity of ~ 13.2 times compared to the sensor evaluation kit was obtained and ~ 8.3 times compared to the already optimized differential setup from [1].

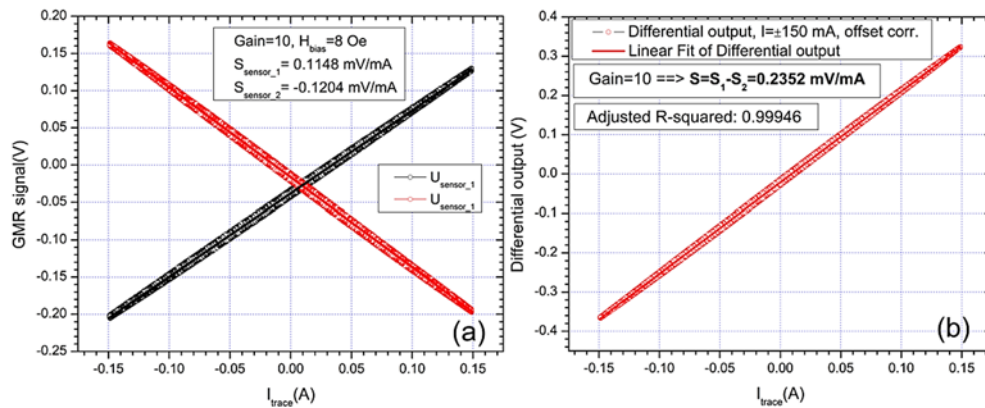


Fig. 7. Differential output of sensors polarized at 8 Oe, DC, ± 150 mA: (a) Individual sensors response; (b) Differential output.

Due to the significant gains in sensor sensitivity, even a 20 mA current can be detected reliably (Fig. 8). The obtained sensitivity for the differential output in this case is $S = 0.2204$ mV/mA. By using the analytical algorithm, for a 20 mA current, an estimated field $B_{tot} = 0.2075$ Oe was produced by the conductive trace band in the sensor area.

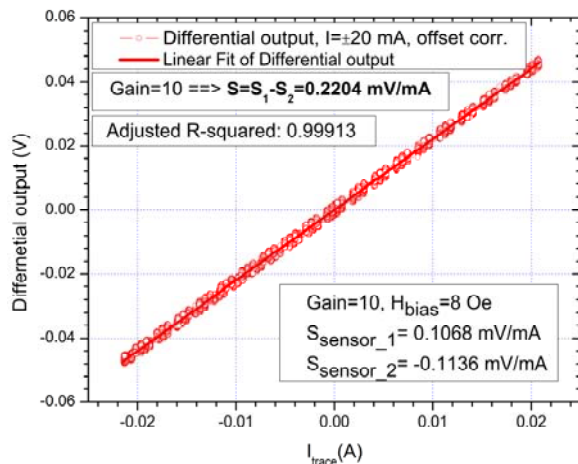


Fig. 8. Differential output of sensors polarized at 8 Oe, DC, ± 20 mA differential output.

This field and sensitivity is sufficient for the detection of MNP (magnetic nanoparticles) with the GMR sensor [8], for example, by sending oscillatory currents through the planar coil which can prove useful for biofunctionalizing the setup if the MNPs are used to label biomolecules. applications and biosensing applications with magnetoresistive sensors. The system can detect currents below 5 mA but the measurement setup can be improved to ensure better linearity and lower noise (for example with further amplification steps EMF shielding of the acquisition lines).

4. Conclusions

The theoretical (analytical method) and operational basis (design rules) for developing a multi-trace GMR-based high sensitivity current sensor PCB has been established. From those we can note: appropriate biasing of the sensors, adequate spacing of components to avoid parasitic magnetic fields, multi-trace and differential or double differential design, appropriate amplification and data acquisition. Versatility in this design allows enhanced characteristics and an increased operation range also for high currents sensing (by integrating a conductive

bar on the PCB backside). The developed system, has shown a theoretical increase of 7 to 10 times increase in sensitivity and an 8.3 to 13.2 increase in sensitivity compared with a single, similarly biased AA003-02 sensor. To sum up, the benefits of the developed design rules for successfully increasing the operational range and usability of the GMR sensor system have been demonstrated with potential to be implemented used for MNPs detection in Lab-on-a-chip applications.

Acknowledgements

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(047)

Planar Hall Effect Sensors for Low Field Detection and Lab on a Chip Applications

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Summary: Disk shaped Planar Hall (PHE) effect sensors, deposited from Permalloy on Si substrate, were used to demonstrate their ability to detect very small magnetic fields and, in particular, their using for magnetic nanoparticles (MNPs) detection. We show that, by using convenient biasing fields, we can fine tune the sensor sensitivity and his field behavior. Sensitivity up to $6 \mu\text{V}/(\text{Oe}\times\text{mA})$ was found for a field range of about ± 10 Oe. To detect MNPs, we used a surface-based detection method in which the presence of MNPs over the sensor surface is sensed through the change of the PHE field characteristics. We found that magnetic moments down to 10^{-6} emu can be detected. These sensors can be integrated in Lab on a Chip devices

Keywords: Magnetic sensors, Anisotropic magnetoresistance effect, Planar Hall effect, Magnetic nanoparticles, Lab on a chip.

1. Introduction

Many applications require magnetic field sensors with high sensitivity, low hysteretic behavior, low noise and low thermal drift. Some special applications require magnetic sensors with tunable properties to fine adapt their field characteristics to the specific use. Various physical phenomena are employed to build magnetic field sensors such as: search-coil, (micro)fluxgate sensors [1], magnetoresistive (MR) and Hall effect sensors, which are based on galvanomagnetic effects in semiconductors and magnetic thin films [2]. Magnetoresistive sensors microfabricated from magnetic thin layers are considered as highly sensitive and are used to detect magnetic fields between 10^{-5} to 10^2 Oe [3, 4]. This paper is focused on using planar Hall effect sensors for magnetic field sensing and magnetic nanoparticles detection. The planar Hall effect (PHE) is a consequence of the anisotropic magnetoresistance effect (AMR), discovered by William Thomson in 1856. AMR effect can be found in bulk ferromagnetic or thin films from Ni, Co, Fe and their alloys. In short, the AMR effect comes from the dependence of the electrical resistivity of a magnetic material on the angle between the direction of electric current and direction of magnetization inside the material. Such that, the measured electrical resistance has maximum value when the direction of current is parallel to magnetization and a minimum value when the direction of current is perpendicular to magnetization. A short theoretical background used to describe the AMR and PHE is presented in what follows [4-6] and referred to Fig. 1. For magnetic thin films, the magnetization is forced to stay in the film plane. Such that, when a current flows through the sample, the AMR effect yields a transverse electric field in a geometry typical for Hall effect but with the applied field contained in the film plane, Fig. 1. A widely used

material is the alloy $\text{Ni}_{80}\text{Fe}_{20}$ (Permalloy) due to its magnetostriction constants close to zero in all crystal directions. Usually, an easy axis of magnetization is defined by H_K through the shape anisotropy ($l > w$) and the intrinsic anisotropy field due to crystalline structure of the magnetic layer and microfabrication process [4, 6]. An exchange biasing field, H_{EB} , may be present, and appears at the interface between the ferromagnetic layer and an antiferromagnetic layer like FeMn or IrMn [3, 6]. This field acts like an external biasing field applied to sensing layer and has the role to keep the initial magnetization state for $H = 0$; usually, H_{EB} ranges between 10 to 100 Oe.

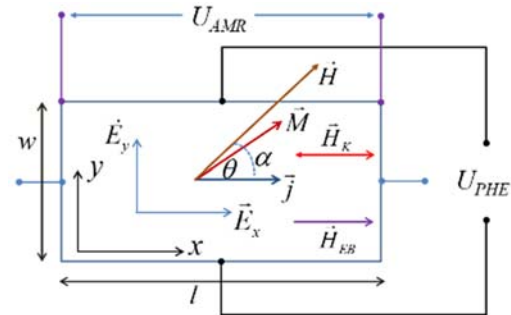


Fig. 1. Schematic representation of a magnetic thin film through which is flowing a current along x axis; M is the magnetization which makes an angle θ with the current due to the applied field, H . A magnetic anisotropy, defined by H_K and an exchange biasing field, H_{EB} , may be present.

The longitudinal, E_x , and transverse, E_y , electric fields can be derived by considering the angular dependency of the resistivity tensor components ρ_{xx} and ρ_{xy} under the applied field H [4, 7]:

$$\begin{aligned} E_x &= j_x \cdot \rho_{xx} = j_x \rho_{\perp} + j_x (\rho_{\parallel} - \rho_{\perp}) \cos^2 \theta, \\ E_y &= j_x \rho_{xy} = j_x (\rho_{\parallel} - \rho_{\perp}) \sin \theta \cos \theta, \end{aligned} \quad (1)$$

with $|j| = j_x$ and $\rho_{\parallel}$, $\rho_{\perp}$ resistivities expressed at saturation field, parallel and perpendicular to the current direction respectively.

The variation of the longitudinal resistance, given by ρ_{xx} , and measured through, $U_{AMR} = E_x \times l$, characterizes the AMR effect. The second term, E_y , shows generation of a signal perpendicular to the current direction in a geometry typical for Hall effect but with the applied field contained in the film plane. This is the PHE signal, U_{PHE} , Fig. 1, and can be expressed by [4, 6, 7]:

$$U_{PHE} = I_{sens} \frac{(\rho_{\parallel} - \rho_{\perp})}{t} \sin \theta \cos \theta, \quad (2)$$

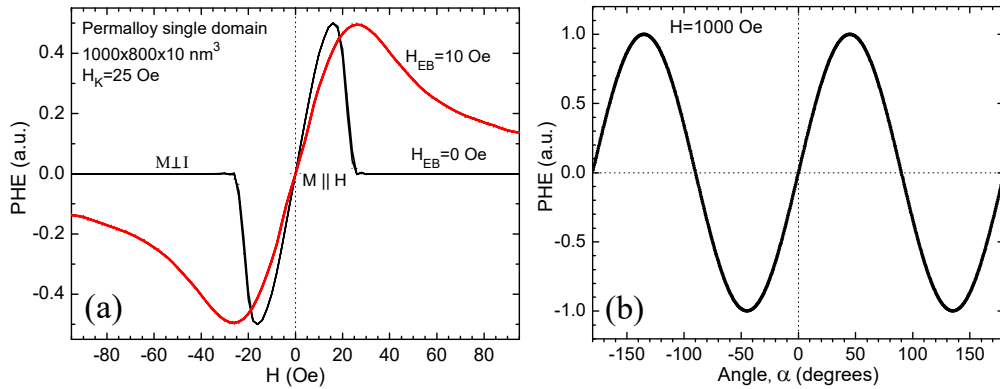


Fig. 2. (a) Simulated field dependencies of the PHE signal for two exchange biasing fields and (b) angular encoder characteristics for a rotating magnetic field, $H = 1000 \text{ Oe}$.

The simulations were made using the Stoner-Wohlfarth (SW) model implemented in SimulMag [9, 10]. These plots, that have a qualitative character, are used to emphasize the main behavior of PHE sensors. We observe that the response of the PHE is linear around zero-field, with a constant sensitivity and the signal is bipolar both for angular and field dependencies. By applying an external biasing field, H_{bias} , or by adjusting, through microfabrication processes, the exchange biasing field along the easy axis [6, 8], we can fine tune the PHE field response. So, it comes that higher sensitivities can be achieved for lower values of H_K and H_{EB} . However, for real structures, smaller values for H_K and H_{EB} means magnetic domains in the sensing layer, that imply nonlinear field dependence and hysteretic behavior of the PHE signal as we will see in the next section. On the other hand, larger values of H_{EB} and H_K bring the sensing layer close to a single domain structure which means a very good linearity of the output signal but with the cost of magnetic sensitivity [4, 6, 8]. The PHE setup is similar to a Wheatstone bridge. Such that, by a careful microfabrication process and an adequate magnetic structure, a very good thermal stability and a noise level of the output signal down to $1.3 \text{ nV/Hz}^{1/2}$, can be achieved [4, 6]. So, the PHE sensors can be used to accurate detect very low magnetic fields, of 10^{-5} Oe . Despite the age of PHE, it is still highly used for

where I_{sens} is the driving current through the sensor ($I_{sens} = j \times w \times t$) and t is the film thickness.

The position at equilibrium of magnetization in the sensing layer, and hence the angle θ between M and j , can be calculated by minimizing the system's free energy density [6, 8]. Note that magnetization rotation can be due to a rotating magnetic field or to a magnetic field, H , applied along y axis ($\alpha = 90^\circ$ in Fig. 1) which inspires two main applications of the PHE sensors: field detection and rotation encoder. Fig. 2(a) shows the simulated field response of the PHE signal for a single domain structure of Permalloy, with dimensions $l \times w \times t = 1000 \times 800 \times 10 \text{ nm}^3$, like in Fig. 1 where $\alpha = 90^\circ$; $H_K = 25 \text{ Oe}$, $H_{EB} = 0$ and 10 Oe respectively. Fig. 2(b) shows the simulated angular dependence of the PHE signal for a rotating field, $H = 1000 \text{ Oe}$.

sensing applications, flexible devices or magnetic memories [4, 6-8] and for fundamental research of some phenomena like magnetization reversal processes in nanostructures, Spin Hall Effect, study of Barkhausen noise in nanostructured systems [11], etc. The PHE sensors are patterned with different geometries in function of the desired applications. The most usual geometries are: cross-shaped [6, 8], elliptical [4, 7] and disk-shaped [10, 12]. Other geometries, that mimic the PHE, consist from ring or diamond shaped AMR resistors connected in a Wheatstone bridge [13]. These sensors are metal-based and can be easily integrated with other components on the same Si chip.

2. Results and Discussion

Disk-shaped PHE sensors, 1 mm diameter and 20 nm thick, were microfabricated on Si/SiO₂ substrate from Permalloy layer, by using photolithography technique. Also, have been patterned Au pads that are in contact with the Permalloy disks, to define the PHE structures, Fig. 3(a). The sensors were passivated with a 200 nm thick TiO₂ layer. A Keithley 6221 programmable current source was used to drive the current through the sensor while a Keithley 2182A nanovoltmeter was used to read the voltage. The chip

was placed inside of two Helmholtz coils that generate H_{appl} and H_{bias} fields like in Fig. 3(a). A computer based DAQ was used to collect and handle data.

2.1. Magnetic Field and Rotation Sensor

Our approach is using disk-shaped PHE sensors in order to avoid in-plane shape anisotropy. This will allow free rotation of magnetization upon an applied

field, H_{appl} . Because there is no exchange biasing field, which assures the initial orientation of M , we used an external biasing field, H_{bias} , which can tune the magnetization state in the sensing layer and, hence, the sensor behavior. This approach gives us the flexibility in building various applications by using a very simple structure. As equation (2) suggests, two main applications can be developed with PHE sensors from Fig. 3(a): magnetic field sensing, Fig. 4(a), and angular encoder, Fig. 4(b).

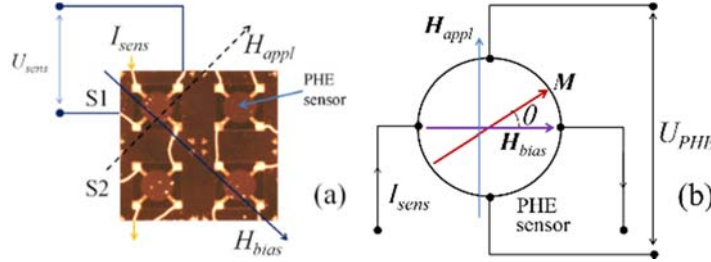


Fig. 3. (a) Top view of the chip with 4 PHE sensors and (b) schematic representation of the disk shaped PHE sensor. The magnetic fields involved and the electrical connections are displayed.

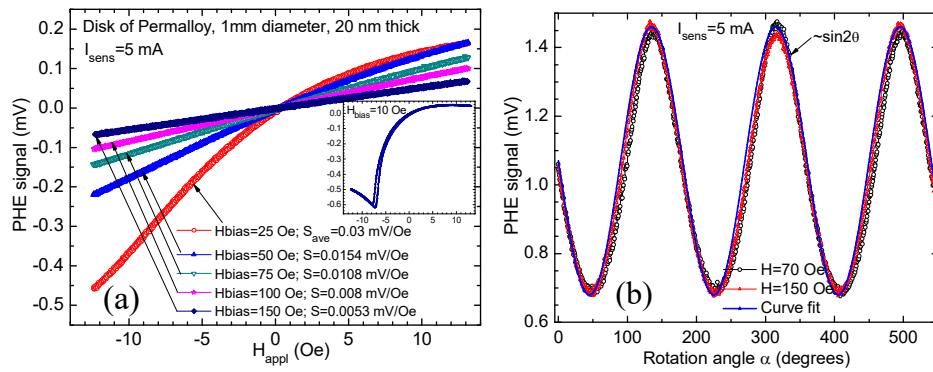


Fig. 4. (a) Low field dependencies of the PHE signal for different biasing fields and (b) angular sensor characteristics for two values of the rotating magnetic field.

The low field dependencies were obtained using the setup from Fig. 3 (a). The applied field is in the film plane, perpendicular to the driving current, I_{sens} , and the biasing field, like in Fig. 1 and Fig. 3. For $H_{bias} < 25$ Oe, the field response $U_{PHE}(H_{appl})$ shows a nonlinear behavior and hysteretic effects (see also the inset from Fig. 4(a)) but the output voltage presents a large variation with the applied field. This is in agreement with the discussion related with data from Fig. 2 and shows that sensors that exhibit hysteresis and low anisotropy present much higher field sensitivity [14]. Such behavior can be beneficial for applications that require detection of magnetic nanoparticles placed above the sensor's surface. The very good linearity of the measured signal for H_{bias} higher than 25 Oe suggests that the main mechanism of the magnetization reversal processes is based on the magnetic moments rotation. Higher values for H_{bias} means a more ordered magnetic state but higher torque is needed to rotate the magnetic moments and the field sensitivity decreases from $15.4 \pm 2.7 \times 10^{-2}$ μ V/Oe for

$H_{bias} = 50$ Oe to $5.35 \pm 4.1 \times 10^{-3}$ μ V/Oe for $H_{bias} = 150$ Oe, as seen in Fig. 4(a). For $H_{bias} = 50$ Oe, the Coefficient of Determination of the fitting line was $R^2 = 0.993$ whereas for $H_{bias} = 150$ Oe, $R^2 = 0.999$ for an applied field variation between -13 to 13 Oe. We observe that by increasing H_{bias} , the maximum value of the applied field for which the sensor has a linear behavior increases [8]. So, a compromise must be chosen in accordance with the application envisioned. The angular sensor characteristics were obtained by using a rotating magnetic field, H , and $H_{bias} = 0$. Because $H_K = 0$ and $H_{EB} = 0$, the magnetization, M , can follow accurately the rotating field, i.e. $\theta = \alpha$ when $H > 150$ Oe, and data can be well fitted with a $\sin 2\theta$ curve, Fig. 4(b).

To illustrate the magnetic moments inside the sensor during the field rotation, we run micromagnetic simulations on a disk-shaped PHE sensor. The results are presented in Fig. 5. The simulated structure is enclosed in a square of $1000 \times 1000 \times 10$ nm³ of Permalloy inside of which has been designed the

disk-shaped PHE sensor. The cell size is $5 \times 5 \times 5 \text{ nm}^3$ and the effective diameter of the disk is 700 nm. The used parameters are: $M_s = 710 \text{ emu/cc}$, exchange constant $A = 1.3 \times 10^{-11} \text{ J/m}$, the anisotropy constant $K_U = 0 \text{ J}$ and $T = 0 \text{ K}$. We used Llg Micromagnetics 4.0 to run the simulations [15].

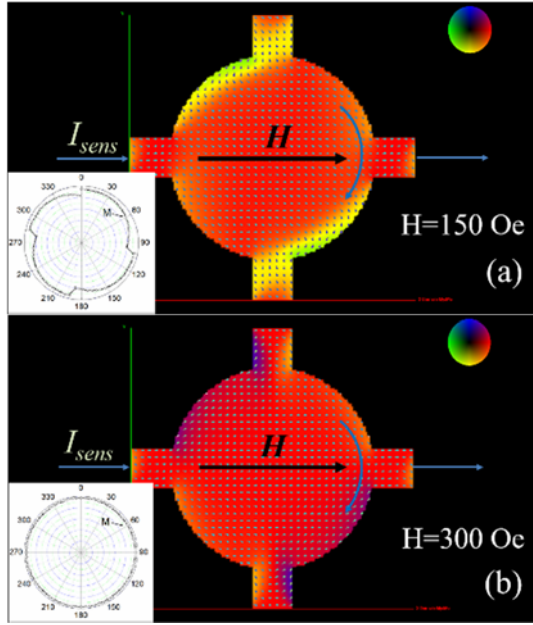


Fig. 5. Illustration of magnetic moments orientations after a full rotation of the field for (a) $H = 150 \text{ Oe}$ and (b) $H = 300 \text{ Oe}$. The color legend illustrates the magnetic moments orientation.

In the lower left corners of Fig. 5(a) and Fig. 5(b) it can be seen the angular dependencies of the magnetization inside the sensor. It comes that for H larger than 150 Oe, the magnetization follows accurately the magnetic field, Fig. 5(b). The magnetic moments orientation is illustrated for $\alpha = 360^\circ$. Despite the fact that these plots, have a qualitative character, because are performed on a sub-micrometer scale structure at $T = 0 \text{ K}$, they are in good agreement with the results from Fig. 4(b).

2.2. Detection of Magnetic Nanoparticles

Finally, we shall discuss detection of magnetic nanoparticles (MNPs) of Maghemite 10 nm in diameter, functionalized with Polyethylene glycol 6000 (PEG 6000), using the disk-shaped PHE sensors presented above. This ability can be useful to develop low cost biodetection systems and lab on a chip devices (LOC) where magnetic sensors are used. In short, magnetic nanoparticles are used to label biological structures of interest which are then detected [16, 17]. Fig. 6 presents the magnetization curves obtained by VSM, at room temperature (RT), for 1.27 mg powder of maghemite nanoparticles functionalized with PEG 6000. From Fig. 6 can be noted the superparamagnetic behavior of these MNPs at RT which is beneficial for detection process because can be minimized false

detection signals when magnetic sensors are used [12, 16, 17]. It has to mention that the blocking temperature for these MNPs is $T_B = 252 \text{ K}$ which is far below the usual working temperatures during the detection experiments.

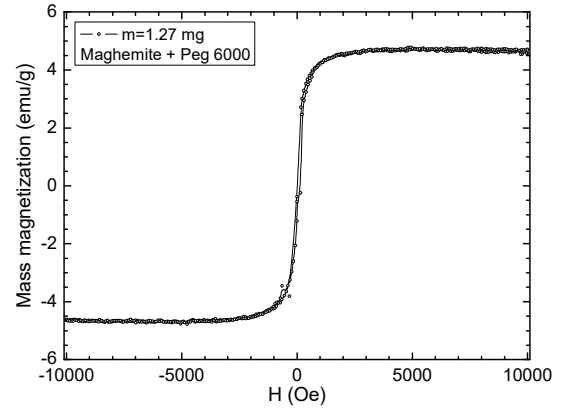


Fig. 6. Magnetization curves obtained by VSM for 1.27 mg powder of maghemite nanoparticles functionalized with PEG 6000.

Two detection schemes were used in the present study to emphasize the importance of the experimental setup. In both cases, a droplet of $0.7 \mu\text{l}$ of aqueous solution with PEG 6000 maghemite nanoparticles was placed above the sensor surface. The mass of MNPs that remain after the solution evaporation was prior estimated of about $9.35 \mu\text{g}$, which means, in terms of magnetic moments using data from Fig. 6, about $2.1 \times 10^{-6} \text{ emu}$ at 10 Oe or $1.87 \times 10^{-5} \text{ emu}$ at 90 Oe.

a) H_{appl} is in the film plane, perpendicular on I_{sens}

The setup is presented in Fig. 3(a). We found that after biasing the sensor by a short pulse of 500 Oe and then turning off H_{bias} , the PHE sensor will keep a relatively well-ordered magnetic state.

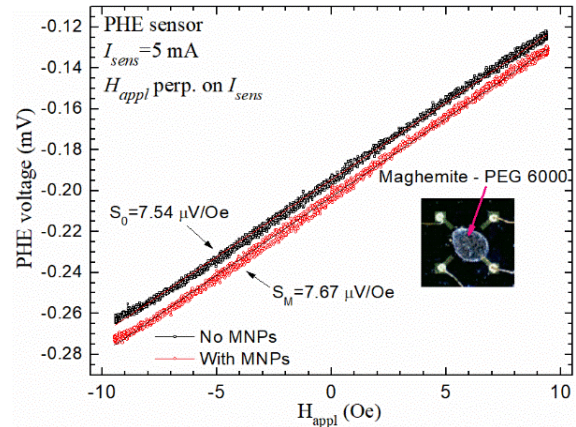


Fig. 7. The low field dependencies of the PHE signal, without and with MNPs of maghemite functionalized with PEG 6000 deposited and dried above the sensor surface; H_{appl} is directed like in Fig. 3(a). Prior each measurement, a short pulse of $H_{\text{bias}} = 500 \text{ Oe}$ has been applied along I_{sens} .

Fig. 7 presents the low field dependencies of the PHE signal measured on the PHE sensor, in the above biasing conditions, without and with MNPs placed on

the sensor surface, like we see in the inset. Prior to each measurement the sensor was biased. During our tests we observed that for applied fields higher than 20 Oe the initial magnetization state will be destroyed and the reversal processes will generate nonlinearity and hysteretic effects like in Fig. 4(a).

These results, which are in good agreement with [18], show a shift of the field dependence of the output signal when the MNPs are on the sensor surface. The output signal presents a very good linearity but, in some conditions where a temperature drift can occur, this detection method cannot offer a net and unambiguous signal to certify the presence of the MNPs. To get a larger signal from MNPs, using this setup, a larger field must be applied, see Fig. 6. However, for fields larger than 20-50 Oe applied in the film plane, the PHE sensor will saturate.

b) H_{appl} is in the film plane, parallel with I_{sens} .

Here we propose an improved detection method based on *localized reversal nucleation induced by MNPs in the sensing layer*. Until now, such method was studied for GMR sensors where micromagnetic simulations and experiments were made in order to increase the dynamic range of these sensors, in terms of MNPs detected [14, 19]. It was shown that micrometer sized structures, with large aspect ratio, have a limited dynamic range [19]. The sensing setup is based on the PHE presented in Fig. 1 and Fig. 3 where, this time, H_{appl} is directed along the driving current I_{sens} . A biasing field, H_{bias} , perpendicular to the sensor surface may be present. This field can enhance the magnetic moments of the MNPs with a very small perturbation of the PHE sensor. The choice of this setup is motivated by the relatively large detection area, typical for PHE sensors, large SNR and a very good thermal stability. In [14] were used CoFe MNPs which present a large magnetic moment that can easily induce reversal nucleation in the sensing layer due to the magnetostatic interaction between MNPs and the sensor. However, the maghemite NPs, often used to label biomolecules, have a lower magnetic moment at small fields. We show that PHE sensors employed in this study can easily detect very low magnetic moments of the maghemite nanoparticles functionalized with PEG 6000. The MNPs were deposited above the sensor surface like described above. Fig. 8 presents the field dependencies of the sensor signal when H_{appl} is swept between ± 30 Oe for $H_{appl} = 0$. The inset shows an image of the Helmholtz coils arrangement and the chip with PHE sensors.

The sharp peaks, symmetrically placed, correspond to coercive state of the PHE sensor without and in the presence of the MNPs of maghemite functionalized with PEG 6000. The magnetostatic coupling between the MNPs and the sensor surface is responsible for the coercive field shift and the change of the signal amplitude in the coercive state where the structure is more sensitive to any field perturbations, Fig. 8. The variation of the sensor signal, ΔU_S , due to the MNPs, increases with H_{bias} because their magnetic moments increase like is illustrated in Fig. 6. So, larger in plane stray fields will come from MNPs. Similar curves for

$H_{bias} = 45$ and 90 Oe were obtained. Fig. 9 presents the dependence of the measured ΔU_S on the biasing field applied perpendicular to the sensor surface for the same amount of MNPs.

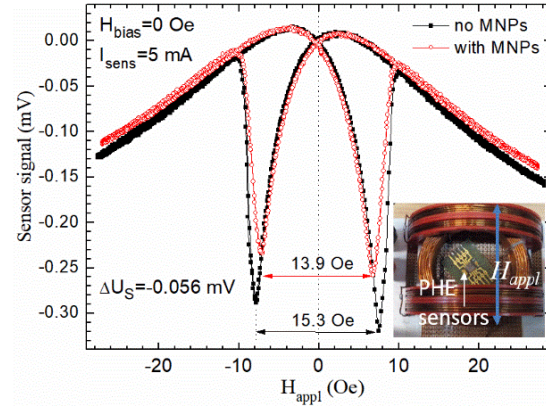


Fig. 8. The field dependencies of the sensor signal, without and with MNPs on his surface, when H_{appl} is swept between ± 30 Oe; $H_{bias} = 0$ Oe. The peaks correspond to the coercive state in the magnetic layer. The variation of the sensor signal, $|\Delta U_S| = 0.056$ mV.

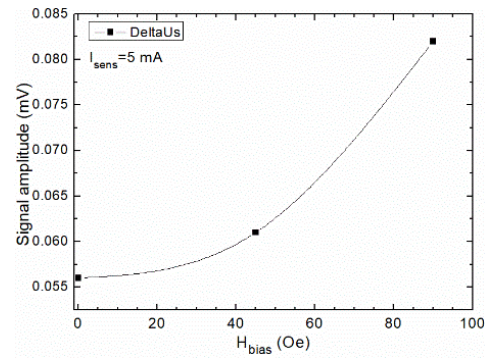


Fig. 9. The variation in absolute value of the sensor signal, ΔU_S , with H_{bias} .

It looks that it is convenient to use higher values of H_{bias} directed normal to the sensor surface to polarize the MNPs. However, for larger values of H_{bias} , the magnetic moments from the sensing layer start to rotate out of the plane in order to follow the field orientation and the extraordinary Hall effect becomes important. For lower values of H_{bias} , the PHE signal is less affected by such a behavior.

Having the above data, we can estimate that for $H_{bias} = 90$ Oe a signal variation $|\Delta U_S| = 0.082$ mV appears for a magnetic moment of 1.87×10^{-5} emu which means a sensitivity of about 0.22×10^{-3} emu/mV. So, it comes that magnetic moments of about 2.2×10^{-6} emu can be detected with this sensor for a signal variation of 0.01 mV. This value corresponds to a mass of about 1.1 μ g of maghemite nanoparticles functionalized with PEG 6000.

3. Conclusions

We microfabricated chips with 4 Permalloy based PHE sensors, 1 mm in diameter and 20 nm thick. Field

detection and angular encoder applications were presented. The field characteristics of the PHE sensors can be fine-tuned through an external field H_{bias} . Such that, for $H_{bias} > 25$ Oe, the output signal has a very good linearity and a field sensitivity up to $15.4 \mu\text{V/Oe}$ for a driving current $I_{sens} = 5$ mA. For $H_{bias} = 25$ Oe the output signal presents a non-linear field behavior but with an average sensitivity of about $30 \mu\text{V/Oe}$. The studied disk-shaped PHE sensors show a very good behavior of the output signal for rotating magnetic fields larger than 150 Oe. Because there is no shape anisotropy, the magnetization follows accurately the magnetic field orientation and the output signal can be very well fitted with a $\sin 2\theta$ curve. Micromagnetic simulations were performed in order to have a better view of this behavior. Finally, we have demonstrated the ability of these sensors to detect MNPs of Maghemite functionalized with PEG 6000 as a starting point to build a LOC biodetection device. We used a detection method based on localized reversal nucleation induced by MNPs in the sensing layer. A sensitivity of about $0.22 \times 10^{-3} \text{ emu/mV}$ was found for a biasing field of 90 Oe. This means that magnetic moments of about 10^{-6} emu can be detected which means in terms of mass $1.1 \mu\text{g}$ of Maghemite functionalized with PEG 6000. The detection limit can be further enhanced by improving the signal processing taking into account not only the signal amplitude but also the phase through the position of the coercive state due to the presence of the MNPs over the sensor surface, Fig. 8. Also, the use of measurement sensor – reference sensor pairs will improve the thermal stability and immunity to unwanted external fields. The using of a tilted biasing field, at small angles to the normal, can have double effect: can allow structure biasing to a convenient state and, also, can polarize the MNPs. Further developments and improvements will be done by integrating these sensors in a microfluidic system.

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(048)

Comparison of a Low-cost System Based on Electrochemical and Optical Sensors for Air Quality Monitoring against Reference Methods

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Summary: In this communication, a comparison between a home-developed portable, low-cost system and reference methods for air quality measurement is performed. For the measurement of the main pollutants, both gaseous (NO₂, NO, O₃, CO) and aerosol (PM₁₀ and PM_{2.5}), electrochemical and optical sensors are used, respectively, all from Alphasense. The device also features the measurement of meteorological data such as temperature and humidity. In addition, different communication modules (Bluetooth and Ethernet, among others) and storage of acquired data are included. The control core of the system is a low power microcontroller from the manufacturer ST (STM32L476), and the whole device is powered through a 230 V AC socket. Air sampling is done actively, and therefore the equipment includes a pump. The reference equipment used for validation belongs to the air quality protection and research network of Extremadura. The complete system has been tested in real conditions for two months next to the fixed reference stations located in Badajoz and Plasencia (Spain). Its general operation has been successful, with strong correlation results, although the calibration of the electrochemical sensors should be modified in future work to improve its performance, especially in low polluted environments.

Keywords: Air quality monitoring, Electrochemical sensors, Particulate matter, Low-cost.

1. Introduction

Atmospheric pollutants are currently a major focus of attention worldwide due to the significant damage they cause. For this reason, the availability of reliable and widely distributed monitoring equipment is of great importance. The analytical methodologies applied in standard air quality monitoring networks guarantee the reliability and continuous availability of results, to comply with environmental protection aims. However, this objective is achieved at a high cost, through sophisticated and costly instrumentation. Automatic analyzers and sampling systems must be installed in large, air-conditioned booths at carefully selected representative locations, provided with high-power electrical connection. Samples taken to laboratories are processed using standardized, high-cost instrumental techniques, operated by highly qualified personnel. As an alternative solution to this drawbacks, a low-cost monitoring device generation is becoming increasingly important [1]. These devices are not intended to replace reference instruments, especially for application purposes, but rather to be a complementary source of information on air quality [2].

2. Materials and Methods

This device includes four electrochemical gas sensors to measure CO, NO₂, NO and O₃, and an optical particle counter sensor to measure PM₁₀ and PM_{2.5}, all of them regulated atmospheric pollutants. The Analogue Front End supplied by Alphasense has

been used for connecting the sensors to the system. An external 24-bit digital-analog converter with 8 inputs is used to read the value of the sensors. On the other hand, the PM sensor used is the OPC-N3 model, also from Alphasense, which also provides humidity and temperature information. However, another humidity and temperature sensor has been integrated, as the previous one may not represent the actual environmental conditions. The control of the system is based on the use of an STM32L476 Nucleo-Board, an ultra-low-power microcontroller based on a high-performance 32-bit core operating at a frequency of up to 80 MHz. On the other hand, it incorporates modules for Low Energy Bluetooth and Ethernet communication in order to receive data on smart mobile devices or data clouds. Moreover, these data from the sensors can also be stored on a microSD card attached to the prototype. Finally, the design includes the possibility of external connection of LoRA or GSM communication modules for wireless data transmission. Fig. 1 shows the designed electronic prototype and points out some of its main elements. The dimensions of the designed electronic printed circuit board are 90 mm × 110 mm.

In order to simplify the connection of the system in different locations, a 230 V AC to 5 V converter and a mains plug have been included. Moreover, the system is equipped with an electric pump that allows the air to be conducted from the outside to the sensor cell. In this way, by maintaining a constant flow of the sample through the sensor cell, interference related to pressure changes will be reduced. This factor is very important in the use of electrochemical sensors. The sensor cell is made of metal in order to reduce the effects of

electromagnetic noise. The entire system has been encapsulated in a watertight box for outdoor placement. A picture of the complete system, detailing the main components, is shown in Fig. 2.

3. Results

The operation of the developed system was successfully tested for two months (August and September 2019) in parallel with the fixed reference stations of Badajoz and Plasencia, in Spain. The data sampling period was 30 seconds, although it was subsequently averaged every 10 minutes for graphical representation and comparison with data from the reference station in $\mu\text{g}/\text{m}^3$. Real-time data was stored in the microSD included in the system.

The preliminary test of the behavior of the sensors presented in this communication was based on visual analysis of the time series of the polluting gases NO_2 , CO and O_3 , and PM_{10} . The O_3 graph is plotted as an example in Fig. 3.

The operation of the ozone sensor seems to be correct, as the signal follows the reference signal quite closely, although its performance could be improved with more accurate temperature and humidity compensation methods. In the case of NO_2 and CO, the results are less accurate, probably because the concentration of these pollutants in Badajoz are quite low, close or below the detection limit of the sensing systems.

Fig. 4 shows, as an example, the comparison between reference and the prototype measurements in PM_{10} .

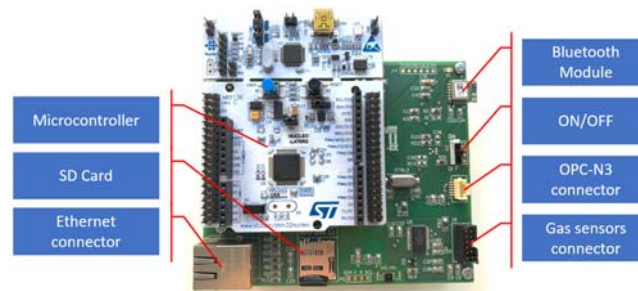


Fig. 1. Electronic system designed.

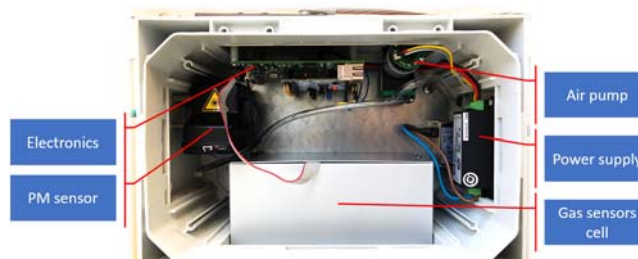


Fig. 2. Complete system developed.

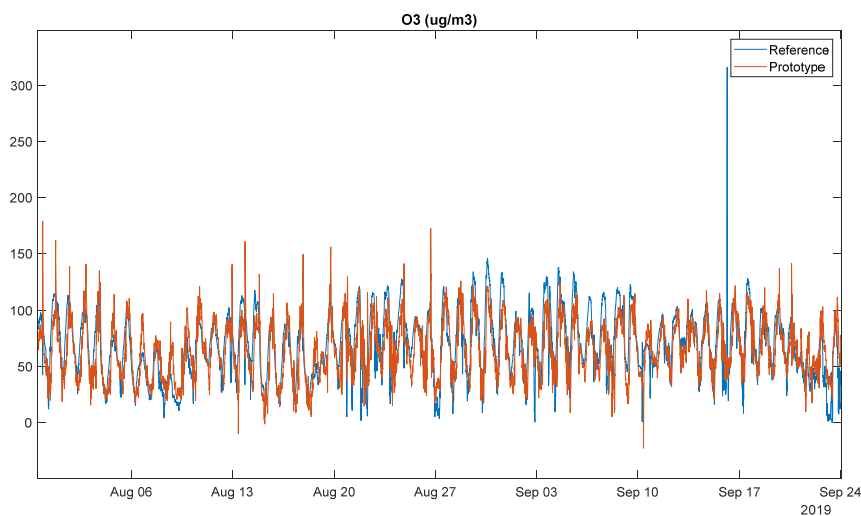


Fig. 3. O_3 time series. Blue: reference, orange: prototype.

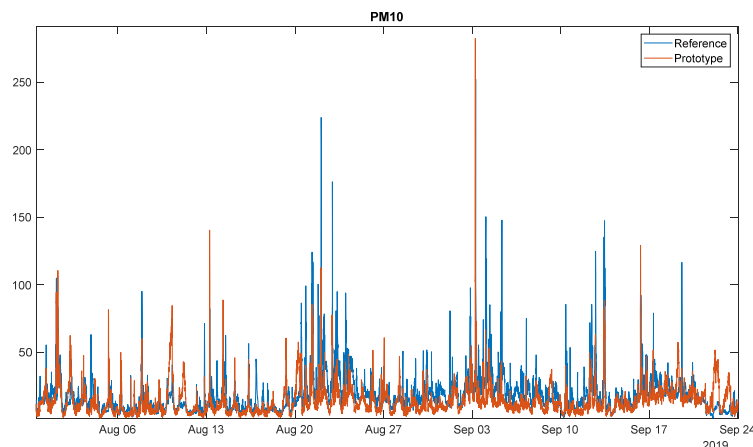


Fig. 4. PM₁₀ time series. Blue: reference, orange: prototype.

4. Conclusions

A portable and low-cost system has been designed for the measurement of the main air pollutants by electrochemical (gaseous pollutants) and optical (PM) sensors. The device was tested for 2 months by comparing its results with those given by a reference station, to study its operation. The system worked correctly during the field campaign, demonstrating that the electronics and design of the device are suitable for atmospheric pollution monitoring under real conditions.

Promising results were obtained for some compounds and PM, whereas other (NO_x) gave a weaker performance, probably due to the low concentration of these pollutants in Badajoz. Calibration refinement and additional validation will

also require new, longer measurement campaigns in more polluted locations.

Acknowledgements

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Dengue Virus Serotype Detection Using Cell-free System Based Biosensor

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Summary: Infection by dengue virus can cause mild fevers or, in some cases, the lethal dengue hemorrhagic fever / dengue shock syndrome. The cause for these lethal symptoms are still being studied but evidence suggest that secondary infection with dengue virus serotype 2 or 3 will increase the chance of developing these symptoms. Thus by defining the infected serotype in patients can help determine the risk in developing fatal symptoms. In this study, we utilize RNA toehold switches in cell-free reactions for detection of dengue virus serotypes. Our result shows that the detection ranges from 50-500 nM of dengue virus target molecules, when the samples were directly add to the reactions. In addition, these reactions were integrated into a paper-based-chip biosensor for dengue serotype discrimination for potential point-of-care applications.

Keywords: Dengue virus serotype detection, Dengue virus biosensor, Paper-based biosensor, Cell-free system, RNA toehold switch.

1. Introduction

Dengue disease is a major health problem in tropical and subtropical regions. The cause of this disease is the dengue virus (DENV) which has four different circulating serotypes (DEN1-4). Patients infected with the virus can experience symptoms ranging from mild fevers (undifferentiated fever or dengue fever) to life-threatening syndromes (dengue hemorrhagic fever / dengue shock syndrome or expanded dengue syndrome). It has not been clearly identified on what factors are causing the lethal symptoms. But there have been reports indicating that the factors contributing to dengue severe symptoms can involve secondary infection with DEN2 or DEN3 [1]. Hence, precise discrimination of the dengue virus serotypes is important for predicting the risk of fatal DENV diseases.

RNA toehold switches, transcribed in cell-free reactions, have been shown to successfully detect RNA viruses [2, 3]. These RNA structures are RNA sequences designed to form a stem loop structure containing a ribosome binding site (rbs) within the loop. Specific sequence can bind to the toehold structure causing the stem-loop to open and exposing the rbs and allow translation of the reporter gene downstream [4]. This opens the opportunity for applying this system into a biosensor device.

In this work we employ the use of RNA toehold switches in cell-free system, designed for specifically detecting different DENV serotypes. These reactions were then incorporated into a fluidic chip for potential prototyping for a biosensor device.

2. Methods

2.1. Preparation of Cell-free System

The cell-free reactions were prepared following Pardee *et al.*, 2014 with some minor modifications. Briefly, the PURExpress® In Vitro Protein Synthesis Kit were mixed according the manufacture protocol. Plasmids containing sequences encoding for the desired RNA toehold switch with lacZ downstream as the reporter gene was added to a final concentration of 30 ng/μl [3]. Four types of plasmids were used altogether in this study, in separate reactions. Each of the plasmids contain sequences encoding for RNA toehold switch specific towards one of the DENV serotype (DEN1 to 4). Chlorophenol red-β-D-galactopyranoside (CPRG) was used in the reactions as the indicator, at a final concentration of 0.6 μg/μl [3]. The RNase inhibitor Murine (NEB, Cat. No. M0314S) was also added to a final centration of 0.4 Unit/μl. After reaction preparation, 1.5 μl of each reactions were spotted onto a Whatman No.1 Chr paper printed with 2 % Alkyl Ketene Dimer (AKD) as hydrophobic patterns to retain the reactions. The paper area containing the reactions were blocked with 5 % BSA prior applying the reactions [3]. Then the reactions were flash freeze in liquid nitrogen and freeze dried for 2 hours.

2.2. Range of Detection

Four types of 30 bp single strand DNA (ssDNA) were design to have the same sequence of the DENV serotypes 1 to 4. Therefore these ssDNA can bind to the four types of RNA toehold switch, corresponding

to the dengue serotypes. These ssDNA were designated as trigger DNA. To determine the range of detection of the RNA toehold switches in the cell-free reactions, all four trigger DNA were added to the reactions that can produce its corresponding RNA toehold switch. The trigger DNA were added at final concentrations of 0, 30, 50, 500, and 5000 nM. Triplicate reactions were prepared for each trigger concentration. The reactions were then incubate at 37 °C for 3 hours.

2.3. Fabrication of DENV Serotype Biosensor Prototype

The four types of cell-free reactions that transcribe the RNA switches for DENV serotypes were integrated into a fluidic chip platform to demonstrate the potential for a biosensor prototype. The fluidic chip was fabricated from two layers of acrylic, the top cover and the channel. The channel allows the sample to flow over all four reactions and into the absorption pad at the end. While the reactions are retained in filter papers printed in AKD.

3. Result

3.1. Detection Range of DENV RNA Toehold Switch

When applying a range of known concentrations of trigger DNA directly to into the corresponding cell-free reactions, the result shows the detection range of the reactions, after incubating at 37 °C for 3 hours (Fig. 1). The concentration containing enough trigger to open the RNA toehold switch will cause the translation of LacZ. The LacZ digests the CPRG changing the reaction color from yellow to purple. Our result shows that the detection range of the RNA toehold switches are 50 nM of trigger DNA (DEN1) and 500 nM trigger DNA (DEN2 to 4) (Fig. 1).

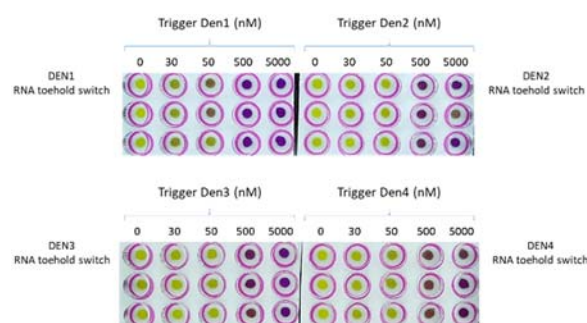


Fig. 1. Detection range of RNA toehold switches for DENV serotype 1 to 4.

3.2. DENV Serotype Biosensor Prototype

In order to test the potential of the biosensor prototype, 1 μ M of trigger DNA for Den1 to Den4 was applied to four separate biosensor. After incubating at 37 °C for 3 hours, the result shows that all of the trigger

DNA were able to react with the correct corresponding RNA toehold switch in the cell-free reaction on the biosensor prototypes (Fig. 2). It is also worth noting that the trigger DNA did not react with other RNA toehold switches that is not its corresponding switch. This demonstrates the specificity of the RNA toehold switches in the reactions (Fig. 2).

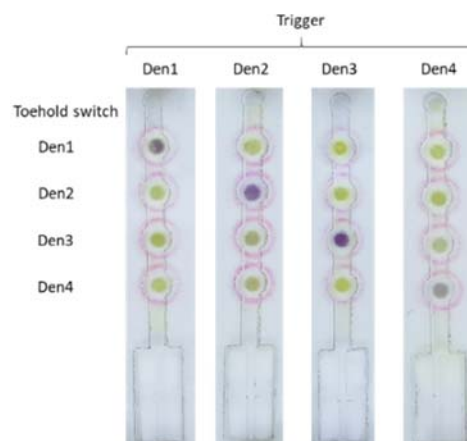


Fig. 2. The prototype for DENV serotype biosensor.

4. Conclusions

Our work demonstrated the possibility of utilizing the RNA toehold switch transcribed in cell-free reactions in a biosensor device for discriminating DENV serotypes.

Acknowledgements

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(051)

Estimation of Attenuation Coefficient from Simulated B-mode Ultrasound Images and Tissue Mimicking Materials

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Summary: We report on the estimation of the attenuation coefficient from simulated B-mode ultrasound breast images. It is well documented that this parameter significantly diverges between normal tissue and malignant lesions. Thus it is of great importance to devise and test algorithms that facilitate accurate estimation of spatially resolved attenuation. *k*-Wave is a Matlab toolbox for the time-domain simulation of acoustic wave fields that facilitates both linear and nonlinear wave propagation in homogeneous and heterogeneous tissue, as compared to strictly linear ultrasound simulation tools (e.g. Field II). *k*-Wave can also be applied to simulate arbitrary distributions (resolved down to single voxel sizes) of parameters including the speed of sound, mass density, scattering strength and to include power law acoustic absorption necessary for simulation tasks in medical diagnostic ultrasound. Via the spectral-log-difference technique, we obtain estimation results from simulated and thus well-known absorption distributions and compare the results to reference values.

Keywords: Ultrasound imaging, Attenuation imaging, *k*-Wave, Ultrasound simulation, Spectral log difference technique, Tissue mimicking materials.

1. Introduction

In addition to use medical ultrasound (US) to image tissue morphology, the characterization of benign and malignant masses based on their mechanical/acoustical properties has been a long term goal [1]. An important property is the absorption parameter α (in units of dB/cm), which follows a frequency power law of the form [2]

$$\alpha = \alpha_0 \cdot f^\gamma, \quad (1)$$

where α is the absorption coefficient, α_0 is the power law prefactor in dB/((MHz) ^{γ}), and γ is the power law exponent in medical imaging typically assumed as $\gamma = 1.0$. Further parameters of interest that distinguish tissue types are the mass density, ρ , and the speed of sound, c . US absorption, as characterized by Eq. (1), for soft biological tissues ranges from $\alpha_0 = 0.03$ dB/(MHz cm) for water up to a maximum of $\alpha_0 = 4.0$ dB/(MHz cm) for certain lesions.

Testing of algorithms for estimating absorption parameters from acquired B-mode images, an example of which is shown in Fig. 3 requires a well characterized set of training data that can be obtained by either tissue mimicking phantoms, exhibiting fairly homogeneous properties that can both be directly measured using the transmission mode or the backscatter mode as in imaging US, respectively.

A US setup as shown in Fig. 1 is used for transmission mode measurements. After a pulse generator excites the US transducer and a needle hydrophone measures the sound pressure level with and without the phantom in place, the absorption for a

given thickness can then be directly determined using the spectral-log-difference technique (SLD) [3].

The setup in Fig. 1 was used to estimate the attenuation coefficient α for different aluminum oxide concentrations (grain sizes 0.3 μ m and 3 μ m) in tissue mimicking phantoms using the SLD method [4] detailed below (for measurement results obtained see Fig. 2).

Another less involved way to obtain calibration data for testing α -estimators is using simulation software to solve the US wave equation including source terms, absorption and dispersion relations and also the nonlinear pressure-density relation. An open source time-domain simulation software package developed for this purpose is *k*-Wave [5, 6].

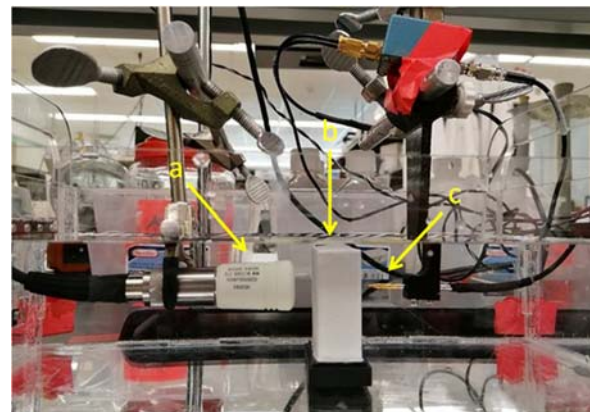


Fig. 1. Setup used for the measuring of tissue mimicking phantoms. (a) Piston-type US transducer; (b) Phantom; (c) Needle hydrophone.

2. Estimation Algorithm

The estimation algorithm is designed to not only estimate a spatially averaged attenuation of the propagating wave but does so with the aim of attaining a very high spatial resolution, too. This spatial resolution is necessary to being able to detect very small and small lesions in the clinical screening process. Fig. 3 shows a cropped B-mode scan exhibiting the typical US speckles that, as will be shown later, are detrimental in the estimation process. Furthermore the distal and proximal data windows are indicated over which the estimation algorithm operates. The sketch in Fig. 4 defines all the indices used in the algorithm. Index $1 \leq \ell \leq L$ is indicating the lateral sector scan line (typically going up to and beyond $L = 200$), and index $0 \leq n < N - 1$ is indicating the depth dimension (typically being sampled at and beyond an $f_s = 200$ MHz sampling rate, so that given the speed of sound, $c \sim 1500$ m/s, on the order of $N = 10^4$ samples are acquired for each scan line for a typical 5 cm depth range).

In order to estimate α_0 over a region of interest (ROI) set by range-gated proximal and distal windows within a B-mode scan of a breast, the same time-gain-compensation (TGC), and the same power settings are used to obtain backscattered signals from different depths over a lateral range of scan lines $\ell_l \leq \ell < \ell_r$ (see Fig. 4) to be averaged over later on.

Define $S_w(f_k)$ the power spectral density of the appropriately time-windowed¹ (subscript $(\dots)_w$) sequence $s_w(n)$ evaluated at frequencies $f_s k/M$; $0 \leq k \leq (M-1)$, with k , the frequency index, f_s the sampling frequency in Hz, and $f_s k/M$; $0 \leq k \leq (M-1)$, the length of the time windowed sequence.

$$\begin{aligned} S_w(f_k) &= S_w\left(\frac{f_s k}{M}\right) = \\ &= \frac{M}{2\pi f_s} \cdot |\mathcal{F}\{s_w(m)\}|^2 = \\ &= \frac{M}{2\pi f_s} \left| \frac{1}{M} \cdot \sum_{m=0}^{M-1} s_w(m) \cdot e^{-j\frac{2\pi}{M}mk} \right|^2 \end{aligned} \quad (2)$$

Now define $B(\ell, n)$ the transducer's output signal representing the TGC-compensated B-mode image with ℓ being the scan line index of the sector scan, and n being the time index representing the depth dimension. Depth $d(n) = n \cdot c/(2f_s)$ (assuming a constant over depth speed of sound c).

Both windows are limited laterally by $\ell_l \leq \ell \leq \ell_r$, thus in total $L_e = \ell_r - \ell_l + 1$ scan lines are averaged over (denoted by $\langle \dots \rangle$) in the following spectral estimation process. The proximal window extends between $n_{p1} \leq n < n_{pm}$, and the distal window

between $n_{d1} \leq n < n_{dm}$ samples. Using above definitions the power spectral density of a depth-windowed region (mean depth for the proximal window d_p , and for the distal window d_d) in a statistically homogeneous tissue, varying only in the attenuation coefficient, and averaged over all $L_e = \ell_r - \ell_l + 1$ selected scan lines is given by (subscripts p and d indicate the proximal and distal ROIs).

$$\begin{aligned} \langle S_p(f, d_p) \rangle &= \\ &= P(f)D(f)A(f, d)B(f, d)e^{-4\alpha(f)d_p}, \end{aligned}$$

for the proximal window, and

$$\begin{aligned} \langle S_d(f, d_d) \rangle &= \\ &= P(f)D(f)A(f, d)B(f, d)e^{-4\alpha(f)d_d}, \end{aligned}$$

for the distal window, where $P(f)$ represents spectrally the combined effect of the transmit pulse and the sensitivity (electro-acoustic and acousto-electric transfer functions) of the transducer. $D(f) = D(f, d)$ denotes the effects of focusing and diffraction that are related to the transducer geometry and its numerical aperture NA at average depth d . $A(f, d)$ is the cumulative attenuation along the propagation path from the surface of the transducer to depth d of the respective depth window which corresponds to the centers d_p and d_d , respectively, of the ROIs. $\alpha(f)$ is the attenuation coefficient within the considered ROIs. $B(f) = B(f, d)$ is a result of the scattering properties of the tissue within the ROIs, namely the effective scatterer size, the scatterer number density, and the mean square variation in acoustic impedance between the scatterers and the background.

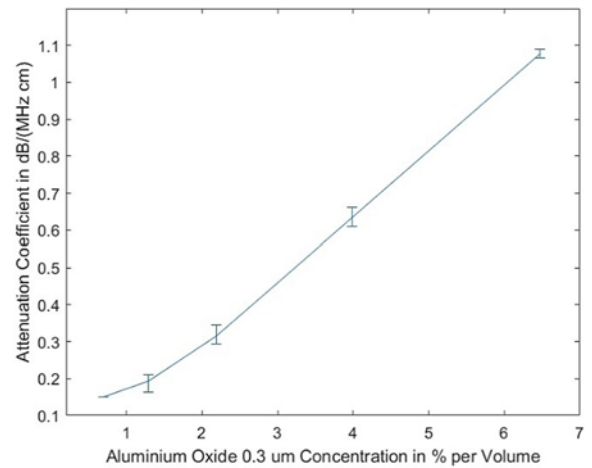


Fig. 2. Attenuation coefficient vs. concentration of aluminum oxide (grain size 0.3 μm) for homogeneously prepared tissue mimicking phantoms (as shown in Fig. 1 (b)).

¹ Please note, this window or tapering function used to suppress spectral side lobes is to be distinguished from the

above mentioned depth windows that allow to select spectra from different depths.

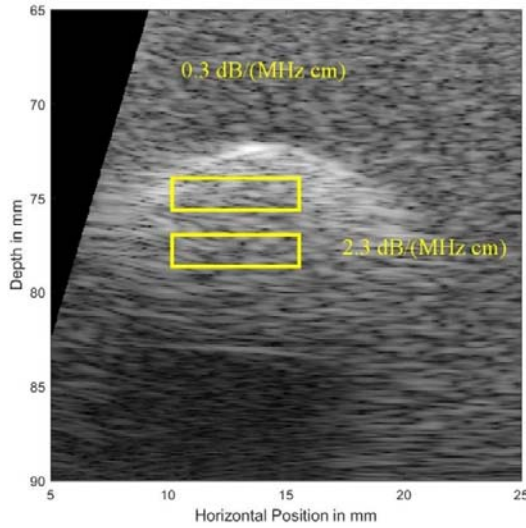


Fig. 3. Simulated US B-mode image, exhibiting typical US speckles and shadowing behind a strongly absorbing region (a simulated lesion). Distal (bottom) and proximal (top) windows are indicated.

Dividing the power spectra obtained from different average depths d_p and d_d and assuming that the tissue within these depth ranges is homogeneous and isotropic, i.e., the scattering term $B(f)$ does not vary with depth, yields:

$$Y(f) = \frac{S_p(f)}{S_d(f)} = \frac{A(f, d_p)B(f, d_p)}{A(f, d_d)B(f, d_d)} \cdot e^{4(d_d - d_p)\alpha(f)}$$

Taking the logarithm (to the base 10 to express the spectra in dB/Hz, and thus the attenuation coefficient in dB/(MHz cm)) yields:

$$\begin{aligned} \log(Y(f)) &= \frac{1}{20} \ln(10) \cdot 4(d_d - d_p)\alpha(f) \cdot \frac{A(f, d_p)B(f, d_p)}{A(f, d_d)B(f, d_d)}, \\ &\approx \frac{1}{20} \ln(10) \cdot 4(d_d - d_p)\alpha(f) = \\ &= \frac{1}{10} \ln(10) \cdot 2(d_d - d_p)\alpha(f) \end{aligned} \quad (3)$$

In Eq. (3) the factor 4 stems from the fact that the attenuation ratio as measured in Np is referring to the attenuation of field quantities (as opposed to dB which gauges power quantities) combined with the round-trip propagation that effectively doubles the path lengths.

The attenuation coefficient of the sample (in dB/(MHz cm)) can be estimated at each frequency component f_k (see Eq. (2)) by calculating γ , the slope of the straight line that fits the log-ratio of the two spectra $Y(f)$, i.e., the slope of the straight line that fits Eq. (3) 3 versus depth.

If the attenuation as measured in dB is assumed to increase linearly with frequency, then the attenuation coefficient's slope is used as a measure for the attenuation in the tissue of interest. The attenuation coefficient can then be simplified to

$$\alpha(f) = \alpha_0 \cdot f, \quad (4)$$

where the parameter α_0 is the attenuation coefficient's slope in dB/(MHz cm) that can be estimated by finding the slope that fits Eq. (4).

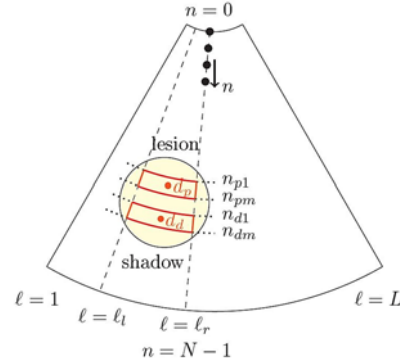


Fig. 4. To define locations and indices within a sector scan used in the estimation algorithm. At the top the US transducer is assumed operating as a phased array.

In the following section the estimation results obtained are shown for a k -Wave simulated and thus very well defined ultrasound phantom.

3. Estimation Results

In order to check the validity of the estimation results as obtained by applying the SLD method the k -Wave environment was used to simulate wave propagation in various media. Initially a homogeneous and isotropic medium was simulated that included only five discretely placed scatterers (at depths of $d = \{5, 10, 15, 20, 25\}$ mm) and α_0 was set at 2.3 dB/(MHz cm) throughout the volume. The transducer (center frequency $f = 5$ MHz) was focused both in azimuth and elevation at 35 mm in depth.

A video sequence was derived clearly showing the wave propagating down range, being scattered and re-scattered off all scatterers and also being strongly attenuated over the distance it propagated. A single frame out of the sequence is displayed in Fig. 5. It clearly indicates the scattered and re-scattered circularly propagating waves emanating from the locations of the scatterers.

Fig. 6 shows the sound pressure amplitude vs. depth in blue and the expected decrease in amplitude (according to the selected medium parameters of 2.3 dB/(MHz cm)) in red. The deviation between the two graphs is attributed to the focusing action of the transducer which is set to focus at a depth of 35 mm in both lateral and azimuthal directions resulting in a less pronounced reduction of sound pressure level than expected.

Clearly visible are the locations of the five scatterers placed at depths of $d = \{5, 10, 15, 20, 25\}$ mm.

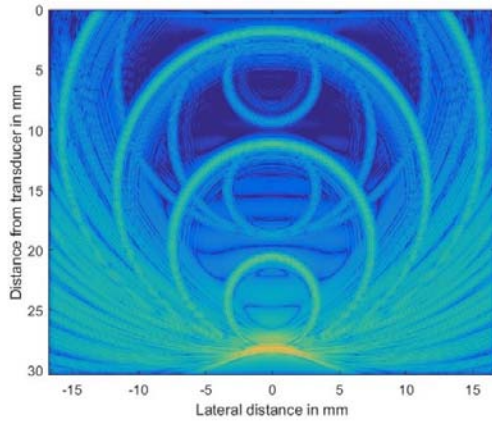


Fig. 5. Ultrasonic wave's envelope propagating downwards, focused at 35 mm and being scattered off discretely placed scatterers. A logarithmic compression is used for display to cope with the large dynamic range of approximately 100 dB, which is caused by the attenuation coefficient of $\alpha = 2.3$ dB/(MHz cm).

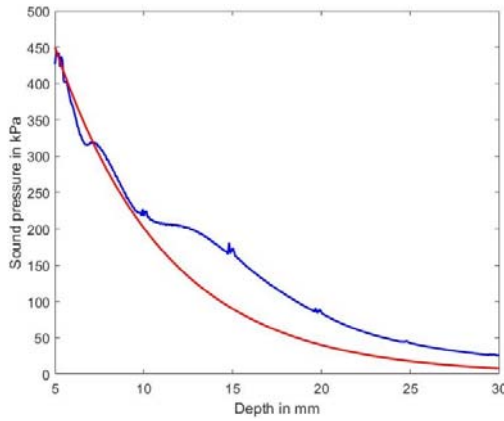


Fig. 6. Decrease of sound pressure amplitude vs. depth in blue and the expected decrease in sound pressure for the assumed attenuation coefficient of 2.3 dB/(MHz cm) in red.

Fig. 7 shows in blue the sound pressure level evaluated along the central scan line of Fig. 3 and $n = L/2$ in Fig. 4, respectively, and in red the expected decrease due to the chosen attenuation coefficient of $\alpha_0 = 2.3$ dB/(MHz cm). Simulation artifacts at very low sound pressure values are seen to increase the noise floor beyond 17 mm depths and those artifacts seem to also cause the less than expected slope towards greater depths.

In a further step a much more general, highly structured medium similar to real breast tissue also including a lesion was simulated (B-mode simulation shown in Fig. 2). Here the typical US speckles are acting as multiplicative noise influencing strongly the SLD-estimate. To obtain valid results the spatial averaging needs to extend way beyond the average size of a speckle. This speckle size is greatly influenced by the numerical aperture of the US transducer, set here to $NA = f/2$, with the focus again set to 35 mm in both azimuth and elevation.

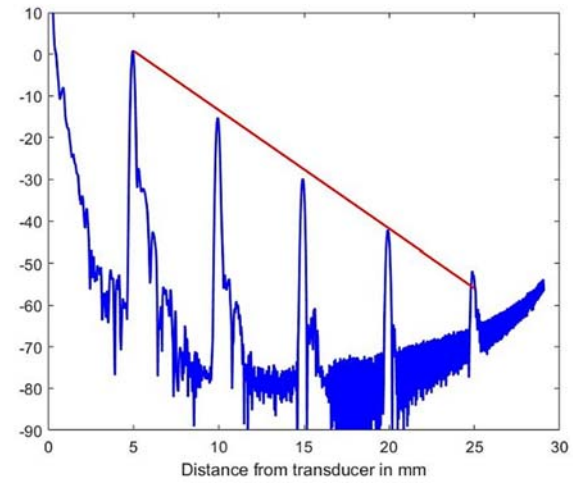


Fig. 7. Decrease of sound pressure level vs. depth estimated from the B-mode image. Indicated in red is the expected decrease in amplitude vs. depth (set to 2.3 dB/(MHz cm)).

Fig. 8 shows the sound wave field at an intermediate time step of a highly structured medium containing random scatterers of high number density evenly distributed over the volume. The only parameter that is varying over space is the attenuation coefficient. It is set evenly at $\alpha_0 = 0.3$ dB/(MHz cm) except for a cylindrical region of diameter 8 mm centered on the symmetry axis at a depth of 22 mm and which has a chosen $\alpha_0 = 2.3$ dB/(MHz cm).

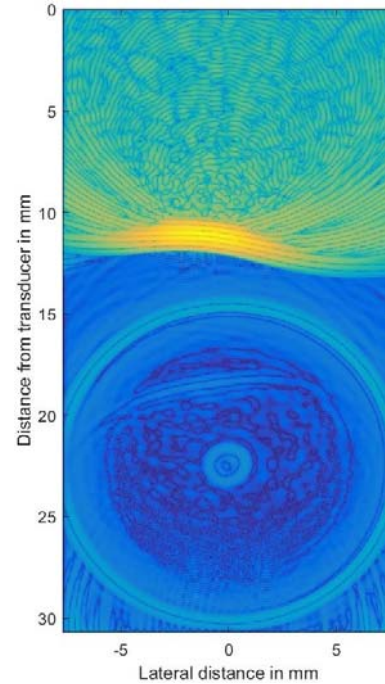


Fig. 8. Sound pressure field at an intermediate time step showing the propagating wave through a highly structured medium that has a strongly attenuating region (diameter 8 mm) centered at a depth of 22 mm.

One can observe US speckles (at depths of less than 12 mm) and the propagating wave being focussed at a scan angle of -5° at 25 mm in depth.

Fig. 9 shows a single frame out of a video sequence displaying on the left the averaged power spectral density $\langle S_p(f, d_p) \rangle$ of the proximal window used as reference in blue, and the power spectral density $\langle S_d(f, d_d) \rangle$ of the distal window in red, and on the right the SLD term $Y(f)$ in dB in blue, and its average over the considered frequency range in red. The estimate for the chosen depth is close to the true value of $\alpha_0 = 2.30$ dB/(MHz cm).

Fig. 10 shows one intermediate processing step where the sound pressure level vs. depth is indicated.

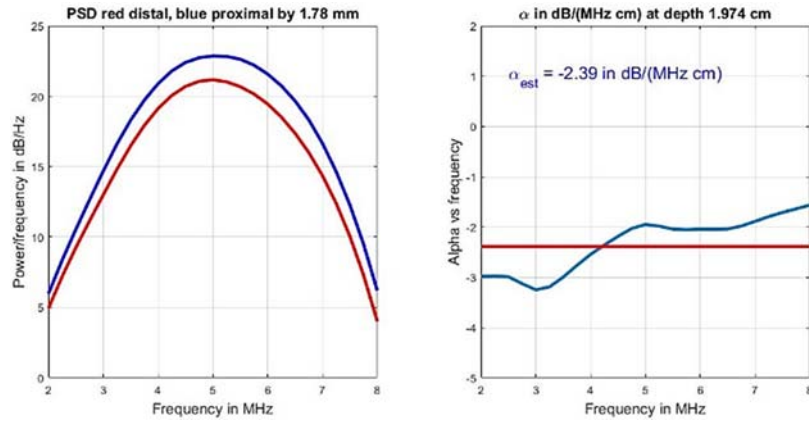


Fig. 9. Intermediate steps of the estimation process. Left: power spectral densities of the proximal window used as reference (in blue), and power spectral density of the distal window (in red). Right: SLD-estimate vs. frequency (blue) and averaged over the frequency range of interest in red.

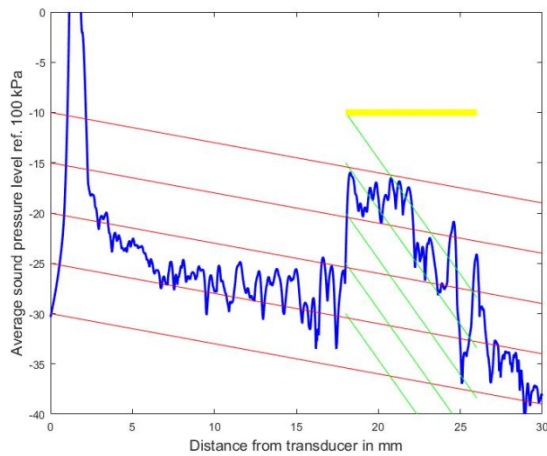


Fig. 10. Averaged sound pressure level in dB (ref. 3 kPa). The red lines indicate the simulated alpha-value of $\alpha = 0.3$ dB/(MHz cm), the green lines indicate the increased attenuation (2.3 dB/(MHz cm)) of an inclusion centered at $d = 22$ mm (radius 4 mm) as indicated by the yellow bar.

ROI needs to be chosen necessary for the detection of very small lesions.

4. Conclusions

We showed that the spectral-log-difference method is able to accurately estimate spatially resolved the US absorption parameter. The open source US simulation

One can note the strong variance of the sound pressure vs. depth that is caused by the effect US speckles have on the SLD estimates. Indicated as a ruler in red is the expected slope of the sound pressure level (assuming $\alpha_0 = 0.3$ dB/(MHz cm)) in the overall volume, and in green the expected slope within the highly attenuating region ($\alpha_0 = 2.3$ dB/(MHz cm)), the yellow bar indicates the area of strong attenuation, the lesion. From this diagram one can conclude that a rather strong variance of the estimator needs to be accepted, in particular, if a small

software *k*-Wave was used in the process of evaluating the estimation results and to compare to measurements on tissue mimicking phantoms.

Acknowledgement

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(053)

Quick Fabrication of Integrated Sensor Interface by Inkjet Printing of Logic Gates and Graphene Inks

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Summary: This paper reports the fabrication of flexible logic gates using printed field effect transistors made by Graphene as semiconductor and honey as gate material. PMOS transistors were made with Graphene and then analyzed to find the voltage characteristics and best operation mode. The PMOS transistors showed a mobility of $120 \text{ cm}^2\text{s}^{-1}\text{V}^{-1}$ and a good slew rate. Details of design and fabrication were studied and reported. The PMOS circuits included Inverters, AND logic, and OR logic. The logic gate performance was compatible to the circuit simulation observed by transistor models. These new logic gates have applications in biomedical interfacing and wearable electronics.

Keywords: Flexible electronics, Graphene, Printed electronics.

1. Introduction

Flexible electronics have a high demand and are among the fastest growing products in the past few years and it is expected that the global market for such products grow from a few billion to hundreds of billions of dollars in the years to come [1]. A flexible circuit is considered product of electronic fabrication that can be bent, stretched, compressed, or undergo physical deformation during its operation [2]. It has numerous applications from foldable cellphones, flexible touch screen displays, wearable electronics to medical devices, biomedical engineering and military applications [3, 4]. With continuous advances in the field of nanotechnology and flexible electronics, there is a need to find materials with flexibility to make active components and digital circuits. This paper will report on application of Graphene ink and inkjet printing to fabricate logic gates using thin-film transistors (TFT). Logic gates are among the basic active components in flexible electronics and fabricating them on a polyamide substrate by a drop-on-demand technique, helps improving flexible circuits.

In this study logic gates are fabricated using printed TFTs made by graphene substrate and honey as gate. Since flexible gates such as dielectric gels showed a good performance under strain and bending, we decided to use honey as the gate material.

Despite the fact that the gel-based logic gates such as honey are primarily considered as components of future flexible and wearable electronic systems, their biomedical interfacing and sensing applications are promising for practical circuits and systems. Interfacing of logic gates with modified electrodes directly connected to the TFT gates will provide an input directly applied to semiconductor devices and helps reduce the noise, power consumption, and delay time in sensing systems. The wide range of such interfacing applications has been illustrated in Fig. 1.

To make a logical AND gate, two TFTs were connected to -15 V in parallel with output Y separated from ground by a resistor.

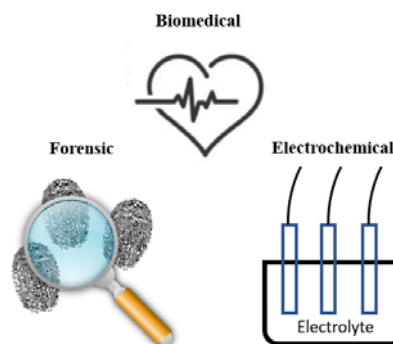


Fig. 1. Different sensing applications of logic gates made by honey as a gel-based TFT gate are illustrated in this picture.

2. Design

Logic 0 for inputs A and B were represented by -15 V, while 0 V represented a logic 1. In our experiments the output values were expected to show a perfect 0V or -15 V because of not using a complementary logic design. The circuit model is illustrated in Fig. 2.

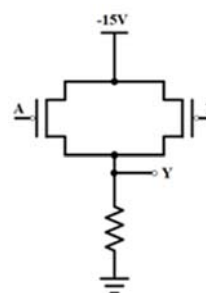


Fig. 2. This picture illustrates a PMOS – AND gate to be created using graphene based TFT.

3. Fabrication

Honey was used as a good substitute for dielectric gates in prototyping of Graphene based transistors [5]. To make logic gates we printed Graphene ink on polyamide substrate. High temperature Kapton tape was used to have both flexible and a high thermal tolerance. After printing the sample was heated for 30 minutes in a 230-degree Celsius oven to improve the conductivity of Graphene. The sample showed a 30 k Ω /square in resistivity. To make a stable connection to the probes for the device under the test two conductive copper tapes were attached to the Drain and Source of the transistors as shown in Fig. 3.

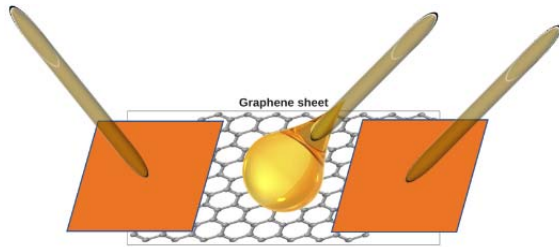


Fig. 3. This picture shows a model honey gate material on the Graphene film transistor.

Graphene transistors were PMOS type and followed conductance provided in the following formula:

$$g_m = \frac{\partial I_D / \partial t}{\partial V_G / \partial t} \quad (1)$$

Measurement of transistors showed an average value of g_m to be 0.2 A/V for the gate voltage of -3 volts and mobility of 120 cm²s⁻¹V⁻¹. Static CMOS logic was not used due to the lack of NMOS transistor using Graphene. The gates in this experiment are dynamic gates made by combination of PMOS transistors and resistors. They have potential applications in electronics such as flexible logic circuits, flexible displays, and biomedical interfaces for wearable electronics.

The fabrication process for building the transistors is as follows. Each transistor is constructed using two different layers. The first layer consists of the printed Graphene, and the second layer contains the source and drain. Between the source and drain honey was used as gate. Fig. 4 shows a sample of AND gate under the test.

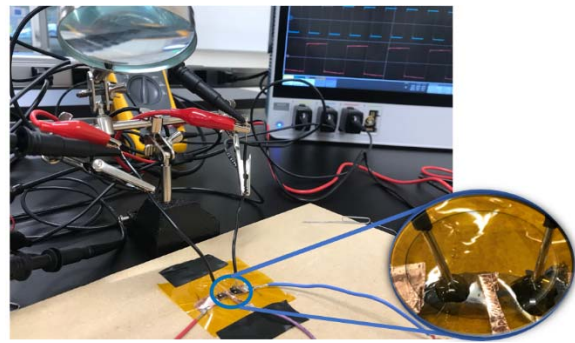


Fig. 4. A fabricated thin film AND gate with Graphene transistors under the test.

4. Conclusions

This study showed that Graphene base and honey gate can be used for fabrication of logic gates in flexible electronics. The technology could be used for prototyping and highly flexible sample can be made. Moreover, the device can be used for interfacing applications and it benefits from direct measurement and immediate command from measures value such as biomedical parameters into transistor gates at the input of logic gates in the circuits.

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Indoor Wayfinding in Large Environments Using Visible Light Communication

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Summary: This paper investigates the applicability of an intuitive wayfinding system in complex buildings using Visible Light Communication (VLC). Typical scenarios include finding places. Data from the sender is encoded, modulated and converted into light signals emitted by the transmitters. Tetra-chromatic white sources are used providing a different data channel for each chip. At the receiver, the modulated light signal, containing the ID and the 3D geographical position of the transmitter and wayfinding information, is received by SiC photodetector with light filtering and demultiplexing properties. Each luminaire for downlink transmission become a single cell, in which the optical access point (AP) is located in the ceiling and the mobile users are scattered within the overlap discs of each cells underneath. The light signals emitted by the LEDs are interpreted directly by the receivers of the users underneath. The effect of the location of the APs is evaluated and a model for the cellular networks is analyzed using orthogonal topologies. A 3D localization design, demonstrated by a prototype implementation, is presented. Uplink transmission is also implemented and the 3D best route to navigate calculated. The results showed that the system allows to determine the position of a mobile target inside the network, to infer the travel direction along the time and to interact with information received optimizing the route towards the destination.

Keywords: Visible light communication, Indoor navigation, Bidirectional communication, Wayfinding, Optical sensors, Transmitter/Receiver, Multiplexing/Demultiplexing techniques.

1. Introduction

Optical wireless communication has been widely studied during the last years in short-range applications. Therefore, communications within personal working/living spaces are highly demanded. Multi-device connectivity can tell users, from any device, where they are, where they need to be and what they need to do when they get there. In future accurate indoor positioning might not be viable by sole utilizing RF communications. Research has shown that compared to outdoors, people tend to lose orientation a lot easier within complex buildings [1, 2]. Fine-grained indoor localization can be useful enabling several applications [3, 4].

To support people's wayfinding activities in unfamiliar indoor environments, a method able to generate ceiling landmark route instructions using VLC is proposed. VLC is a data transmission technology [5, 6]. It can be easily used in indoor environments using the existing LED lighting infrastructure with few modifications [7, 8]. This means that the LEDs are twofold by providing illumination as well as communication. Research is still needed to design LED arrangements that can optimize communication performance while meeting the illumination constraints.

Each luminaire for downlink transmission become a single cell in which the optical access point (AP) is located in the ceiling. Data from the sender (the map

information and the path messages necessary to wayfinding) is encoded, modulated and converted into light signals emitted by the transmitters. Tetra-chromatic white sources are used providing a different data channel for each chip. The use of white polychromatic LEDs in the ceiling offers also, the possibility of Wavelength Division Multiplexing (WDM) which enhances the transmission data rate. To receive the mapped information generated from the ceiling light in visual light form the users are equipped with a receiver module that displays this fabricated information in the mobile terminal. Receiver modules includes a photodetector based on a tandem a-SiC: H/a-Si:H pin/pin light controlled filter [9-11] that multiplexes the different optical channels, performs different filtering processes (amplification, switching, and wavelength conversion) and finally decodes the encoded signals, recovering the transmitted information. This kind of receiver has proved to be adequate when used in large indoor environments with a 2D building model [12]. However, vertical positioning is also important [13, 14]. 3D tracking is difficult by the need for additional coverage for passages between floors (e.g. stairs, elevators).

In this paper, a dynamic LED-assisted positioning and navigation VLC system is proposed. A 3D model for the building is established. The transmitted information, indoor position, motion direction as well as bi-directional communication are determined.

2. VLC Dynamic Navigation System

2.1. VLC Virtual Scenario

When we are looking for the shortest route to a place, we want to be guided on a direct, shortest path to our destination. A destination can be targeted by user request to the central manager (CM).

The dynamic navigation system is composed of several transmitters which send the map information and path messages required to wayfinding. Mobile optical receivers, using joint transmission, extracts their location to perform positioning and, concomitantly, the transmitted data from each transmitter. To synchronize the signals from multiple LEDs, the transmitters use different ID's, such that the signal is constructively at the receiver. Bidirectional communication between the emitters and the receivers is available in strategic optical access point (Li-Fi zone).

The indoor environment chosen is a shopping center with several floors. The 3D virtual floorplan with the information about the number of floors and the nodes localization is draft in Fig. 1. The ground floor is level 0, and the user can go both below and above from there. Each unit cell can be referred as $C_{i,j,k}$ where i, j, k are respectively the line, the row and the level of the top left node of each unit cell. To exemplify, Cell $C_{2,1}$ is depicted in the figure for levels -1, 0, 1 and 2. Therefore, each node, $X_{i,j,k}$ carries its own color, X , (RGBV) as well as its horizontal (line and row) and vertical (level) ID position in the network (i,j,k). For data transmission commercially available polychromatic white LEDs were used.

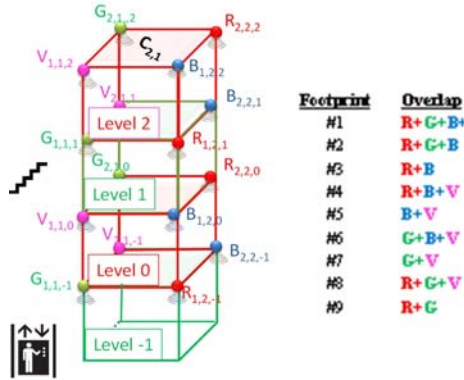


Fig. 1. The 3D virtual floorplan.

In Fig. 2 the ceiling plans for the LED array layout in a 3D building, for even and odd floors is shown (R,G,B,V are the modulated color spots for data transmission in each level). A multi-layer orthogonal geometry with lines and rows for each floor was considered for a square unit cell framed with four tetra-chromatic LEDs. Only one chip, in each node, is modulated, the Red (R; 626 nm), the Green (G; 530 nm), the Blue (B; 470 nm) and the Violet (V; 390 nm), the other have a dc driving current for white perception [3, 15].

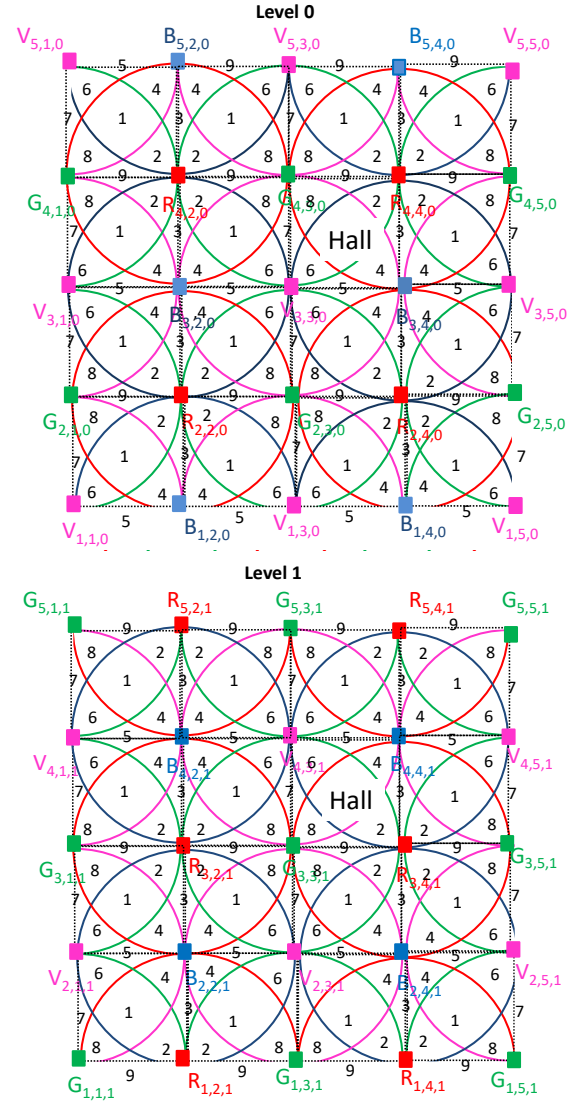


Fig. 2. Illustration of the proposed scenario. Ceiling plans for the LED array layout in a 3D building, for even and odd floors. The footprint regions assigned to the overlaps are pointed out in Fig. 1.

To receive the information from several transmitters, the receiver must be positioned where the circles from each transmitter overlaps, producing at the receiver, a multiplexed (MUX) signal that, after demultiplexing, acts twofold as a positioning system and a data transmitter. The overlap regions, the footprints, are also pointed out in Fig. 1. Nine reference points, for each unit cell, are identified and give a fine grained resolution in the localization of the mobile device across each cell. The user positions is represented as $P(x, y, z)$ by providing the horizontal positions (x, y) and the correct floor number z . By integrating floor number information into the previous 2D system [16], the overall performance of the system will not be significantly affected as this time the 2D positions are more important and the floor detection accuracy is high enough to handle automatic floor plan changes. Also, there needs to be stairs or an elevator to enable a connection between the floors.

The VLC photosensitive sensor, in the receiver module, is a double pin/pin photodetector based on a tandem heterostructure, p-i(a-SiC:H)-n/p-i(a-Si:H)-n sandwiched between two transparent contacts [3]. The device is an active filter able to identify the wavelengths and intensities of the impinging optical signals. Its quick response enables the possibility of high speed communications [17]. Bi-directional communication between VLC emitters and receivers is available at a VLC ready handheld device through a control manager (CM) interconnected with a billboard receiver located in a Li-Fi zone.

Different users, are considered. Depending on the time available, they can find a friend, shop, have a meal or rest. When arriving, they notify the controller manager (CM) of their location (x,y,z), asking for help to find the best way for their needs. A code identifies each user. If a user wishes to find a friend both need previously to combine a common code for the schedule meeting. The first arriving initiates the alert notification to be triggered when the other is in his floor vicinity and generates a buddy list for the meeting. The buddy finder service uses the location information from the network's VLC location from both users to determine their proximity and sends a response message with the location and path of the meeting point.

2.2. Encoding Techniques: Communication Protocol

An on-off keying modulation scheme was used to code the information. To create a communication protocol and overcome the technology constraints, a 64 bits data frame was designed. In Fig. 3 it is displayed the representation of one original encoded message, in a time slot.

In Fig. 3, the transmitted node packet ($R_{4,4,0}$; $G_{4,3,0}$; $B_{3,4,0}$; $V_{3,3,0}$) from cell $C_{4,3,0}$ (ground floor Hall; Fig. 2) are displayed. Several control fields are used depending on the type of transmitter. The synchronism (Sync) is the first and the next are used for the identification of the ceiling. A stop bit is always used at the end of each frame. For the luminaires, the cell ID (x,y,z) begin the second block while for the CM transmitter a pattern [000] precedes this identification.

Those sequences are followed by a new block (pin_1) with the password of the user if a request is needed, if not this block is set at zero and the user only receives its own location. The last block is used to transmit the wayfinding message and may include, in the beginning, the code of the request meeting (pin_2).

The same synchronization header [10101], in an ON-OFF pattern, is imposed simultaneously to all emitters. Each color signal (RGBV) carries its own ID-BIT. Cell's IDs are encoded using a binary representation for the decimal number. At the beginning of the binary code of the z coordinate an extra bit was added to represent the floor's sign: setting that bit to 0 is for a positive number, and setting it to 1 for a negative number, the remaining 3 bits indicate the

absolute value of the z coordinate. The last bits, in the frame, are reserved for the message send by the $X_{i,j,k}$ node (payload data). When bidirectional communication is required, the user has to register by choosing a user name with 4 decimal numbers, each one associated to a color channel. Here, each channel (RGBV) needs a binary 4-digid code. The decimal numbers assigned to each ID block are pointed out in the Fig. 3. Results show that $R_{4,4,0}$, $G_{4,3,0}$, $B_{3,4,0}$ and $V_{3,3,0}$ are the transmitted node packets, in a time slot, from an AP ($C_{4,3,0}$) located in level 0. In this location, an identified user 9119 [1001, 0001, 0001, 1001] receives his response message [wayfinding needs] from the infrastructure.

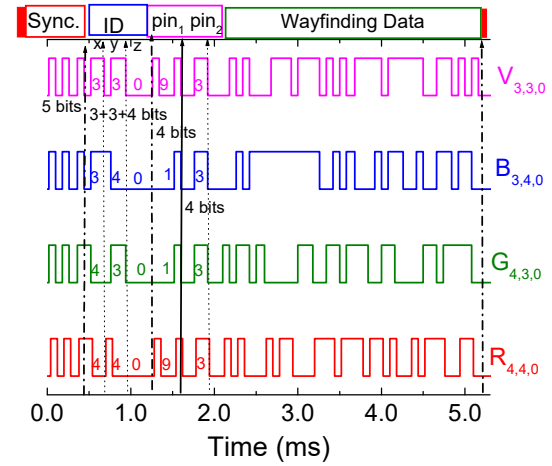


Fig. 3. Data frame structure. Representation of one original encoded message, in a time slot: $R_{4,4,0}$; $G_{4,3,0}$; $B_{3,4,0}$ and $V_{3,3,0}$ are the transmitted node packet from the unit cell $C_{4,3,0}$ in the network.

3. Results

3.1. Travel Direction: Wayfinding Services

To compute the point-to-point along a path, we need the data along the path. As a PoC, in the lab, a navigation data bit transition was tested by moving the receiver along a known pattern path.

In Fig. 4, the MUX signal acquired by a user, as well as the decoded information, is displayed in different instants. The visualized cells, paths and the reference points are also shown. The user enters the floor 1 (see Fig. 2) by line #7 ($C_{4,1,1}$), it goes to position #1(t_0) being directed by the CM into the pretended directions ($C_{4,2,1}$ # 5) where he arrives at t_3 passing through footprints #3 (t_1) and #1(t_2) from the next cell.

Results show that the location and path of a mobile receiver was obtained based on the dynamic LED-based navigation system. As the receiver moves between generated point regions, the received information pattern changes. For instance, when the receiver moves in the reverse direction, from $C_{4,1,1}$ #3 to $C_{4,1,1}$ # 1 two different ID channels are added ($G_{3,1,1}$ and $V_{4,1,1}$). Here, the 4-bybinary bit code has changed from [1010] to [1111]. In the forward direction ($C_{4,2,1}$)

the added channels have the same color but different ID ($G_{3,3,1}$ and $V_{4,3,1}$). So, just by tracking the path in successive instant the direction of the receiver is known. Going forward corresponds to crossing lines #3 and #7, turning left cross of line #5 and turning right to line #9.

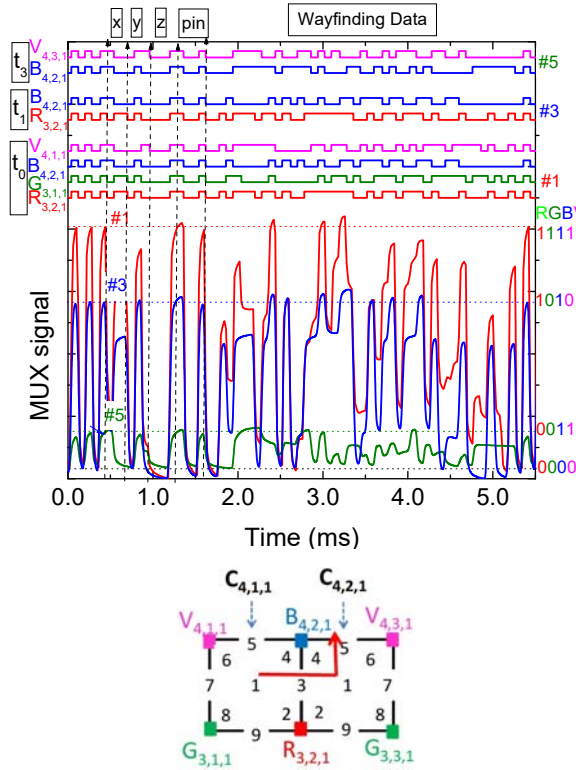


Fig. 4. Fine-grained indoor localization and navigation in successive instants. On the top the received channels packets are decoded at each time step [R, G, B, V].

3.2. Bi-directional Communication: Buddy Services

Bi-directional communication is available at ready handheld device through a control manager (CM) interconnected with a billboard receiver located at each unit cells in a Li-Fi zone ($C_{i,j,k}$, #1). In Fig. 5, the MUX synchronized signals from two identify users are displayed. In the right side, the match between the MUX signal and the 4-binary code is pointed out. On the top the decoded channels packets are shown [R, G, B, V].

Taking into account Fig. 2, results show that in a time slot, two identified users (“2015” and “7261”; pin_1) have successively (t_0 , t_1) request to the buddy wayfinding services, the right track (wayfinding data) for their previously scheduled meeting, pin_2 [0011/3;]. At the right hand of the figure the scenario is illustrated. In the proposed scenario, user “2015” initiates the alert notification ($C_{4,1,1}$; t_0) to be triggered in his floor vicinity (level 1) and initializes the buddy list that includes all the users who have the same meeting code (pin_2 : [0011]; 3). User “7261” arrives later ($C_{2,3,-1}$; t_1), identifies himself and uses the same

code in the buddy wayfinding services to track the best way to the scheduled meeting. For route coordination the CM sends a personalized “response” message at the requested position ($C_{2,3,-1}$). In this message the buddy finder service uses the location information from both to determine their proximity and sends a response message to user “7261” with the best route to the meeting.

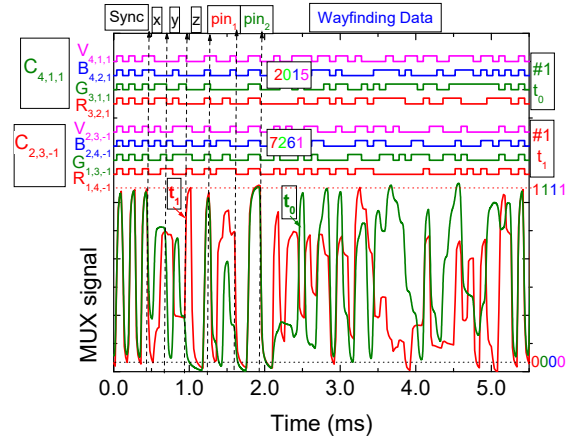


Fig. 5. MUX/DEMUX signals assigned to a “request”. MUX signal received by the CM receiver from two users (“2015” and “7261”) at different locations ($C_{4,1,1}$; #1 and $C_{2,3,-1}$; #1) in successive instants (t_0 and t_1). On the top the transmitted channels packets are decoded [$X_{i,j}$].

4. Conclusions

This paper proposes a generating method of ceiling landmark route instructions using VLC. For lighting, data transmission and positioning, white LEDs were used. A SiC optical MUX/DEMUX mobile receiver decodes the data and, based on the synchronism and ID of the joint transmitters, it infers its path location, timing and user flows.

A 3D building model for large indoor environments was presented, and a VLC scenario in a three level building was established. The communication protocol was presented. Bi-directional communication between the infrastructure and the mobile receiver was analyzed. Global results show that the location of a mobile receiver, concomitant with data transmission is achieved. The dynamic LED-aided VLC navigation system enables to determine the position of a mobile target inside the network, to infer the travel direction along the time and to interact with received information.

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Redesign of the Vehicle Trajectory Inside an Intersection Using Visible Light Communication

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Summary: In this work the communication between the infrastructures and the vehicles, between vehicles and from the vehicles to the infrastructures is performed through Visible Light Communication (VLC) using the street lamps and the traffic signaling to broadcast the information. Data is encoded, modulated and converted into light signals emitted by the transmitters. Tetra-chromatic white sources are used providing a different data channel for each chip. As receivers and decoders, SiC Wavelength Division Multiplexer (WDM) devices, with light filtering properties, are used. To control the arrival of vehicles to an intersection and schedule them to cross at times that minimize delays are primary objectives. A Vehicle-to-Everything (V2X) traffic scenario is established and bidirectional communication between the infrastructure and the vehicles is tested, using the VLC request/response concept. A phasing traffic flow is developed as a proof of concept. The experimental results confirm the cooperative VLC architecture showing that communication between connected cars is optimized using a request/response concept. An increase in the traffic throughput with least dependency on infrastructure is achieved.

Keywords: Vehicular communication, Light fidelity, Visible light communication, White LEDs, SiC photodetectors, OOK modulation scheme, Traffic control.

1. Introduction

The Visible Light Communication (VLC) holds special importance when compared to existing forms of wireless communications [1, 2]. VLC seems to be appropriate for providing wireless data exchange for automotive applications. Visible light represents a new communication opportunity for vehicular networking applications. The communication is performed through VLC using the street lamps and the traffic signaling to broadcast the information.

An Intersection Manager (IM) can increase the throughput of the intersection by exchanging information with and directing incoming Connected Autonomous Vehicles [3-5].

Our goal is to increase the safety and throughput of traffic intersections using VLC connected cooperative driving. Two emerging technological trends are redesigning the physical world: the self-driving and the remote driving [6, 7].

Vehicular Communication Systems are a type of network in which vehicles and roadside units are the communicating nodes, providing each other with information, such as safety warnings and traffic information [8]. Communication between fixed locations and vehicles (Infrastructure-to-Vehicle, I2V) between vehicles (Vehicle-to-Vehicle, V2V), and between vehicles and fixed locations (Vehicle-to-Infrastructure, V2I) is essential to transfer information in real time. The I2V applications focus on utilizing the traffic related infrastructure, such as traffic light or streetlight to communicate useful information.

The proposed system is composed of several transmitters, the street lights and the traffic signals, which transmit map information and traffic messages required to the moving vehicles. Data is encoded, modulated and converted into light signals emitted by the transmitters. Then, this information is transferred to receivers installed in the vehicles. Every street light has their differentiable Identifications (IDs) for the generation of the visible light signal that transmits the map information through a Visible Light Transmitter module. Tetra-chromatic white sources are used providing a different data channel for each chip. Every vehicle is equipped with a receiver module for receiving the mapped information generated from the street. The receiver modules include a photodetector based on a tandem a-SiC:H/a-Si:H pin/pin light controlled filter [9-11] that multiplex the different optical channels, perform different filtering processes (amplification, switching, and wavelength conversion) and decode the encoded signals, recovering the transmitted information.

In this work, a two-way communication between vehicles and the traffic lights is implemented. The redesign of the trajectory is presented. Street lamps and traffic lights broadcast the information. The On-vehicle VLC receivers decode the messages and perform V2V distance measurements. A V2X traffic scenario is proposed and characterized. A phasing traffic flow is developed as a Proof-of-Concept (PoC). The simulated results confirm that the redesign of the intersection and its management through the

cooperative request/response VLC architecture allows to increase the safety and to decrease the trip delay.

2. Vehicular Communication Using VLC

2.1. Redesign Concepts

The redesign of the traffic-actuated controller uses vehicle request/respond message information to generate phase durations appropriate to accommodate the demand on each cycle. Examples of the representation of a redesigned phasing diagram, a functional area with two-way-two-way intersection and a timing function configuration are presented in Fig. 1. In Fig. 1a, a phasing diagram is displayed. Each Timing Function (TF) controls only one movement. Since two movements can proceed simultaneously without conflict as shown in Fig. 1b, hence two of the timing functions will always have simultaneous control, as exemplified in Fig. 1c. The problem that the traffic-actuated intersection manager has to solve is to allocate the reservations among a set of drivers in a way that a specific objective is maximized. Signal timing involves the determination of the appropriate cycle length and apportionment of time among competing movements and phases. The timing apportionment is constrained by minimum “green” times that must be imposed to provide pedestrians to cross and to ensure that motorist expectancy is not violated.

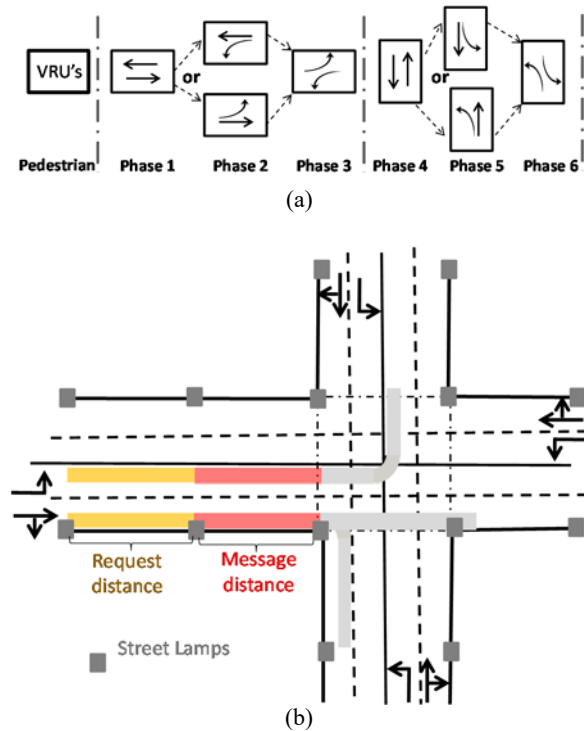


Fig. 1 (a, b). a) Representation of phasing diagram.
b) Physical area and channelization.

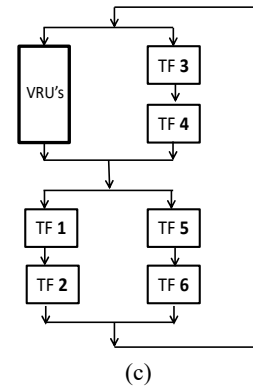


Fig. 1 (c). c) Timing function configuration.

The use of both navigation and lane control signs to communicate lane restrictions is demanding. Downstream from that location (request distance in Fig. 1b), lane restrictions should be obeyed. Vehicles may receive their intentions (e.g., whether they will turn left or continue straight and turn right) or specifically the need to interact with a traffic controller at a nearby crossroad (message distance in Fig. 1b). In the sequence, a traffic message coming from a transmitter nearby the crossroad will inform the drivers of the location of their destination (i.e., the intended intersection exit leg).

2.2. V2X Communication Scenario

A V2X (I2V2V2I2V) communication link, in a light traffic controlled crossroad, was simulated. The crossroad link is displayed in Fig. 2.

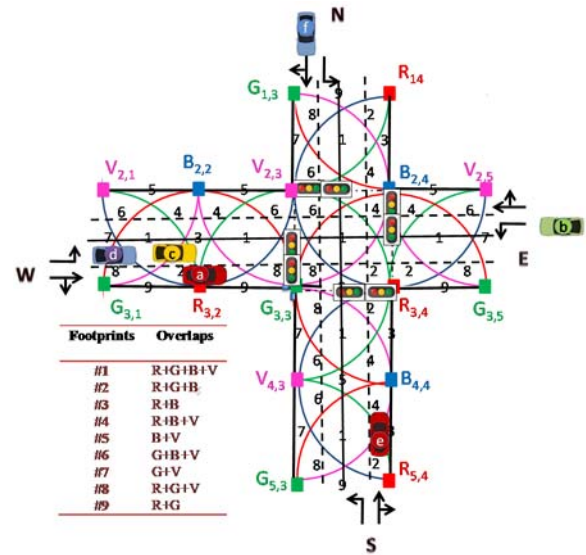


Fig. 2. V2X lighting plan model and generated joint footprints in a crossroad (LED array=RGBV color spots).

To build the I2V it is proposed a simplified cluster of unit square cells in an orthogonal topology that fills all the service area [13]. To realize both the communication and the street illumination, white light

pattern [000000] means that a response message is being sent by the controller manager. Here, the signal controller responds to the request of the vehicle located in position # 8 ($R_{3,2}$, $G_{3,3}$, and V_{23}) at the request time (request distance). This response is received in the unit cell adjacent to the crossroad (message distance, Fig. 1b) that shares a common node ($R_{3,2}$) with the request distance (see Fig. 3).

In Fig. 4, a MUX signal due to the joint transmission of four R, G, B and V optical signals, in a data frame, is displayed. The bit sequence (on the top of the figure) was chosen to allow all the *on/off* sixteen possible combinations of the four input channels (2^4).

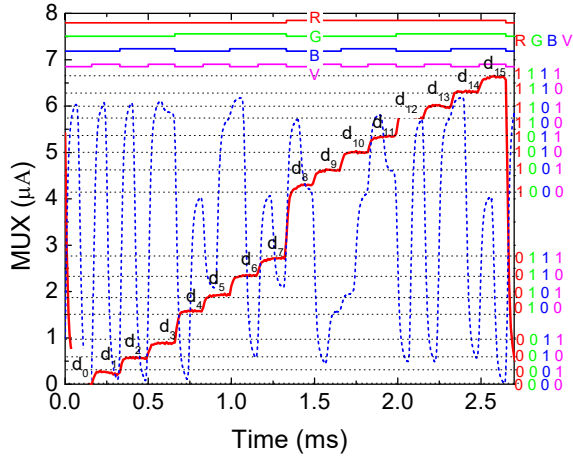


Fig. 4. MUX signal of the calibrated cell. On the top the transmitted channels packets [R, G, B, V] are depicted. A received MUX signal is also superimposed to exemplify the decoding algorithm.

Results show that the code signal presents as much separated levels as the *on/off* possible combinations of the input channels, allowing decoding the transmitted information [9]. All the levels (d_0 - d_{15}) are pointed out at the correspondent levels, and displayed as horizontal dotted lines. In the right hand side, the match between MUX levels and the 4 bits binary code assigned to each level is shown. For demonstration of the decoding technique, the signal transmitted in Fig. 3a and received, in the same frame of time, is also added (dotted curve). Hence, the signal can be decoded by assigning each output level (d_0 - d_{15}) to a 4- digit binary code [X_R , X_G , X_B , X_V], with $X=1$ if the channel is *on* and $X=0$ if it is *off*.

3. Results

3.1. Led Assisted Navigation

As a PoC, performed in the lab, a navigation data bit transition was tested by moving the receiver along known pattern path as shown in Fig. 5.

Here, it is displayed the MUX signals received when Vehicle *a* enters the crossroad in position #8 (t_1) and it goes straight to position #2 (t_2) (Phase1, TF1), while vehicle *c* turn left, moving across position #1

(Phase2, TF2). Results show that, as the receiver moves between generated point regions, the received information pattern changes. The vehicle speed can be calculated by measuring the actual travelled distance overtime, using the ID's transmitters tracking. Between two consecutive data sets, there is a navigation data bit transition (channel is missing or added). It was observed that when the receiver moves from #8 to #2 one ID channels was lost ($B_{2,4}$) and one are added ($V_{2,3}$). Here, the 4-binary bit code has changed from [1101] to [1110] while Vehicle *c* and *d* change theirs from [1111] to [0011] and Vehicle *b* to [1100]. The receivers compute the geographical position in the successive instants (path) and infer the vehicle's speed. In the following, this data will be transmitted to another leader vehicle through the V2V communication or to control manager at the traffic light through V2I.

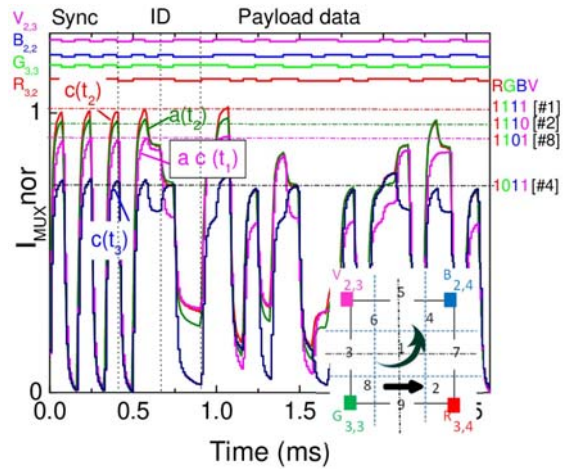


Fig. 5. Normalized MUX signals acquired by a receiver at the crossroad, in positions #1, #2, #4, #6 or #8. On the top the transmitted channels packets [R, G, B, V] are decoded.

3.2. Cooperative System

To model the worst-case scenario, vehicles approaching the intersection from different flows are assumed to have a conflicting trajectory (Fig. 2). Two instants are considered for each vehicle, the request time (t) and the response time (t'). All the requests contain vehicle positions and approach speeds. If a follower exists (Vehicle *d*), the request message from its leader includes the position and speed previously received by V2V. This information alerts the controller to a later request message (V2I), confirmed by the follow vehicle. In the PoC we have assumed that $t_a < t_c < t_d$, and $t_a < t_b < t_c$.

As an example, in Fig. 6, the I2V MUX signals received and decode (on the top of the figure) by the receivers of the vehicles *b*, *e* and *f* are also displayed at request (t_e) and (t'_b and t'_f) response times. In the right side, the received channels for each vehicle are identified by its 4-digit binary codes and associated positions in the unit cell.

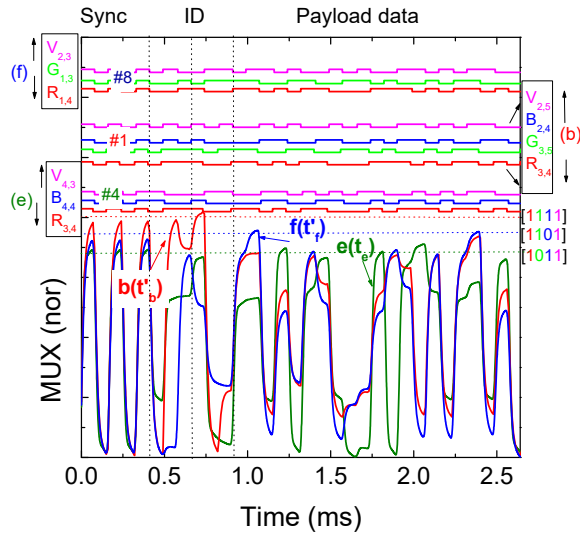


Fig. 6. MUX signals and the assigned decoded messages (at the top of the figure) from vehicles *b*, *e* and *f* at different request and response times (I2V).

After decoding we have assigned position #4 ($R_{3,4} G_{4,4} V_{4,3}$) for Vehicle *e*, position #1 ($R_{3,4} G_{3,5} B_{2,4} V_{2,5}$) for Vehicle *b* and position #8 ($R_{1,4} G_{1,3} V_{2,3}$) to Vehicle *f*, respectively at their request and response times t_e , t'_b and t'_f . Here, $t'_e < t'_f$.

3.3. Traffic Signal Phasing in a V2X Communication

A phasing diagram and a timing function configuration were presented in Fig. 1, for functional areas with two-way-two-way intersection. A traffic scenario was simulated (Fig. 2) using the new concept of VLC request/response messages. A brief look into the process of timing traffic signals is given in Fig. 7.

Redesign traffic-actuated controller uses *a*, *b*, *c*, *d*, *e* and *f* vehicles requesting and responding message information to generate phase durations appropriate to accommodate the demand on each cycle. Each driving vehicle is assigned an individualized time to request (t) and access (t') the intersection. The exclusive pedestrian stage, “Walk” interval begins at the end of Phase 5. (Fig. 1).

A first-come-first-serve approach could be realized by accelerating or decelerating the vehicles such that they arrive at the intersection when gaps in the conflicting traffic flows and pedestrians have been created. However, a one-by-one service policy at high vehicle arrival rates is not efficient. From the capacity point of view it is more efficient, if Vehicle *c* is given access at t'_c before Vehicle *b*, at t'_b to the intersection and Vehicle *d* is given access at t'_d before Vehicle *e*, at t'_e then, forming a west left turn of set of vehicles (platoon) before giving way to the fourth phase (north and south conflicting flows), as stated in Fig. 7.

The speed of Vehicle *e* was reduced, keeping a safe distance between Vehicle *e* and Vehicle *d*.

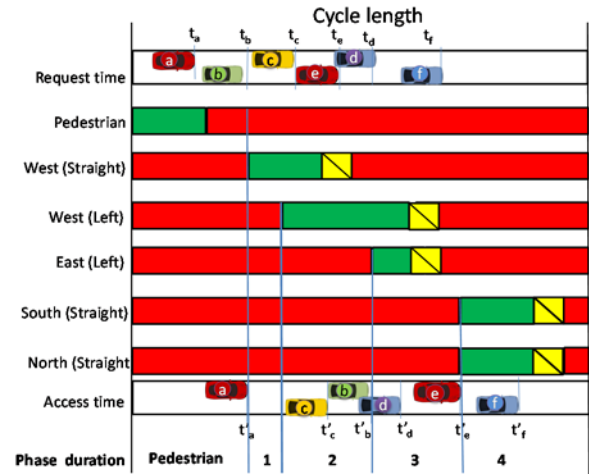


Fig. 7. Requested phasing of traffic flows: pedestrian phase, Phase 1 (W straight flow), Phase 2 (W straight and left flows), Phase 3 (W and E left flows), Phase 4 (N and S straight flows). $t_{[x]}$ is the request time from the Vehicle *x* and $t'_{[x]}$ the correspondent response time from the manage controller.

4. Conclusions and Future Trends

This paper presents a new concept of request/response for the redesign and management of a trajectory in a two-way-two-way traffic lights controlled crossroad, using VLC between connected cars. The connected vehicles receive information from the network (I2V), interact with each other (V2V) and also with the infrastructure (V2I), using the request redesign distance concept. In parallel, a control manager coordinates the crossroad and interacts with the vehicles (I2V) using the response redesign distance concept. A simulated traffic scenario was presented and a generic model of cooperative transmission for vehicular communication services was established. As a PoC, a phasing of traffic flows is suggested. The simulated/experimental results confirmed that the proposed cooperative VLC architecture is suitable for the intended applications. The introduction of VLC between connected vehicles and the surrounding infrastructure allows the direct monitoring of relative speed thresholds and inter-vehicle spacing.

In order to evolve towards real implementation, the performance of such systems still needs improvement, namely the distance between conflicting vehicles along with the trajectories of other opposing vehicles should also be monitored and optimized. As further work, the research team plans to finalize the embedded application, for experimenting in several road configurations with either static or moving vehicles.

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Multifunctional Carbon Fiber Reinforced Polymer Composite Structures: Reinforcing and Sensing

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Summary: In our research, we investigated how electric current connected to multifunctional carbon fiber would affect the dynamic mechanical properties of the polymer matrix composite. The reinforcing carbon fiber, which could have additional functions as well (e.g., self-sensing) was connected to an electric circuit, then we carried out a dynamic three-point bending test. We varied the amperage and connection time of the current and measured bending stiffness and strength. Based on the results, we concluded that there is a current limit (0.5 A), under which the electric current does not modify the dynamic bending properties. We measured the generated Joule heat. We determined the same current limit where the specimen would not heat up too much, although resistance heat should be considered at any amperage connected for longer time. Usually, when carbon fibers are used multifunctionally for both reinforcing and sensing, a low current is applied (~100 mA), therefore it would not change the mechanical behavior of the composite.

Keywords: Multifunctional carbon fiber, Smart material, Self-sensing, Joule heating.

1. Introduction

Carbon fiber reinforced polymer composites (CFRP) are widely used in many industries due to their low density, high strength, load-optimized stiffness and excellent corrosion resistance. One of these industries is transportation, where electric vehicles are dynamically developing, and an important goal is to increase their range. This can be achieved by either increasing the capacity of the batteries or reducing consumption, for example by reducing the weight of the vehicle. One solution is to use reinforced plastic, such as fiber-reinforced composites. Additional weight reduction can be achieved by combining functions: a component can perform multiple tasks (e.g. load-bearing, signal transferring, self-repairing, and energy storage), therefore, the non-load-bearing part can be eliminated [1]. In our previous review article, we categorized multifunctional CFRPs based on their secondary function and application [2]. For instance, monitoring the resistance of a carbon fiber bundle in a specimen would create a structural health monitoring sensor. Bashmal *et al.* [3] investigated the linearity of the sensitivity of carbon fiber. They tested different conducting areas and lengths by varying the number and length of the carbon fiber tows they used. They concluded that specimens with a small number of filaments (3K or under) have linear characteristics and are suitable for low strain levels.

With carbon fiber-based sensors, it is possible to detect deformation, fiber breakage, delamination and failure as the resistance of the carbon fibers will increase during those effects. Wang *et al.* [4] performed drop tests on carbon fiber-reinforced polymer composites with different layer orders while measuring the change in the electrical resistance of the specimens. They investigated several configurations:

in one case, surface resistance was measured, which in their opinion could be used to detect fiber breakage, in another case, the change in resistance was measured through the width of the composite specimen, which they thought was more suitable for detecting delamination. They also mentioned that by measuring the change in resistance, smaller defects can also be detected compared to other non-destructive processes (e.g. ultrasonic testing).

Luan *et al.* [5] used a novel manufacturing technique to create a multifunctional composite: they developed an additive manufacturing process with two printer nozzles and integrated an injection system. With this method they built specimens with continuous carbon fiber and thermoplastic matrix to test the reinforcing and sensing function and injected epoxy resin as curing agent separately for self-healing function. They performed compression tests and three-point bending tests on the material. They achieved continuous and *in-situ* detection of damage. They found that self-healed specimens have better mechanical properties than the original specimens because of the poor initial adhesion between the fiber and matrix and the improved adhesion after the first test caused by the cured epoxy.

With new materials, for example nanocomposites, new opportunities are available. Shape memory polymers could change their geometry triggered by heat, mechanical stress or an electric current. This additional function can be used in different applications where weight saving is crucial or the available space is limited (e.g. space structures, artificial muscles, civil applications) [6]. Other nanocarbon allotropic materials, for example graphene nanoplatelets can also be used for sensing, they can even be applied on originally non-conductive materials. Sethy *et al.* [7] spray coated a glass fiber

reinforced composite beam with this material. They investigated the sensitivity of this sensor with different mechanical tests and obtained a high gauge factor, which makes it a feasible material for multifunctional applications.

In parallel with the ongoing development of electronic, information and telecommunications technologies, harmful electromagnetic radiation called electromagnetic interference (EMI) is also a growing problem. It can interfere with electronic devices, which can be particularly risky for applications such as medical technology, the aerospace and military industries, and can also be harmful to health. To avoid these, Faraday-cage-based equipment and materials are used, which partially or completely surround the device responsible for radiation, partially transmit, reflect or absorb electromagnetic radiation, thus limiting the adverse effects. Aripin *et al.* [8] used carbon fiber reinforced thermoplastic matrix tapes and made three dimensional specimens with different layer orders. Based on both electromagnetic insulation and mechanical properties, they concluded that unidirectional tapes are best for EMI shielding and structural functions. They explained it by the good through-thickness conductivity and interlaminar adhesion of the CFRP layup.

Another example of multifunctional CFRP is the resistive heating of carbon fibers, which can be used to help the crosslinking process, create the necessary heat for bonding and welding, or deice airplane wings. Weiland *et al.* [9] simulated and experimentally validated the effectiveness of self-heating CRFP. They also made mold from CFRP. They summarized their results in such a way that although CFRP is suitable for *in-situ* heated mold material by resistive heating, the induced temperature will not be homogeneous due to the relatively low thermal conductivity. With their simulation, they found that varying cross sections of the CFRP tool results in a more homogeneous heat distribution.

Welding components together requires pressure and heat. Using the self-heating property of CFRP makes it possible to locally heat the parts to weld them. Flanagan *et al.* [10] used carbon fiber reinforced polyether ether ketone (PEEK) composite parts and they heated them by Joule-heating generated by induced alternating current. They inspected the welded assembly by mechanical testing and optical microscopy and concluded that although it had similar mechanical properties to adhesive bonded parts, it had voids, delamination and warpage caused by the welding process. They predict that this manufacturing process could be used in different industries, for example for automotive component assembling.

Pan *et al.* [11] also used the self-heating effect of carbon fiber reinforced PEEK. They added graphene to enhance conductivity and to achieve better heat distribution, therefore decreasing the power needed for larger scale parts. In their research, they tested the self-heating function by simulating cold application, where icing of the part can occur, therefore efficient deicing

is needed. They achieved rapid heating and deicing by using the *in-situ* heating concept.

The generated heat can be a disadvantage in the case of the electrical multifunctional application of CFRP: when the carbon fibers are used as sensors, the generated heat changes the resistance of the sensor, which would cause inaccurate sensing. In our recent article, we presented how a change in temperature would result in a change in resistance (Fig. 1).

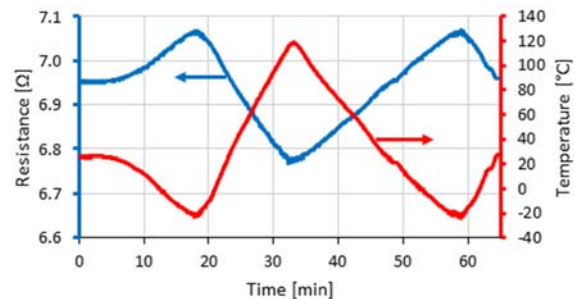


Fig. 1. Resistance as a function of temperature during a heating-cooling cycle for heat dependence measurement [12].

In that article, we showed that carbon fibers have a negative linear temperature coefficient, which means that with increasing temperature the resistance of carbon fibers decreases. Their heat sensitivity is significant, a temperature difference of 100 °C results in the same resistance change as breaking the specimen would cause, therefore this effect should be taken into account and it should be compensated for in sensing applications [12].

The development of electric vehicles includes examining possibilities of how the body could also be an energy storage component. The so-called structural supercapacitor is a capacitor-like energy storage unit in which the electrodes are usually activated carbon fiber layers with an electrically insulating layer between them. In addition to energy storage, they also have load-bearing capacity [13].

Using structural batteries, for example combining lithium-ion battery with carbon fiber reinforced polymer composite is ideal for weight and volume saving. Moyer *et al.* [14] used the carbon fiber's mechanical properties for the structural function and the electric properties for the current collector function. They used this concept in a CubeSat, which is a miniature satellite type. As the outer dimensions are strictly given, all the necessary parts should fit inside. They eliminated the traditional battery by using their multifunctional composite, therefore saving volume for other instruments.

The use of supercapacitors and structural batteries raises several safety issues: what happens to the stored electrical energy in the event of a collision, and how electric current affects the dynamic mechanical properties of CFRP.

Sierakowski *et al.* [15] investigated the relationship between the impact resistance of composites and the electric current conducted through samples using drop

tests. A direct current of 25 A and 50 A was passed through the pieces for different times. They observed that if the current is passed through the specimens immediately before the mechanical impact, the impact resistance increases, but if they connected the current for longer, impact resistance decreases. The negative effect of prolonged current was explained by resistance heating, but the authors did not explain why a brief electric current would improve the mechanical properties.

Barakati and Zhupanska [16] modelled the behavior of composite plate under a mechanical impact load. They used the model to investigate the response of a composite plate during the impact, as they subjected the specimen to an electric current and placed it in a magnetic field. Their calculated results showed that the applied electromagnetic field reduced the magnitude of vibration and improved damping properties. They also claimed that maximum deflection and stress are also affected by the electromagnetic field, but it also depends on the layer order.

Amaro *et al.* [17] investigated the effect of electric current on carbon fiber-reinforced composites, with particular reference to their impact resistance. In their experiments, unidirectional prepreps were used, from which composites of different layer orders were formed, and then subjected to drop tests. The electric current flowed through the thickness of the specimens, perpendicular to fiber orientation. A decrease in impact resistance was observed on specimens in which the electric current flowed for a longer time. They repeated the drop test on the same specimen until the tup (the tip of the falling weight) fully broke through the specimen. Based on those results, they concluded that electric current has a negative effect on impact fatigue life.

The aim of this study is to show how electric current affects the mechanical properties of a carbon fiber composite structure as there are opposing results. Therefore, we investigated short and long connection times and different amperage.

2. Experiments

In our experiments, we investigated how the electric current flowing through the specimen would affect impact bending stiffness. We made glass fiber (225 g/m² areal density, Owens-Corning Composites LLC, USA) reinforced epoxy (MR 3016 with MH 3124 hardener, 100:40 mixing ratio, Ipox Chemicals, Hungary) specimens and laminated a carbon fiber (Sigrafil C T24-5.0/270-E100, SGL Carbon SE, Germany) layer into them as the penultimate layer (Fig. 2). We cured the specimens at 80 °C for 4 hours.

We tested the specimens with a three-point bending-like impact test: we modified the specimen support and the tup of a falling-weight impact tester (Ceast Fractovis 9350, Ceast, Italy) to form an arrangement similar to the Charpy impact test, but with a falling weight moving vertically instead of the pendulum. The carbon fibers were on the drawn side

of the specimen. Falling height was 638 mm, the speed of impact was 4.43 m/s, and impact energy was 15 J.



Fig. 2. Side and top view of a specimen.

In order to apply current to the carbon fibers, we used copper blocks clamped onto the dry carbon fibers sticking out of the specimen. The copper block had to be fixed to the machine, so we adjusted the support with two custom parts made by fused deposition modelling (Fig. 3).

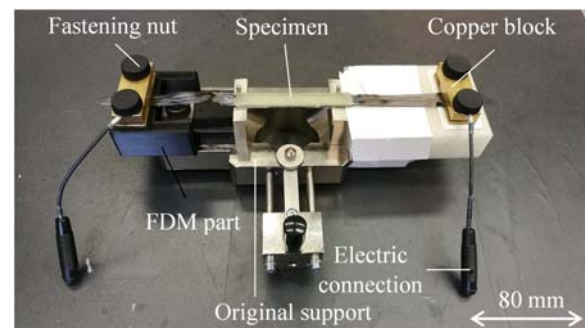


Fig. 3. Custom designed support for dynamic testing and copper block positioning.

We applied different currents (0.5; 1; 1.5; 1.75; 2; 2.25 A) to the carbon fibers reinforcing the composite through the copper blocks with a power supply (GW Instek GPD-3303S, GW Instek, Taiwan). We varied the connection time (2; 10; 30; 60; 90 s) of the flow of electric current before the impact and switched the power supply off immediately after the impact test. We prepared 12 specimens in one batch, from each batch, and tested 3 pieces as references without current.

We measured the temperature of the specimens and how it changed due to the electric current. We used a thermal imaging camera (A325SC, FLIR Systems, USA) to monitor the temperature change under the aforementioned current values for 90 s (the longest connection time).

3. Results and Discussion

To determine which is the highest current and longest connection time to apply to the specimen we started with temperature measurement. We calculated the surface average of the maximal temperature after 90 s connection time (Table 1).

The measured data show that even the lowest inspected current heated up the specimen from room temperature in 90 s. However, at the lowest current,

the temperature stabilized after the electric power was connected (Fig. 4). Based on our previous research [12], this increment doesn't affect the feasibility of the carbon fiber as sensor, but Joule heating should be taken into account when an electric current is used. As expected, with increasing current, the maximum temperature increased. With increasing current, the temperature did not stabilize in 90 s – it would have warmed up further (Fig. 4). At a longer connection time and higher current (see Fig. 4 2.5 A), the specimen become so hot that it degraded the epoxy resin and the support, therefore we maximized the current and connection time.

Table 1. Average maximum temperature on the surface of the specimen after 90 s connection time.

Current [A]	Average maximum temperature [°C]
0.50	31
1.00	49
1.50	75
1.75	83
2.00	95
2.25	127

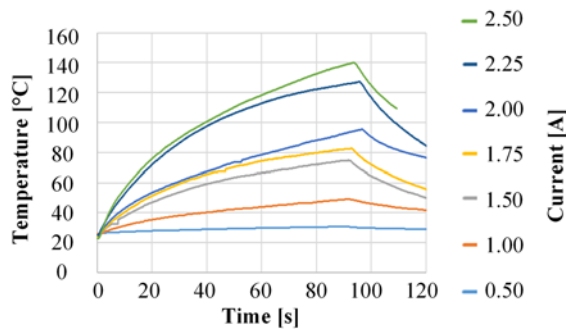


Fig. 4. Surface temperatures reached at various amperages in 90s.

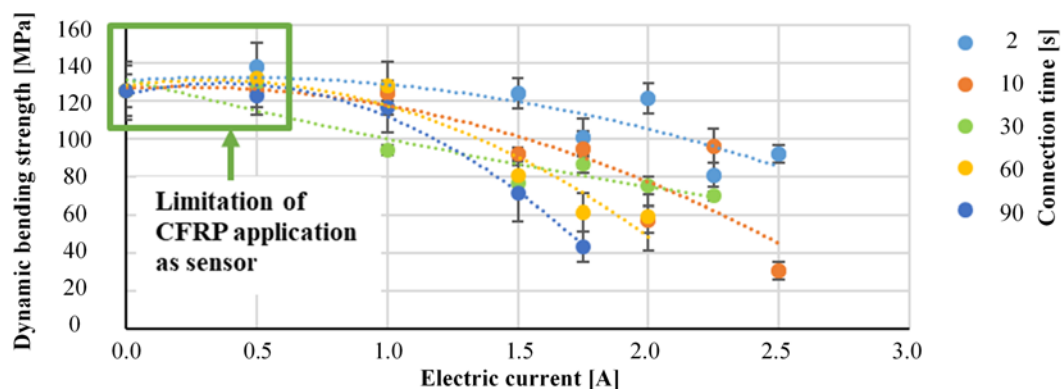


Fig. 5. The effect of electric current and connection time on the dynamic bending strength of carbon-glass hybrid reinforced composite.

4. Conclusion

Reinforcing carbon fibers nowadays are used for additional functions as well, such as sensors. In these applications, a current flows through the composite,

After temperature measurement, we carried out the mechanical test in order to determine the dynamic bending behavior. Based on the maximal impact force (F) and specimen geometry, we calculated the dynamic bending strength (σ_{db}) of the specimen (1).

$$\sigma_{db} = \frac{3FL}{2bh^2}, \quad (1)$$

where L is the support span, b is the width of specimen and h is the thickness of the specimen. The dynamic bending strength was plotted as a function of the current (Fig. 5), so that the effect of the magnitude and the time of the current on bending strength became visible. The diagram clearly shows that bending strength decreases with increasing current and connection time. At higher currents or longer connection times, the Joule heat heats up the specimen and this can lead to softening.

The dynamic bending modulus (E_{db}) was calculated based on the same specimen geometry and the slope of the force-deflection diagram (2).

$$E_{db} = \frac{1}{4} \frac{L^3}{bh^3} \frac{F}{f}, \quad (2)$$

where f is the deflection at maximum force. The calculated dynamic bending modulus values are in Table 2, the reference value (without current) was 3630 ± 205 MPa. It can be seen from the table that increasing the current and connection time decreases modulus. Therefore, if carbon fibers are used as a sensor, it is advisable to use a current of less than 0.5 A so that the mechanical properties of the structure are not affected. In practice, this can be solved by, for example, sampling every half minute with the current switched on for a short time only.

which can alter the mechanical properties of the material. In our research, we investigated the effect of the magnitude of the current and connection time on the dynamic bending properties of composite specimens. The results indicate that CFRP is well

suited for multifunctional purposes, but particular attention needs to be paid to the change in mechanical properties due to the heating effect of the current. Either a low current electric circuit should be used

($I < 0.5$ A), or the electric current should only be switched on for a short period of time (less than 2 s in the case of a higher current).

Table 2. Dynamic bending modulus in the case of different currents and connection times in MPa. X – the specimen become too hot to handle safely, n.a. means that appropriate data was not available.

Time [s] Current [A]	2	10	30	60	90
0.50	3687±125	3626±327	3426±204	3659±205	3689±242
1.00	3393±412	2879±258	2617±269	3586±573	3217±412
1.50	2297±782	1842±636	2014±263	2036±284	1634±568
1.75	1687±126	1909±918	1924±234	1212±396	1387±258
2.00	3473±240	n.a.	1874±294	981±202	1279±256
2.25	n.a.	2752±228	1589±140	×	×

Acknowledgements

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(057)

Indoor Navigation Based on Visible Light Communication Using an a-SiC:H Photodetector

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Summary: This paper presents the use of a selective device based on a-SiC:H/a-Si:H for the photodetection of visible signals emitted by red, green and blue emitters in a Visible Light Communication (VLC) system. The VLC system employs RGB white LEDs to provide both illumination and information transmission. The acquisition and processing of the measured photocurrent allows the identification of the induced optical excitation, which encodes the spatial position. The system is designed so that the detector's spatial location can be obtained based on the identification of the received optical signals. The methodology used for the photocurrent signal processing involves Fourier transform analysis for frequency identification and the use of a photodetector with spectral selective properties of wavelength identification. A full characterization of the photodetector is presented together with the physical operation that plays the key role in the detection of the output photocurrent.

Keywords: Amorphous SiC, Optoelectronic, Spectral sensitivity, White LEDs, Visible light communication, Indoor positioning.

1. Introduction

Visible Light Communication (VLC) technology constitutes a data communication tool based on the use of visible light that is used for illumination and data transmission [1, 2]. These optical links use LEDs to encode and transmit the signals, and electronic devices generally containing photodiodes to detect the incoming signals from the light sources [3]. It is a free-space, line of sight communication technology [4, 5]. Several applications are currently being developed, spanning from low data rate applications, such as indoor positioning and navigation to more demanding bandwidth applications [6, 7].

In this paper, we propose the use of a multilayered a-SiC:H device [8] to perform the photodetection task at the reception end of a VLC system designed for navigation purposes [9, 10]. The emission optical system is based on RGB white LEDs with the internal red, green and blue emitting chips being modulated at specific frequencies through coding bit sequences normally based on On-Off keying [11, 12]. The photodetector works as an active optical filter operating in the visible part of the electromagnetic spectrum [13]. Its sensitivity can be adjusted through external optical bias supplied by a steady state optical source. This is the key for light wavelength discrimination [14, 15].

The paper is structured as follows. In Section 2, the design of the VLC system is described. Section 3 contains results and respective discussion. Section 4 presents main conclusion and a few guidelines for future work.

2. Architecture of the VLC System

The proposed optical system can be described using the block diagram shown in Fig. 1. It contains two main sub-systems, one related to the input optical signal and a second one to the output electrical signal.

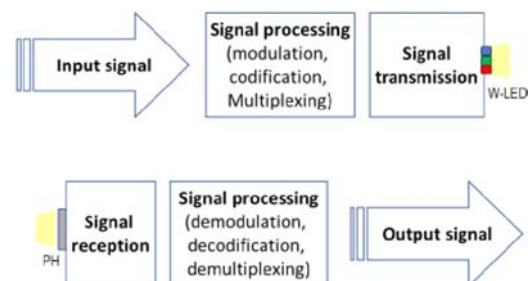


Fig. 1. Block diagram of the visible light communication system.

The input optical signal sub-system includes four white LEDs in which the internal emitting red, green and blue chips are modulated with specific frequencies and bit sequences inducing different optical excitations in the view angle of each LED. The output signal sub-system includes a large area a-Si based photodetector operating in the visible spectrum with a selective spectral photo response depending on external conditions (electrical and/or optical bias). The photocurrent signal produced by the photodetector is dependent on the input optical system. Signal processing at this stage includes demodulation of the electrical signal for demultiplex the different components assigned to each input wavelength,

frequency analysis, and bit sequences decoding. The optical sources are commercial white LEDs with red, green and blue emitters (w-RGB LEDs). Specifications of the w-RGB LEDs are displayed in Table 1.

Table 1. RGB LEDs Specifications.

	Blue	Green	Red
Central wavelength (nm)	464	535	630
Linewidth (nm)	22	48	13
Optical intensity (mcd)	180-505	560-1400	355-900
Far field pattern, 2 θ (°)	180-505	560-1400	355-900

The commercial tri-chromatic white LEDs are designed for general lighting, providing a wide divergence angle around 120 °C, which fits simultaneously the requirements for illumination and communication services as it provides a wide signal distribution. The luminous intensity of each emitter can be independently adjusted; however, it must be regulated to ensure a stable, white perception by the human eye.

The proposed optical sub-system for the VLC system (Fig. 2) includes several navigation cells with four white LEDs modulated with different bit sequences. Green emitters transmit the identification code of the navigation cell. Red emitters are used to transmit the western cardinal directions and the blue chips the eastern directions. North and South directions are ascribed to the modulation frequency. Frequency of the North positions is half of the one used for the emitters assigned to the South directions.

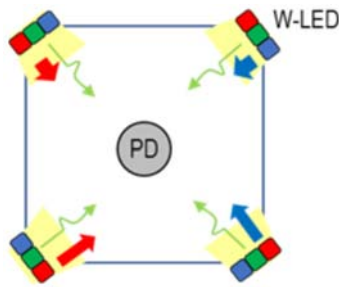


Fig. 2. Configuration of the luminaires used for illumination and communication.

The optical signal generated by the four LEDs and detected by the photodetector is spatially assigned its position, which provides a direct correlation of the measured induced photocurrent to the detector position.

3. Device Configuration and Characterization

Fig. 3 shows the semiconductor configuration used as integrated photodetector and tunable filter device. It is a monolithic heterojunction composed by two pin structures based on a-Si:H and a-SiC:H built on a glass

substrate covered by two transparent electrical contacts. The device was fabricated by PECVD at 13.56 MHz radio frequency.

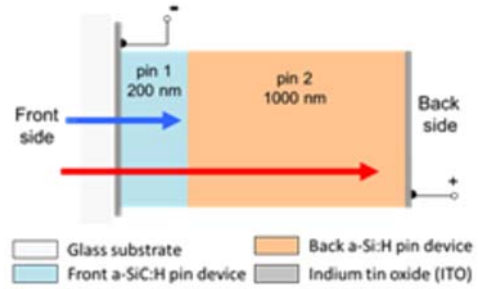


Fig. 3. Device configuration.

The intrinsic absorber materials of the pin devices were designed to confer the ability to detect separately short and long visible wavelengths. This assumption led to the engineering design of both intrinsic layers in what concerns thickness and material optical absorption coefficients, resulting in the choice of a narrow (200 nm) high bandgap (2.1 eV, a-SiC:H) material for the front layer and a thicker (1000 nm) lower (1.8 eV, a-Si:H) bandgap semiconductor for the back structure. Thus, the front p-i-n structure (pin1) exhibits high absorption to the blue light (shorter wavelengths) and high transparency to the red wavelength while the back structure (pin2) high absorption of the red wavelength. In Fig. 4, it is displayed the photodetector transient photocurrent under light produced by the blue, green and red emitters. In this experiment, each emitter was electrically modulated by a rectangular impulse generating a rectangular optical impulse on the device front side. The photocurrent was measured using steady state background lights from the front and back sides and without optical bias.

Results show that under back violet background illumination the gain is high at short wavelengths (blue light) and decreases for longer wavelengths (green and red). The device behaves as a short-pass filter. Under front violet background light, the device works as a long-pass filter for wavelengths above 500 nm, blocking the shorter wavelengths. The shift of the irradiation device side (front/back) allows tuning the short- and long- spectral regions, while the medium region (475 nm – 530 nm) can only be tuned by using both active filters. Under front illumination the red part of the spectrum is enhanced while under back illumination the main enhancement occurs for the blue signal. The quantification of the signal amplification under front and back bias is determined by the optical gain (α_F and α_B for the front and back gains, respectively), defined at each wavelength (λ) as the ratio between the signal magnitudes measured with and without optical bias.

For the red, green and blue light the front optical gain under violet background light is, respectively, 5, 3.5 and 1.3, while for the back optical gains, the values are 0.6, 0.6 and 1.7.

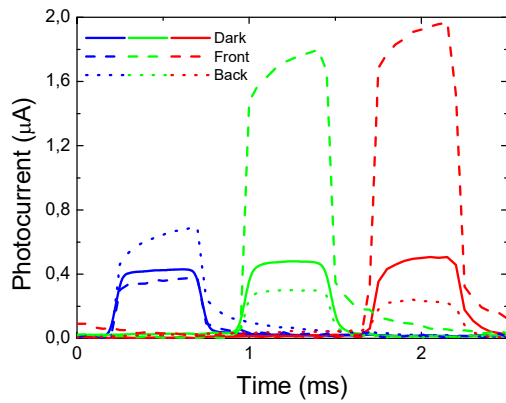


Fig. 4. Photodetector photocurrent measured without optical bias and under front and back optical bias under blue, green and red light of the white LEDs.

3. Results and Discussion

Fig. 5 displays the output photocurrent signals, measured with and without front and back optical violet bias, in a specific position corresponding to the optical excitation from the red and blue emitters at the same frequency and a green signal for identification purposes. The condition assigned when all optical signals are off corresponds to the reference level. The single optical signals are displayed at the top of the figure to help the reader with the different on-off optical states.

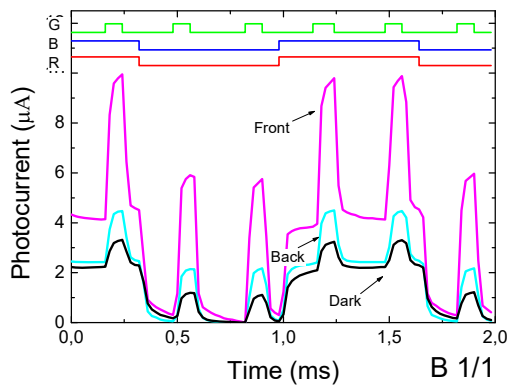


Fig. 5. Photocurrent signal measured without and under front and back optical bias.

In this specific spatial position, the device receives three different optical excitations from the red, blue and green emitters. Under front background light the photocurrent is enhanced compared to the signal measured without background light, as the presence of the red and green light signals induce large amplification. Under back background light, the photocurrent signal is like the one measured without background, as there is a balance between the amplification factor due to the blue signal and the attenuation ($\alpha_F < 1$) from both red and green lights.

Based on the measured photocurrent, we have implemented a computer algorithm to automatically

decode the individual photocurrents assigned to the different input optical signals. We started by extracting the green signal from the photocurrents, which was achieved by sampling the photocurrent signal at the specific time slots where the green signal logic state is unknown and subtracting it to the contiguous time slots where green is known to be absent (Fig. 6).

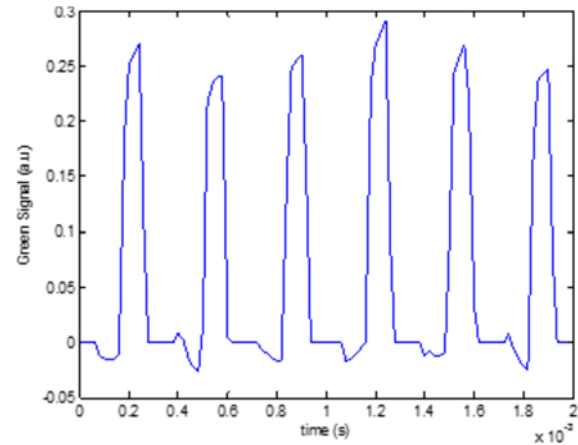


Fig. 6. Extraction of the signal encoded in the green channel.

The second step of the proposed algorithm is to filter out the green signal by simply subtracting the extracted green signal from the photocurrents. The third step takes advantage of the color filtering properties of the sensor. We use front optical biasing to detect green and red wavelengths and back biasing for the blue light. The modulus of the complex Fourier coefficients is computed. Decision about the presence of the frequency in the signal is taken if the evaluated coefficient is higher than a specific threshold. A dedicated algorithm built with MatLab was developed to automatically decode the location position of the detector.

In Fig. 7, it is displayed the Fourier coefficients obtained from the photocurrent signal measured under front and back optical bias. As the output signals can be described by square waves, they contain only components of odd-integer harmonic frequencies. The Fourier analysis shows that the detected frequency is around 750 Hz, which indicates that it is due to emitters located at the top of the unit cell (excited with R and B signals at this frequency). The observed low intensity signal at 1500 Hz is the resultant second harmonic caused by a non-perfect square wave. Results show that there is balance between both Fourier coefficients, which informs about the simultaneous presence of red and blue signals. This is clearly noticeable by the high peaks of the front photocurrent (represented in the graphs by the magenta line), as in the idle bits all emitters are set to ON and this combination results in photocurrent amplification when the device is soaked by steady state illumination from the front side. In contrast, the same signal under back background light is decreased.

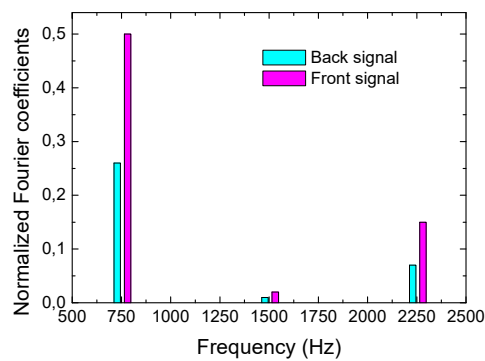


Fig. 7. Fourier coefficients obtained from the photocurrent signal measured under front and back optical bias.

4. Conclusions

A spectral selective photodetector operating in the visible light spectrum was used to measure the output photocurrent supplied by red, green and blue emitters of RGB white LEDs in a visible light communication system. The device, designed specifically, to work as an active optical filter allowed the identification of the optical excitation of each of the emitters, and consequently the spatial position of the detector. Fourier analysis and device optical gains were used to perform the output signal decoding.

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(059)

Nitrogen, Phosphorous and Potassium Detection in Soil Using MEMS Sensor

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Summary: Agriculture as an occupation holds a very important role. The plant growth is affected by the climate and fertility of the soil. We can test the fertility of the soil and improve it based upon the needs of the crop to be grown. Nitrogen, Phosphorous, and Potassium are the major macronutrients found in soil. The use of MEMS-based sensor provides high sensitivity along with compact implementation in optoelectronic devices. A cantilever beam sensor has been modelled in this paper for the detection of NPK nutrients in soil. The cantilever beam sensor is based on the concept of local heating due to the absorption of light by the targeted nutrient.

Keywords: NPK nutrients, MEMS, Cantilever, Sensor, Soil sensor, Absorption, Simulation.

1. Introduction

Understanding the fertilization mechanism is an important part of agriculture. The detection of the concentration of nutrients in soil allows us to plan on the type of fertilizers to be added to the soil. Nitrogen, Phosphorous and Potassium (NPK) are known as the macro-nutrients in soil as their concentration is higher than the rest of the nutrients in soil. The absorption wavelengths of the NPK nutrients in soil are as seen in Table 1 [1].

Table 1. Absorption Wavelength of NPK nutrient in soil.

Nutrient	Absorption Wavelength (nm)
Nitrogen	470
Phosphorous	540
Potassium	640

The MEMS based sensors currently being studied for soil sensing use ISFET with ion-selective electrodes [2]. For development of these electrodes, specific polymers have to be synthesized for detecting each nutrient [4]. Due to this, maintaining uniform sensitivity for all nutrients in a single MEMS sensor is difficult. The MEMS sensor in this paper makes use of a single cantilever beam setup (Fig. 1), which can be reused for sensing all three macro-nutrients, by using a tunable laser which emits light of wavelength equal to the absorption wavelength of the targeted nutrient.

2. MEMS Sensor

2.1. Working of the Sensor

When an atom is impinged by an electromagnetic radiation with wavelength equal to the absorption wavelength of the atom, then the electrons of that atom

are set into vibrational motion. During its vibrational motion, the electrons interact with the neighboring atoms in such a way as to convert its vibrational energy into thermal energy.

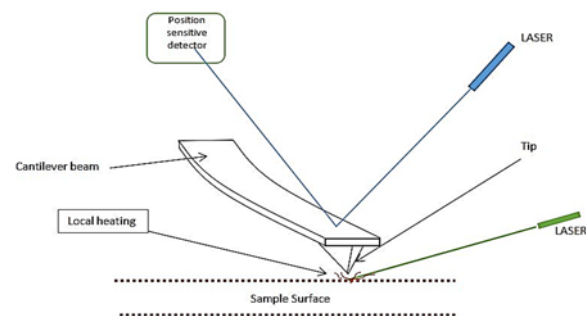


Fig. 1. MEMS sensor model.

We consider a model as shown in Fig. 1. The cantilever beam's tip is kept close to the soil sample's surface. The surface right beneath the beam tip is illuminated by a laser source of wavelength equal to the absorption wavelength of the targeted nutrient. This absorption will lead to vibrational motion of electrons and eventually release thermal energy. The thermal energy radiated near the end of the cantilever will lead to bending of the cantilever due to thermal expansion. The beam will go back to its equilibrium position after it cools down. These oscillations of the beam are noted by using another laser bouncing off the top of the free end of the beam onto a position sensitive detector. These oscillations can be recorded and studied to calculate the concentration of nutrients in the sample.

2.2. Simulation of the Cantilever Beam

The 'Heat transfer in solids' model of the COMSOL software is used to simulate the behavior of the cantilever beam under thermal expansion.

One end of the cantilever beam is kept fixed while the beam end with the tip is kept freely moving. A bimaterial cantilever is used in this model [5]. The model uses Aluminum as the material for the top block and the triangular tip and Silicon Oxide for the bottom block. The 2D cantilever beam model has a length of 900 μm and a height of 50 μm each for both the blocks of the beam. The triangle tip is taken to be 50 μm in height. The fixed end of the beam is kept at 293.15K and the free end is kept at 294K. The temperature gradient along the beam can be seen in Fig. 2.

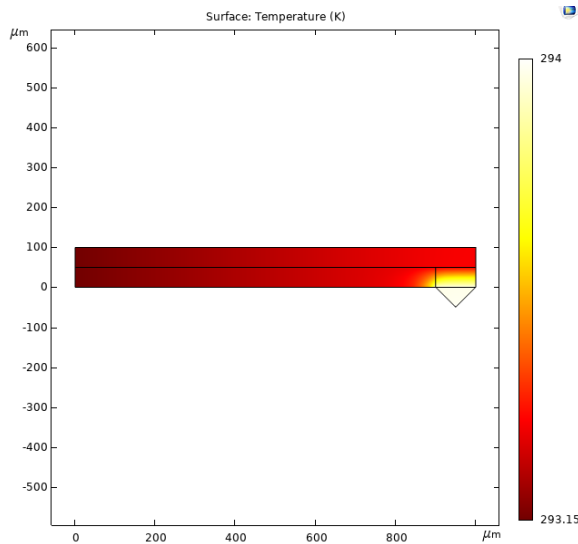


Fig. 2. Surface temperature of cantilever beam.

In Fig. 3, we can see the bending of the cantilever beam under thermal expansion. The one-dimensional graph of Fig. 4 gives us the total displacement of the free end of the cantilever beam as it bends under thermal expansion. We note down the displacement of the free end of the beam to observe the sensitivity of our model [3]. The displacement of the free end of the simulated cantilever beam at different temperature differences between the fixed and free end of the beam can be seen in Table 2.

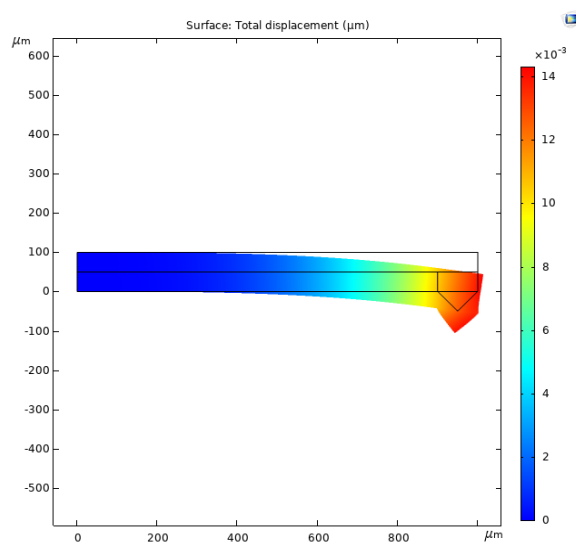


Fig. 3. Total Displacement of beam.

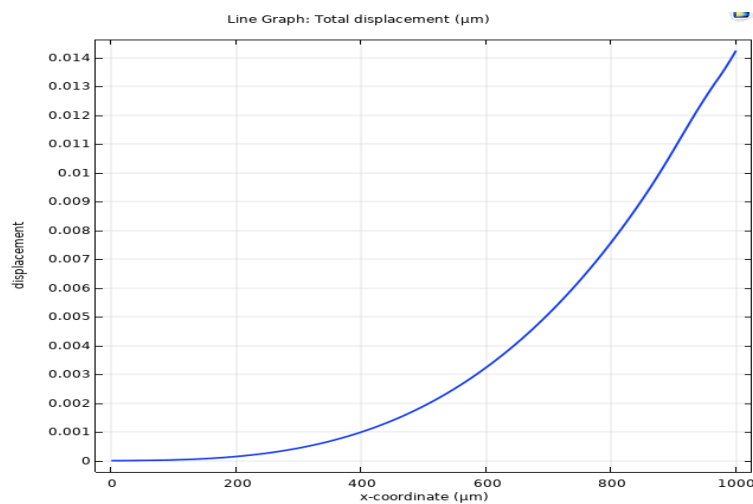


Fig. 4. Displacement plot of cantilever beam.

Table 2. Displacement of cantilever beam for different temperature differences.

Temperature at fixed end (K)	Temperature at free end (K)	Displacement of free end (μm)
293.15	298	0.0804
293.15	294	0.0141
293.15	293.5	0.0058

4. Conclusions

We have designed a MEMS model based on the cantilever beam setup for detection of NPK nutrients in soil. The cantilever beam bends under thermal expansion caused by the absorption of light by the targeted nutrient. The samples with higher

concentration of the targeted nutrient will show higher bending of the beam compared to the case where concentration of the targeted nutrient is low. This model provides an alternative to the use of polymer-based MEMS soil nutrient sensors, with the help of a simple cantilever beam design.

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A Review on Detection of Particulate Matter Using the New Developed Camera-based Optical Sensors

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Summary: Numerous researches and several depth reports have revealed that exposure to particulate matter (PM) can contribute to respiratory, cardiovascular, and cerebrovascular diseases and increase the risk of mortality. As a consequence, ambient air quality regulations and guidelines have been developed by different organizations to restrict the PM allowable concentration. To this aim, there is a considerable demand for the development of particle sensors, characterized by high accuracy and reliability. Traditional detection methods of particulate matter concentration need professional hardware, which is incredibly expensive. However, low-cost, mobile air quality monitoring devices such as optical particle counters (OPCs) have a lack of accuracy. In this framework, researchers have investigated the use of 2D camera sensors instead of single-pixel photodiodes for the development of the optical PM sensors. This paper reviews the recent developments and future trends of newly designed camera-based optical sensors for real-time detection of particulate matter.

Keywords: Air quality monitoring, Camera-based optical sensors, Particulate matter concentration, Real-time detection, Low-cost monitoring.

1. Introduction

Strong growing evidence shows the hazardous effects of ambient particulate matter in the development of chronic diseases from non-accidental cardiovascular to lung cancer [1]. Several studies also have revealed the direct association between exposure to PM and the increasing rate of mortality [1, 2]. Due to these health and even environmental outcomes, governments have defined air quality guidelines and standards regarding the limitation of particulate matter concentration [3].

Recently, a variety of devices based on traditional approaches such as gravimetric technique, microbalance method, and electrical principle for the PM monitoring have been developed. These professional instruments are accurate enough, but they are expensive and bulky [4, 5].

Optical systems based on the light scattering method enable real-time detection of particle concentration using optical particle counters (OPCs) and condensation particle counters (CPCs) [6]. OPC and CPC are able to detect the scattered light from particles and classify them at the minimum size of 50 nm and 2 nm, respectively.

However, the major drawback of the light scattering method is the increasing probability of coincidence at high densities which affects the accuracy of the detection system [7, 8]. For these reasons, it is essential to design an inexpensive and accurate sensor for real-time measurement of the particulate matter.

Currently, the researchers are introducing newly designed camera-based optical sensors using different software and hardware.

Optical particle sensors for air quality have received an increasing attention over the past decade [9]. Nowadays, the newly designed camera-based sensors have been developed by researchers to provide a portable and accurate detection system. This paper presents an overview of the most recent advances in proposed sensors.

To the best of our knowledge, this is the first review to focus explicitly on camera-based 2D particulate matter sensor for air quality detection.

This paper is organized as follows. Section 2 presents the most significant advances in camera-based optical PM technology. A comprehensive comparison of recent sensors structures is discussed in Section 3. Finally, Section 4 gives some conclusions.

2. Newly Designed Camera-based Optical Particle Sensors

2.1. Smartphone-based OPC

Combining the small and low-cost optical dust sensor with a smartphone seems a promising approach to avoid any external and sophisticated platforms for transmitting the measurement data [10, 11]. Generally, the functional principle of the sensor is based on nephelometry. In [12], the particle movement is easily detected by using a 2D matrix sensor instead of a single-pixel photodiode. The particles cross the incident light and illuminate by the collimated LED light of the phone. The airborne dust particles scatter the light in all directions, and the camera records a sequence of images.

It was proposed a magnifier-based method to represent individual scatter traces of the particles. In

the magnifier approach, the camera is not located at the focal point of the optical lens, and therefore the virtual image of the dust particles in observation volume can be obtained [12]. Fig. 1 shows the complete design of the light scattering sensor using a smartphone.

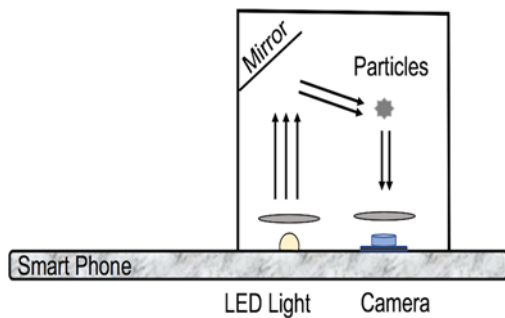


Fig. 1. The third-generation design of the optical sensor [12].

2.2. Filter-based Particle Sensor

Another impressive low-cost sensor, based on a glass-fiber filter and camera can be used for PM monitoring in the urban environment [13, 14]. The particulate matter transported by the airflow goes into the inlet sampling and pass through the filter with the help of a small low power pump (Fig. 2).

Inside the chamber, the filter surface illuminates by a set of multi-wavelength LEDs light. The front camera periodically takes photos to detect the captured particles on the filter.

Finally, image processing software provides the size distribution of the particulate matter from recorded images [15]. The image processing software uses the OpenCV library and performs four processing phases to analysis the captured image.

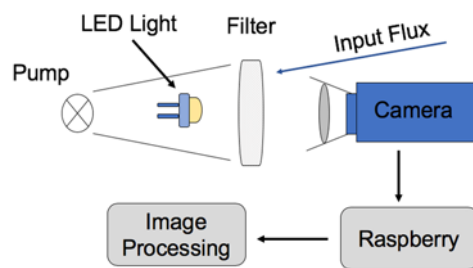


Fig. 2. Block diagram of the filter based particulate matter sensor [14].

2.3. Dust Deposition Camera Particle Counter

A novel approach, based on CMOS image sensor using image processing techniques, classification algorithms and pattern recognition, is proposed [16, 17].

In the proposed sensor, the dust directly deposits on the CMOS image sensor (without any lens). The sensor was placed at 45° angle towards the ground to restrict

the overlapping due to particle accumulation [16]. An LED lamp light illuminates the sensor surface to obtain high-contrast photos in different environmental conditions.

The control software allows taking a picture periodically and saves it as an uncompressed JPEG. Image processing algorithms enhance the captured images characteristics (e.g. contrast). Consequently, the particle size and shape can be obtained from preprocessed images by means of classification algorithms. Fig. 3 shows the image acquisition system.

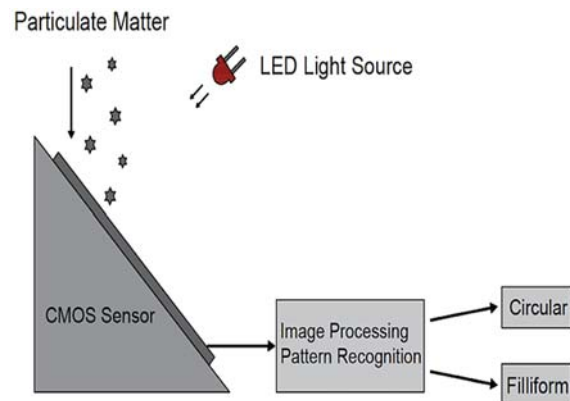


Fig. 3. Dust sensor setup for based on deposition on the CMOS camera [16].

2.4. Holographic Particle Counter

A new holographic particle counter (Fig. 4) can determine the number concentration of the aerosol particles in the lower micrometer-sized [18-20].

As discussed before, one of the significant limitations of the commonly used light scattering based devices is the rise of the probability of particle coincidence in the high-density area [21]. As an attempt to fill the gap, a 2D camera sensor instead of a single photodiode can be employed to detect the full 3D information of sensing volume.

In the particle imaging unit, every single particle goes into the sampling channel and exposes to the low power diode laser beam. It diffracts the incident beam and creates the circular interference fringe pattern in the detection plane. The particle number concentration can be obtained using pattern recognition on the recorded images [22].

Fig. 4 shows the complete geometry of the holographic particle counter. The proposed setup can attach to the Condensation Nucleus Magnifier (CNM), which works based on the principle of magnifying the small particles before exposure to the laser light. Therefore, the system is intended as an acceptable alternative to condensation particle counters (CPCs).

The benefit of this configuration is that it does not require a narrow inlet for counting each particle in the observation volume, and it is independent of the speed and direction of the particles.

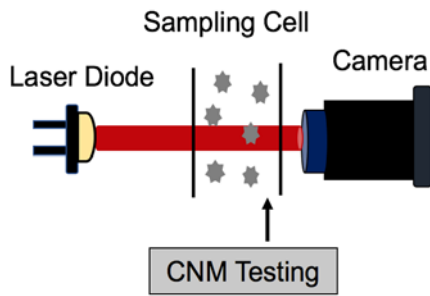


Fig. 4. Holographic optical counter [19].

3. Comparison and Discussion

Tables 1 and 2 show the measurement specifications of the newly designed camera-based optical particle sensors for air quality monitoring. In particular, the Table 1 compares the typical detection range of the PM based on functional operation, the different structures, and other limitations of the sensors.

In the inhalable coarse particles (PM_(10-2.5)) range, the landscape of ultra-low-cost instruments is

dominated by miniaturized camera-based OPC. The accuracy of the detection system almost depends on the phone model. Besides, the proposed sensor can be quickly attached and removed from the phone.

The low cost and simple design COMS image sensor is able to provide detailed information on the shape of the particles in addition to their size. In this framework, it is possible to identify dust origins. It should be noted that the sensor can be efficiently used for the fixed indoor environment such as museums or cleanrooms.

As is evident, the mobile filter-based sensor using the wide wireless connectivity seems a promising candidate for low-cost PM monitoring. However, the sensor still needs some improvements in terms of the sampling system, detection size range and dimension.

Among the others, holographic particle counter seems a compelling alternative to the condensation particle counter (CPC) for determining the airborne particle number concentration of lower than μm -rang. The sensor enables not only accurate measurement of the low particle number concentration but also allows determination of higher particle densities.

Table 1. Comparison of camera based PM sensor.

Sensor Type	Camera-based OPC	Holographic based sensor	Filter based sensor	CMOS sensor
Working principle	Light scattering + Image processing	Holography + Image processing	Image processing	Image processing
Size fraction	PM _{10-2.5}	PM above 3-4 μm after condensation	PM above 3 μm	PM above 3 μm
Minimum Resolution	–	Single frame: 4.16 ($\frac{\mu\text{m}}{\text{ccm}}$) 20fps: 0.208 ($\frac{\mu\text{m}}{\text{ccm}}$)	–	–
Detection system	Size detection / Mass concentration	Number Concentration	Size detection / Mass concentration	Size detection / Shape
Reference system	SMPS/ TSI APS	TSI-3775	Commercial particulate monitoring station	–

Table 2. Comparison of components of novel sensors.

Sensor Type	Camera-based OPC	Holographic based sensor	Filter based sensor	CMOS sensor
Light source	Smartphone flash LED	Laser diode (635nm)	3RGBLED 1 IR LED 1 UVLED	White LED
Detector	Smartphone camera	UI-3082SE-M Camera IDS (COMS sensor)	RPI NOIR Camera V2	Life Cam HD-3000 (COMS sensor)
External Lens	Large magnification lens	-	Macro Lens 15×	Without lens segment of camera
Dimension	Miniaturized	Small (Without CNM)	Small	Small

4. Conclusions

Hazardous effects of particulate matter in human health and environment has stimulated growing demands for developing the low cost, portable and reliable PM monitoring sensors.

However, the mobile PM detection sensors are a much less known area compared to significantly numerous reports of the classic air monitoring techniques. This paper reviews the development and current status of newly designed optical camera-based sensors for real-time PM monitoring.

The proposed techniques appear very promising for the development of the PM air monitoring sensors, but there are still some issues with the increasing of the detection size of the particles as well as their accuracy.

Since demands for such a system considerably increase, we can estimate that new and more accurate devices based on new structures and processing methods may appear in the near future.

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Novel, VOC-free-renewable Solid-state Biopolymer Detector for Mid-to-extreme Radiation Field Monitoring

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Summary: A novel, VOC-free-renewable solid-state biopolymer detector is presented for applications pertaining to mid-to-extreme gamma-electron beam dosimetry applications. Polylactic-Acid (PLA) based “green” biopolymer was used to derive PLAD – a renewable, inexpensive solid-state radiation detector that promises to offer significant potential benefits over when compared with state-of-art detectors for detecting gamma-electron and also other forms of ionizing radiation. Experimental data are presented on well-correlated measurements of ionizing radiation in the mid-to-extreme dose range to readily discernible physical property changes such as relative viscosity, embrittlement, and molecular bonding.

Keywords: Polylactic Acid, PLA, PLAD, Radiation, Gamma, Electron, FTIR, Detector, Dose.

1. Introduction

Nuclear safety is of paramount concern in world-wide nuclear and radiation based facilities – especially, for those handling and reprocessing involving extreme radiation doses, and esp. for actinides (e.g., Pu/Am/U) that are also high-impact alpha-spontaneous fission neutron emitters; as also, medical and sterilization facilities deploying high energy photon and e-beam accelerators. As such, nuclear radiation instrumentation for safety and optimal functioning of such facilities with varying missions is always an important factor [1].

For the past over 80+y, radiation monitoring technology has remained largely the same [2-4] – relying primarily on sensor technologies which require monitoring for the tell-tale charge buildup in ionized gases/solids (e.g., fission chambers using HEU and compensated ion chambers), or monitoring of light flashes from scintillation or thermoluminescence. At the high cost end, complex radiation spectrometers can cost in the \$M range (as may be deployed at high powered accelerator driven spallation sources or research reactor facilities and requiring skilled scientific staff), down to a range of \$1-10K for portable survey meters, and even at as low as \$10 for commonly used personnel dosimeters (e.g., TLDs) the aggregate personnel monitoring costs for complex-wide needs can rise to several \$M/y. In addition, present day nuclear instrumentation is varied in construction and deployment according to its application; i.e., different individual systems for neutron/gamma/beta/alpha /fission radiation detection.

This paper provides results from scoping studies that address the need for improvements to present-day nuclear instrumentation by means of development and testing of a novel, ultra-low cost, solid state polylactic acid (PLA) bio-polymer based nuclear detector instrumentation referred to herein, as PLAD.

2. Methodology & PLAD Sensor Overview

At Purdue University (PU), we have researched [5-7] the potential for inexpensive, renewable (corn-soy derived) “green” PLA-based polymers to be used for radiation detectors we call “PLAD”. PLA is a biopolymer (derived from dextrose sources such as corn/soy/...) that appears to offer significant potential benefits as a detector for various forms of ionizing radiation, as well as for dosimetry. Fig. 1 shows a typical commercially-purchased PLA (MW ~10⁵) resin bead (~2-3mm OD) from Natureworks, LLC. Effectively, each such resin bead serves as a solid-state ionizing radiation detector.

Q) What may be so compelling for PLAD as a potential radiation detector? It is well-known that ionizing radiation interaction will produce atom dislocations and other effects in virtually every material to varying extents for example via neutron irradiation of steels leading to embrittlement. However, such dislocations and property changes are not readily discernible for monitoring in real-time using commonly available devices and techniques. What is needed is a material that responds well even in harsh environments to varied forms of ionizing radiation to produce well-correlated property changes to simple physical properties (e.g., density, relative viscosity, hardness, molecular weight, etc.) that are amenable to rapid-fire and cost-effective measurements using common laboratory equipment. This is where PLAD technology appears to show significant potential. Successful implementation may provide for ultra-low cost (10-100× lower than even TLDs), safety-security-environment friendly alternate to expensive nuclear instruments such as HEU-fission chambers and compensated ion chambers.

3. Results

Scoping studies with gamma/electron radiation reveal exciting possibilities such that PLADs may find use in high mixed field neutron-gamma-beta radiation monitoring in near-real time via correlation of absorbed incident nuclear radiation to changes in key physical properties such as hardness, relative viscosity (RV), friability (F), heat of fusion (Hf), melting temperature and FT-IR based molecular bonding. Some of these results are presented herein. Details are presented elsewhere [5-7].

Fig. 2 presents results of impact of Co-60 generated (1.2-1.3 MeV) gamma dose on PLAD's changes in RV. Experiments were conducted using PU's Gamma-Cell irradiator. RV is a readily determined parameter (involving dissolution and noting of time to dispense a set quantity relative to that of a baseline solution). As noted therein, a distinct first order linear variation is observed for doses ranging from 0 kGy to about 40 kGy (40,000 Rad). A second linear RV vs Dose variation is noted for >40 kGy dose.

Fig. 3 presents results for irradiation dose versus grinding efficiency (GE). Once again, irradiated resin pellets can be simply ground in commercial (off-the-shelf) grinders and sieved. Once again, the variation of GE vs dose is seen to remain linear over 0-100 kGy (0 – 100,000 Rad) ranges which is relevant for high dose environments such as encountered commonly for medical sterilization and also in the vicinity of spent nuclear fuel (SNF) wherein the radiation field intensity can readily range $\sim 10^4$ R/h even 2y post-shutdown [3].

Fig. 4 depicts percent change in normalized FT-IR signals for peaks of interest in irradiated samples. Horizontal axis represents dose in kGy. An interesting trend is noted for the COC-to-COH bonds of PLA and appears to present a useful metric for monitoring for radiation dose. Similar results have also been observed with e-beam radiation [6].

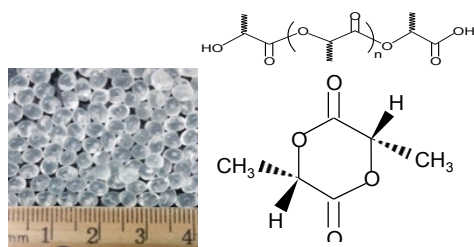


Fig. 1. PLA resin beads (Natureworks, LLC) and molecular structures (repeat & dimer) [6].

4. Conclusions

PLAD response to mid-to-extreme gamma-electron dose correlates nicely with RV, GE and FTIR molecular bond variation; other parameters such as melting temperature as well as melting temperature and material hardness also appear to provide first-order monitoring enablement to varying extents for change

correlated dose [6]. Future work will include the integral effects of neutron-alpha radiation effects.

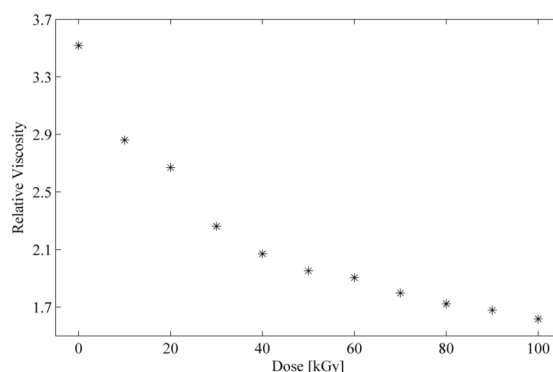


Fig. 2. Variation of RV vs Irradiation Dose [6].

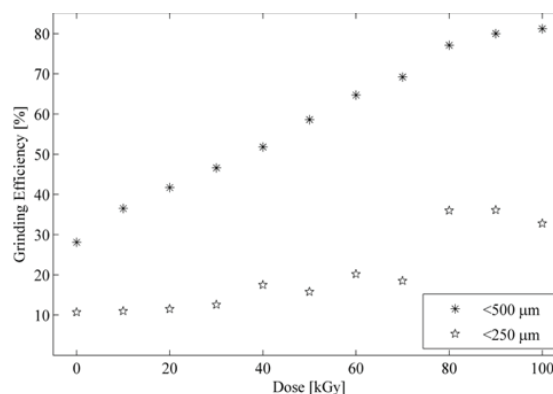


Fig. 3. Embrittlement/grinding efficiency (GE) vs Dose [6].

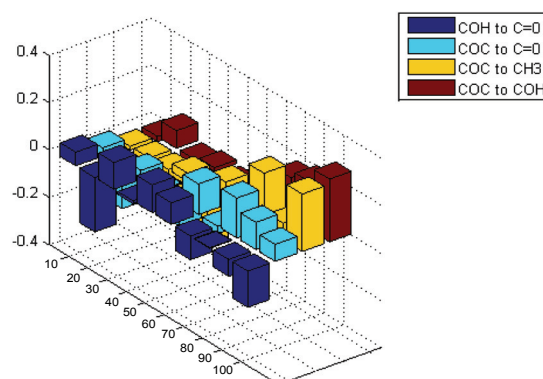


Fig. 4. FT-IR peak variations with bond type & dose [7].

Acknowledgments

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Optical Solutions for Estimation of the Particulate Matter Distribution from Internal Combustion Engine

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Summary: The diesel cars are equipped with a particulate matter filter to reduce the emissions from vehicle exhaust. However, several researchers have reported high filter failure rates. In this framework, regular checking of the vehicle emission by the use of cheap and simple test equipment is essential. Among different sensor technologies, optical particle counters (OPCs) are widely used for real-time, mobile detection of the aerosol particles from a low concentration environment (e.g., clean room) to highly concentrated urban areas. Since their detection accuracy mostly depends on the monitoring of single particle scattering, decreasing the probability of particle coincidence in observation volume is necessary. The proposed work relates to the design of new different layouts, including low-cost optical particle counter for real-time detection of the particulate matter concentration after vehicle tailpipe. The main objective is to detect particulate matter emissions based on single-particle assumption to achieve very low coincidence errors. The final layout shows that the optical sensor can obtain acceptable accuracy by the use of four ejectors in parallel and one compressor.

Keywords: Exhaust emission, Optical particle counter (OPC), Coincidence error, Ejector, Particulate matter detection, Real time detection.

1. Introduction

Evidence from the last decade shows that particulate matter (PM) pollution is one of the leading public health concerns since it increases the risk of mortality [1]. The characteristic of this atmospheric particulate matter can vary considerably depending on its origin, i.e. natural (e.g., volcanic activities) or anthropogenic (e.g., combustion of fossil fuels). Particulate emissions from fuel combustion by internal combustion engines (e.g., diesel engines) represent one of the most significant problems in ambient air pollution.

Diesel vehicles are equipped with highly efficient diesel particulate filters (DPF), which reduce the emissions of soot particles. However, the removal of the DPF and the broken or manipulated filter can significantly affect the PM emission. In this framework, it is essential to recognize such big polluters.

The on-board diagnostic check is not a reliable option for the detection of the performance or filter faults. One way to accomplish this task is the periodical roadworthiness check [2] of the emissions after tailpipe using low weight and cheap equipment. To this end, the selection of a suitable particle sensor is a necessity.

There are numerous devices based on different working principles (e.g., gravimetric method, microbalance method, high-frequency acoustic techniques, etc.). However, the available measurement systems are typically complex, expensive, bulky, and stationary [3].

In the last decades, low-cost, light-scattering-based optical particle counters are attracting increasing

attention because of their cost and ability to provide real-time measurement [4]. Although, one of the main problems of these devices that affect the accuracy of the measurement system is the coincidence error. In principle, coincidence error is caused by the simultaneous presence of several particles in observation volume [5]. In order to handle this issue, the single-particle assumption is mostly used for all optical particle counters.

In this work, the authors show a design of the straightforward and portable layout, using low-cost OPC after vehicle tailpipe to obtain very low coincidence errors by the single-particle assumption.

It is, therefore, of crucial importance to increase as much as possible the sensitivity in the detection of particulate matter concentration from diesel exhaust vehicles. Therefore, to enhance the detection accuracy and to decrease the coincidence error, different solutions were suggested. The final layout, with four parallel ejectors and one compressor, seems to be the best configuration for high precision detection by OPC.

In the case that the vehicle emission is at a higher concentration, there is the ability to increase the dilution factor by the use of more ejectors in parallel.

The paper organizes as follows. Section 2 describes the general principle of the optical particle counter and mains solution for the use of OPC after vehicle tailpipe. Section 3 presents the first sensor layout by reduction of sensing volume of OPC. Section 4 shows a layout including one ejector after exhaust to detect the PM emission, whereas Section 5 suggests the use of several ejectors in parallel to increase the dilution factor. Finally, some conclusions draw in Section 6.

2. Designing the Layout using OPC after Tailpipe

The Mie theory explains the theoretical solution of Maxwell equations for spherical particles that interact with electromagnetic waves. Generally, the intensity of the scattered light depends on the wavelength, the particle diameter, the reflective index and the scattering angle [6].

The measurement principle of the optical particle counters is typically described by the scattering theory (e.g. Mie theory or Ryleigh theory). The particles cross the laser light beam and scatter the incident light in all direction. The scattered light detected by the photodetector convert to the current and amplify by the use of OPC electronics.

The main limitation of these counters is the increasing probability of coincidence of consecutive particles at higher particle density. In low-cost optical sensors, there is no real knowledge about the number of particles present in observation volume. In this case, the sensor can detect the total sum of the light scattered intensity from N particles, instead of single particle detection.

In diesel engines vehicles, the DPF filter decreases the concentration of particulate matter emission by two orders of magnitude [7]. However, the tailpipe PM concentration is still too high to avoid the probability for the occurrence of coincidences in the sensing volume of the optical particle counter.

Reducing the observation volume or diluting the airflow in the inlet to the counter are two primary solutions to decrease the coincidence error [8]. These are the main ideas for designing the sensor for detection of the exhaust PM vehicle by the use of OPC.

3. Reduction of the Observation Volume of the OPC (First Layout)

There is a variety of OPC with different detection range accuracy in the market. Among several low-cost dust sensors, the OPC-N3 (AlphaSense company) [9] is selected for its sufficient capability to monitor the accurate environmental PM concentration at a reasonable cost. However, to the best of our knowledge, there is no literature study in which the optical particle sensor is used in such a way to obtain a real-time measurement of the PM vehicle concentration.

For the design of the proposed sensor, as a first solution, we consider the reduction of observation volume for diminishing the coincidence error. Generally, the maximum measurable concentration is limited by the minimum size of the sensing volume that can be realized [5].

The requested reduction of the sensing volume of the sensor can be achieved by narrowing the inlet of the sensor. In fact, the reduction of inlet sampled airflow passed through the observation volume should guarantee the presence of one particle in the sensing cell at a time.

It should be noted that the maximum number concentration of the diesel engine [7] for the minimum detection size of the OPC-N3 is about 500 ($1/cm^3$).

Therefore, the required size of observation volume, considering the single-particle assumption, can be achieved with an inlet duct diameter of 0.121 cm.

However, the main concern is that it is challenging to construct such a small inlet duct that satisfies the requirements.

Fig. 1 presents the relationship between the concentration of diesel engines emission and the diameter of the inlet sampling for the measurement of only one particle in the cylindrical sensing volume. For the defined sensing volume, the diameter of the laser is considered 0.5 mm bigger than inlet duct diameter.

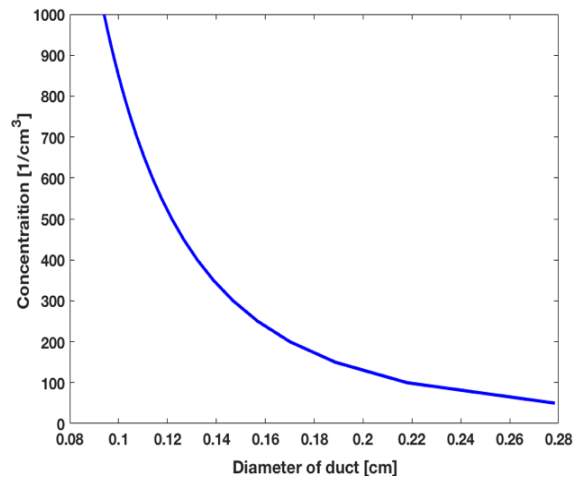


Fig. 1. Number concentration vs diameter of the inlet.

4. Dilution System before OPC (Second Layout)

The proposed layout adopts an ejector dilutor [10] before the OPC to decrease the concentration of the sampled exhaust emission.

The exhaust gas from vehicle tailpipe flows through the ejector by the use of a steel tube. Besides, the adoption of the appropriate filter before the ejector eliminates all airborne particulate matter in the environmental air. To obtain a constant emission flow rate in the mixing zone, the use of a single head pump before the fluid part of the ejector is necessary. Figure 2 shows the complete design of the proposed system after the vehicle tailpipe.

Using the characteristic curve of the ZH05DSA-06-06-06 ejector, the maximum dilution factor (defined as the ratio of airflow rate and exhaust flow rate) at a pressure supply of 0.6 MP is about 1.93.

The real and diluted concentration can be defined by the following formula:

$$C_d = \frac{C_r}{df} \quad (1)$$

where C_d is the concentration of the diluted emission, C_r is the real concentration of the exhaust emission, and df is the dilution factor.

Technical specification of OPC-N3 shows the maximum coincidence probabilities of 0.84 and 0.28, and maximum particulate number concentrations of 1000 ($1/cm^3$) and 0.500 ($1/cm^3$) respectively.

In this framework, for the real number concentration of 500 ($1/cm^3$), the proposed dilution factor is still not sufficient to have a low coincidence error by the OPC (with respect on what recommended in the data sheet).

To this aim, it is discussed the design of the second layout in Section 2.3.

5. The Third Layout Detection System

Fig. 3 shows a different layout, including more ejectors in parallel to decrease the number concentration of particulate matter to the feasible value of around 60 ($1/cm^3$) that can be obtained by the dilution factor of 8 (eq.(1)).

The environmental air flows through all ejectors except the last one. The fraction of the raw tailpipe emission sucks by the suction part of the last ejector. In the proposed layout, the OPC fan can adjust the flow rate to the appropriate value by connecting the sensor indirectly to the main tube.

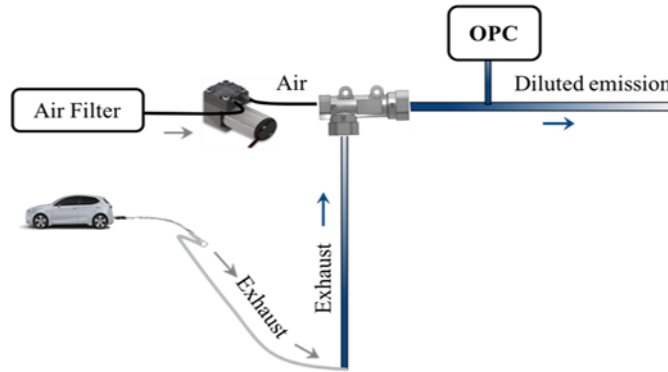


Fig. 2. Total design for monitoring of PM after vehicle tailpipe using a low-cost optical particle counter

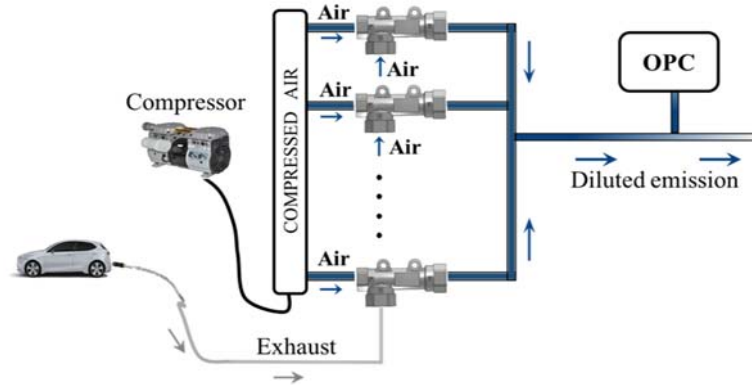


Fig. 3. Detection system using parallel ejectors.

Being Q_{supply_air} the volumetric flow rate of the supply air, $Q_{suction_air}$ the flow rate of the suction air and $Q_{exhaust}$ the flow rate of the exhaust gas in the suction part of the last ejector, the dilution factor calculate using the given formula:

$$DF = \frac{\sum_{i=1}^N Q_{supply_air} + \sum_{i=1}^{N-1} Q_{suction_air}}{Q_{exhaust}} \quad (2)$$

Finally, the dilution factor by using the four ejectors in parallel and choosing the supply pressure of 0.5 MPa is about 8.15. The volumetric flow rate of the mixed emission and proposed compressor can be

determined as a value of 69.66 (l/min) and 49.73(l/min) respectively.

6. Conclusions

This work shows the feasibility of a low-cost layout, including an optical particle counter, to measure particulate matter concentration after vehicle tailpipe. In particular, a layout with four ejectors in parallel and one compressor has been designed to dilute the vehicle tailpipe emission exhaust. In the proposed setup, to decrease the maximum

concentration of the exhaust emission to the value of 60 ($1/cm^3$), it can be chosen a dilution factor of 8.

It is worth to note that, since the OPC-N3 minimum detection size is about $0.35\mu m$, the system can only measure part of the accumulation exhaust particles. Future works will be carried out to enhance the detection accuracy and to calculate the temperature and humidity of the mixture emission before the OPC inlet.

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Sensor Fusion for Accurate Human Body Temperature Measurement at a Distance

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Summary: Thermal cameras can be used to detect radiating heat from the human body. The human body temperature is usually measured from the forehead surface. Due to their inferior accuracy when compared to traditional hand-held thermometers, thermal cameras are not designed for medical use. The present work proposes a novel concept device that combines generic thermal imaging with LiDAR data and environmental temperature and humidity for an accurate human body temperature measurement. First, the LiDAR estimates the distance to the subject, then employing a computer vision pipeline to fuse visible spectrum stream and thermal stream, the human forehead temperature is detected. Experimental results show that the temperature measured from about 60 cm (2 feet) from the subject is not significantly different than the temperature obtained using a medical thermometer.

Keywords: Thermal camera, LiDAR, Sensor fusion, Computer vision, Deep learning, Elevated body temperature.

1. Introduction

At present, the majority of the human body temperature measurement devices are accurate only if the measurement is made through touching the subject or getting at 10-20 millimeters of the human body. However, manual handling of a medical thermometer is only possible with a dedicated human operator. For mass screening of population, the medical-style temperature measurement is non-practical [1]. The non-invasive detection of human body temperature has seen a peak interest after 2003, when severe acute respiratory syndrome (SARS) virus affected the planet [1]. In the context of the new pandemic caused by Coronavirus (COVID-19), the accurate measurement of human body temperature at a distance becomes a key technology for mass health screening.

Thermal cameras are used in many industries for different objectives, such as automotive, security, thermal insulation inspection, battery charging process monitoring, etc. However, the generic thermal cameras are not the mainstream solution for human body temperature measurement in medical scenarios [2]. This is valid especially for low-cost thermal sensors, where the accuracy error margin for the detected temperature can be as high as plus minus 5 % [3].

Usually thermal cameras provide digital images with the highest values of intensities corresponding to the hottest temperatures in the scene. Detecting fever of the human body in unconstrained environments was investigated in [4], where classical computer vision techniques are used to separate the high intensity pixels from the thermal camera corresponding to the human face after calibrating the algorithm using a black body with a known temperature.

Thermal camera and depth map generating devices are employed in [5] to measure skin temperature with the aim of obtaining a thermal footprint of a person in indoor environments. A thermal camera system is proposed in [6] for continuous body temperature measurement using computer vision for face detection and object tracking. The temperature measurements obtained in [6] are reported to have a low root mean square error (RMSE) compared a medical thermometer.

To overcome existing pitfalls inherent with available low-cost thermal cameras employed in human body temperature measurement, the present work aims to build a reliable device for reduced constraints temperature measurement. The scientific and practical contributions of the present work are the following:

- Using low-cost, off-the-shelf sensors for building the human body temperature measurement device;
- Fusing information from the different sensors in a robust software architecture for accurate temperature measurements at 40 cm – 60 cm;
- Automatizing the information flow through a computer vision pipeline that outputs the temperature from the human forehead without a human supervisor.

The remainder of the paper is organized as follows: Section 2 describes the hardware components employed to build the device. The proposed measurement methodology is depicted in Section 3 and the experimental results are reported in Section 4. Further enhancements to the proposed device using deep learning are described in Section 5 and the conclusions are drawn in Section 6.

2. Hardware Design

2.1. Employed Sensors

First, a generic, low cost thermal camera is employed. The thermal camera that is employed is a FLIR One Pro [1], with dedicated connection for mobile phones via USB-C or micro USB. A fan cooling system has been attached to the camera to avoid the interference of high temperatures due to continuous operation. This can be observed in Fig. 1.

For accurate distance measurement to subjects a LIDAR-Lite v2 from PulsedLight has been used. The LIDAR sensor connects via Inter-Integrated Circuit

(I2C) communication protocol as a slave and measurements are requested by the main microcontroller unit (MCU), in a separate process, every 0.02 seconds. The measured distance has an accuracy of +/- 2.5cm.

Environmental temperature and humidity inputs are being provided by the HTU21D digital sensor on the same I2C interface. The sensor can gather temperature readings from (-40 °C) up to 125 °C and humidity readings in the 0-100 % range.

All peripherals are connected to a Raspberry Pi 4 Model B/4GB, running Debian OS 10.0.

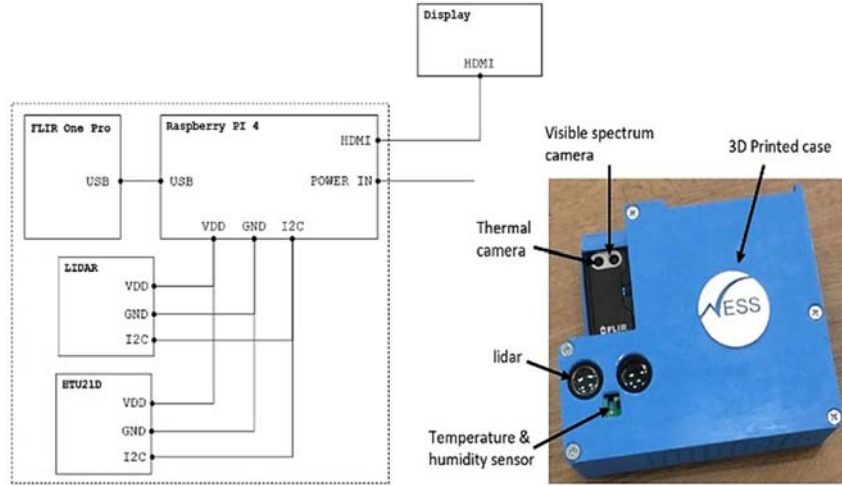


Fig. 1. Device Design: Left: Multi-sensor connectivity: PI4 – Raspberry Pi 4, HTU21D – Sensor for environmental temperature and humidity; Right: final device.

2.2. Thermal Imaging to Raw Temperatures

In the thermal image obtained from the FLIR One Pro camera each pixel corresponds to an infrared radiation spot. This radiation comes from the subject that stays in front of it, but also from the environment that emits indirect radiation or attenuates radiation components as the user moves towards the sensor [7]. The camera manufacturer provides a calibration procedure that takes into account 3 factors [8]:

1. Emission from the object, $W_{ob_emitted} = \varepsilon \cdot \tau \cdot W_{obj}$, where ε is the emissivity of the object, τ is the transmittance of the atmosphere;
2. Reflected emission from ambient sources is equal to $W_{re} = (1 - \varepsilon) \cdot \tau \cdot W_{amb}$, where $(1 - \varepsilon)$ is the reflectance of the object;
3. Emission from the atmosphere, $W_{atm_emitted} = (1 - \tau) \cdot W_{atm}$, where $(1 - \tau)$ is the emissivity of the atmosphere.

The final radiation spot input for each thermal pixel is calculated using the sum of the above factors as follows:

$$W_{tot} = \varepsilon \cdot \tau \cdot W_{obj} + (1 - \varepsilon) \cdot \tau \cdot W_{amb} + (1 - \tau) \cdot W_{atm} \quad (1)$$

During calibration the values taken by the above parameters are $\varepsilon = 0.97$ object emissivity human bodies, W_{amb} and W_{atm} depend on τ the transmission through the atmosphere factor. τ is calculated using d , the distance to subject, T_{atm} , temperature of the environment fed by the local sensors and relative humidity readings from the sensor. The FLIR calculations are done using the Waldemar [9] and FLIR Toolkit [10] equations below:

$$\tau(d, \omega) = K_{atm} \cdot \exp[-\sqrt{d}(\alpha_1 + \beta_1 \sqrt{\omega})] + (1 - K_{atm}) \cdot \exp[-\sqrt{d}(\alpha_2 + \beta_2 \sqrt{\omega})] \quad (2)$$

$$\omega(\omega\%, T_{atm}) = \omega\% \cdot \exp(h_1 + h_2 \cdot T_{atm} + h_3 \cdot T_{atm}^2 + h_4 \cdot T_{atm}^3) \quad (3)$$

As described by Quang et al. [7], the scaling factor for the atmosphere damping is fixed at $K_{atm} = 1.9$, attenuation for atmosphere without water vapor ($\alpha_1 = 0.006569$, $\alpha_2 = 0.012620$), attenuation for water vapor ($\beta_1 = -0.002276$, $\beta_2 = -0.006670$), and $h_1 = 1.5587$, $h_2 = 6.939 \times 10^{-2}$, $h_3 = -2.7816 \times 10^{-4}$, and $h_4 = 6.8455 \times 10$. All values were taken from the thermal image file, from the metadata overhead.

The influence of atmospheric humidity as well as the camera-to-object distance can be observed in Fig. 2 and errors highlighted in [11] and [12] show the importance of accommodating for external factors from (2) and (3) which cannot be disregarded as such.

All raw temperature readings from the FLIR one driver rely on dynamic inputs from the LIDAR and the environmental temperature and humidity sensors. To achieve this, separate processes have been assigned to each sensor and all readings are gathered by the thermal image process for final calculations. Data is collected using ZeroMQ concurrency framework.

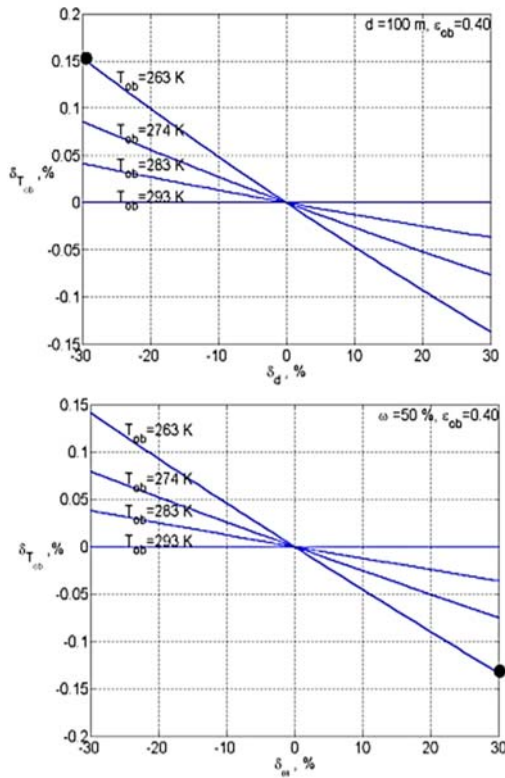


Fig. 2. Top - Influence of camera-to-object distance d setting error on object temperature T_{ob} measurement error at, $d = 100$ m and $\omega = 50$ %, Bottom - Influence of atmospheric relative humidity ω setting error on object temperature T_{ob} measurement error [11].

3. Proposed Measurement Methodology

At the core of the proposed human body temperature measurement device is a computer vision framework that organizes the flow of information from various sensors, as seen in Fig. 3. The video streams that are provided by the FLIR One pro camera are not identical in resolution. The visible spectrum stream has a resolution of 1440×1080 pixels with 3 channels and the raw thermal stream has a resolution of 120×160 pixels, monochrome.

3.1. Computer Vision Pipeline

The video stream produced by the visible spectrum camera is initially employed to detect the face of the

person. For face feature detection, the popular Dlib library [13] is employed. Dlib library, originally proposed in 2009 [14] has two main components: a linear algebra component for fast feature computations and a machine learning component. The version of Dlib that was employed was trained to detect 69 facial landmarks.

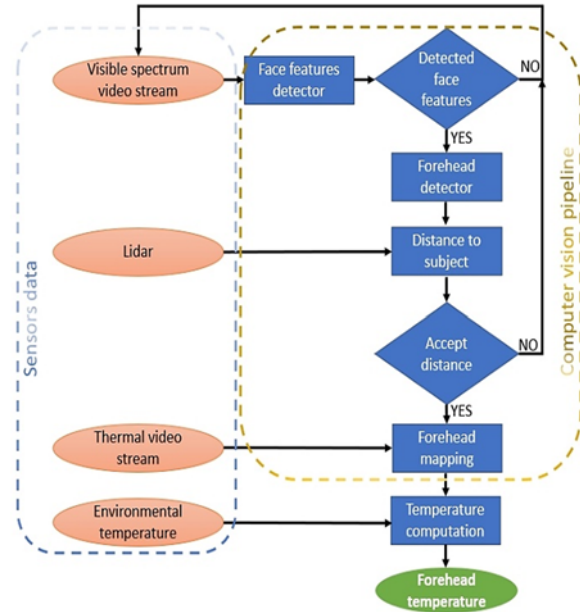


Fig. 3. Information flow for the proposed device.

Among the facial landmarks detected by the Dlib library, the eyebrows were used to detect the forehead. The line that joins the 2 eyebrows was considered as the bottom of the forehead. A rectangle that is centered above the line will represent the forehead, or region of interest (ROI). The location of the ROI is subsequently mapped into the thermal stream by applying heuristically image alignment techniques between the 2 frames.

After the distance to the subject is measured using the LiDAR sensor, and is confirmed to be in a pre-defined range, the maximum temperature of the ROI in the thermal stream is fed as input into an equation for determining the accurate temperature.

4. Experiments

The experiments were organized according to a protocol that can be employed in generic workplaces, such as hospitals, IT companies (Fig. 4), factories or institutions. The participants were asked to stop in front of the device on lines marked on the floor at distances of 40 cm, 50 cm and 60 cm. A single participant shall use the device in a single moment in time, i.e. no other human will step into the field of view of the thermal camera. The individual will then wait for the device to measure the forehead temperature at each of the 3 distances. Afterwards, the forehead temperature is measured using a medical thermometer. The 4 temperatures corresponding to an individual are

recorded. No personal identification information is stored for a person, only an ID number for experiments.

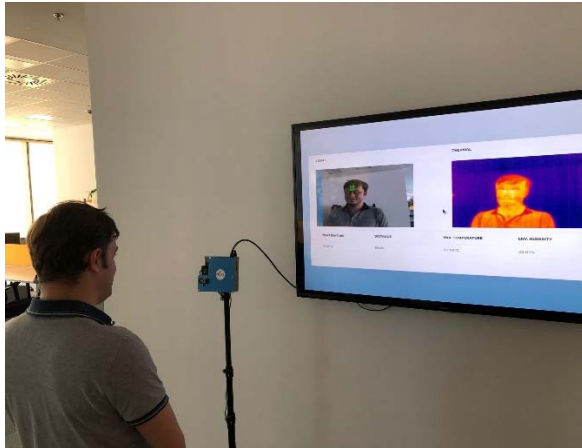


Fig. 4. Proposed device deployed at Ness office in Timisoara.

For performance measurement, the RMSE was computed between the distributions corresponding to the temperature measurements made by the device at the 3 distances and the distribution composed of the temperatures obtained from the medical thermometer. The properties of the 4 distributions for 10 individuals can be observed in the Table 1 below.

A statistical significance test was performed, with the null hypothesis being that the temperatures measured by the proposed device are from a different distribution than the temperature measured by a medical thermometer. The values obtained for the 3 distances are considered to be drawn from normal distributions, with the means and standard deviations computed from the temperature samples. A t-test was performed, that allows a comparison between the average values of the two data sets and determine if they belong to the same population. In the initial experiments, the number of degrees of freedom for the t-test is equal to the double of the number of the individuals minus 2. The t-values obtained for the 3 distances below are 1.08 for 60 cm, 0.47 for 50 cm and 0.16 for 40 cm. Setting a significance level of 0.05, from the statistics table the t value read $t_{0.95}$ is 1.73. The conclusion was that the temperatures measured by the device are not statistically different than the temperatures measured by the medical thermometer for all the distances.

Table 1. Human body temperature measurements.

Distance to subject	Temperature Measurements – Mean and std. deviation		
	Proposed device	Medical grade thermometer	Statistically significant
40 cm	36.5 (0.29)	36.4 (0.17)	NO
50 cm	36.42 (0.32)		NO
60 cm	36.34 (0.35)		YES

When deployed in production, for enhancing the process of staff thermal scanning, the device is mounted at the entrance point in a company. When an employee arrives at the office, he/she will position in front of the device at a distance ranging anywhere between 40-60 cm. A large display will provide the information required to the person, i.e. the distance to the device, environmental temperature and humidity and the forehead detected temperature. If the user is too close or too far from the device, a message will be shown on the display.

5. Enhancements Using Deep Learning

Convolutional neural networks (CNN) are a powerful tool for performing image processing tasks without going through the feature engineering stage. In the present work, a CNN is proposed to map the pixels from the thermal sensor to the temperature that was read by the medical grade thermometer.

The pixels corresponding to the forehead obtained from the employed computer vision pipeline will be resized and stored in a database as monochrome images. Since the employed thermal stream has a 120 by 160 pixels resolution, the region corresponding to the forehead will be a relatively reduced size matrix, for example 10 by 10 pixels. Therefore, the first convolutional layer of the CNN will have the input size an array of 100 values. Subsequent convolutional layers will encode the activation maps of the initial layer and the final part of the network will be represented by classical fully connected layers of neurons. The CNN can have as additional inputs the information received from the LiDAR and the environmental temperature sensor.

The output of the CNN will be a single value, obtained after training using as loss function the mean square error (MSE) between the output of the final neuron and the ground truth values obtained by using a medical thermometer. Gradually, during training, the parameters of the CNN will be optimized to correct the errors and estimate the forehead temperature as close as possible to the values obtained from the medical thermometer.

To employ the deep learning approach to the proposed device, however, it will need sensor readings from numerous persons with higher body temperature than the normal temperature, i.e. ill individuals as training data. For this reason, the deep learning approach is the subject of future research. The research question that will prove interesting to be investigated is: can the deep learning approach provide the same reliability after removing some of the sensors required by the present hardware solution?

6. Conclusions

Generic thermal cameras are not used for medical purposes in human body temperature measurement. In

the context of a global pandemic, in numerous companies and institutions the thermal scanning of the staff is not accurately performed. The present work proposes a non-invasive human body temperature measurement by fusing information from various sensors. The experimental results show that there is no statistical significance between the temperatures measured by the proposed device and the medical grade thermometer typically used in hospitals. Future work will focus on integrating deep neural networks to map the region of pixels from the thermal sensor corresponding to the human forehead to a value read by the medical thermometer.

Acknowledgment

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Beehive Air Measurements Using E-nose – Practical Aspects

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Summary: Honeybee poluation is increasingly threatened worldwide. Proper beekeeping pracices may change this tendency. Chemical properties of air inside the hive are a source of useful information for designing relevant actions. The aim of this paper is to present selected practical aspects regarding application of E-nose for beehive air measurements in field conditions.

Keywords: Air quality, Gas sensor, E-nose, Portable device, Bee.

1. Introduction

Bees are critically important for the environment and to the economy. Unfortunately, their population is disappearing at an alarming rate throughout the world.

Proper management of the apiary can prevent bee diseases and, at the same time, allow to obtain high quality products, respecting consumer's health [1]. The appropriate information is essential for this purpose. It can be obtained on the basis of regular observations or the measurement of the appropriate parameters. The first approach is: (1) readily available to bee specialists; (2) time consuming; (3) subjective and strongly dependent on beekeeper experience; (4) time and labor consuming. The best source of highly reliable information about the condition of bee colony or events that may require the beekeeper's action is beehive monitoring based on measurements of the appropriate parameters.

Currently, beehive monitoring is focused on the continuous, automatic determination of temperature, air humidity, analysis of sound and vibration of a beehive, counting of outgoing and incoming bees as well as the number of bees in the hive entrance area, video observation, weighing the colony, as well as determining the chemical content of beehive air. Unfortunately, chemical properties of air in hives are relatively rarely considered as a source of information, which is useful for the management of the apiary.

This problem can be overcome by using electronic nose technology. Usually, the Electronic Nose is understood as an instrument, which "comprises an array of electronic chemical sensors with partial sensitivity and an appropriate pattern recognition system capable of recognizing simple or complex odors." In [2], we have shown, that devices based on gas sensor array and module for classification of beehive air quality allow to detect varroosis, one of the most dangerous diseases of honey bees.

The aim of this paper is to present several practical aspects regarding application of E-nose for beehive air measurements in field conditions. Our attention was mainly focused on issues related to gas sampling and

the influence of environmental conditions on the measurement process.

2. Results

In our work, it was assumed, that beekeepers need mainly portable, direct reading instruments, which are simple in use. These device should work with minimal operator attention/intervention; offer good precision of classification and provide a real time, continuous, online information. The instrument, which fulfils these requirements was conceptualized in the Laboratory of Sensor Technique and Indoor Air Quality Studies, WRUST and it was made by InSysPom, see Fig. 1. The device is based on semiconductor gas sensors.



Fig. 1. The instrument for classification of beehive air quality.

Online measurements were preferred in our approach, because in this case, the influence of the instrument on the bee behavior is significantly reduced. Online measurements require dynamic, extractive sampling. We found that the location of the sampling probe inside beehive is of the utmost importance for the measurement process. The

placement of the probe near the beehive entrance (Fig. 2, left), which is favoured by beekeepers, typically causes strong fluctuations of the gas sensor signal (Fig. 3a). When accessing the beehive from the top, and locating the probe between combs (Fig. 2, right) smooth sensor signals are recorded typically (Fig. 3b). We think this effect is related to the movement of bees.



Fig. 2. Sampling probe location in the beehive.

Filters are important element of each gas sampling system. They protect sensors against particulate matter. It was observed that mounting a syringe filter on the gas line lowered the signals of individual sensors and changed their mutual layout, see Fig. 4.

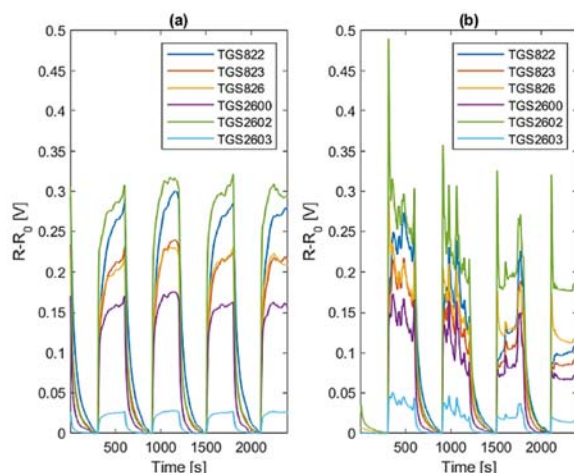


Fig. 3. Gas sensors responses to beehive air when the probe was located: (a) between combs, (b) near the beehive entrance.

During transport between the beehive and gas sensor device, the gas sample flowing inside Teflon tube is under influence of ambient conditions. Physical conditions inside and outside the beehive are usually different. Water condensation was observed inside the Teflon tube during sampling on the day, when the ambient temperature did not drop below 24 °C, see Fig. 5. This effect has an adverse impact on measurement results. Upon tubes clogging with water, filters may be wetted. It should be noted, that removing water vapour from the gas sample is not an easy task.

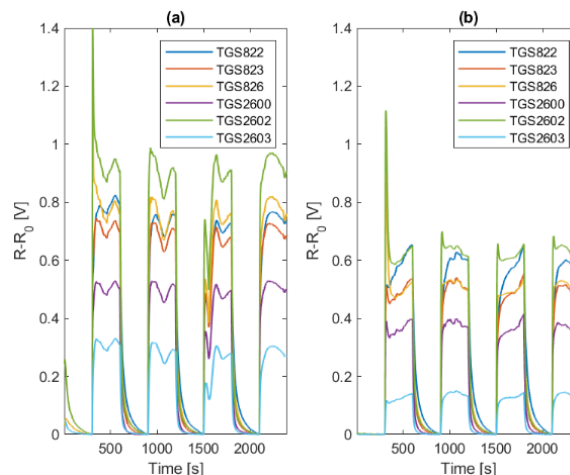


Fig. 4. Gas sensors responses to beehive air when the gas line was: (a) Not fitted, (b) Fitted with the syringe filter.



Fig. 5. Water vapour condensation inside the tubing which transported the gas sample between the beehive and the measurement device.

3. Conclusions

Our studies clearly show, that many factors have to be taken into account when designing the E-nose, which is dedicated for the measurement in the field conditions.

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(001)

Opto-galvanic Spectroscopy of Kr 84 / Ne Transitions in the Range of 1270 nm–1640 nm for Use as Optical Wavelength Standards

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Summary: In this work the optical absorptions of the Krypton 84 $1s_2 - 2p_8$ between 1270-1640 nm are analysed to be used for an optical wavelength standard in this regime. The spectrum was measured using opto-galvanic spectroscopy which shows higher sensitivity than direct optical absorption spectroscopy. More than 30 usable absorption lines have been determined. In the analysis, it was shown, that absorptions of isotopes of Kr show wavelength shifts of several nm from Kr 84. Here, all absorption lines of the Krypton / Neon mixed gas were summarized which can be used in the whole ITU transmission range of 1270 nm to 1640 nm. Additionally, a wavelength drift of the detected lines was measured using a climate chamber between 10 °C and 50 °C. Here we propose a grid of 27 stable absorption lines of Kr to be used as a wavelength standard for optical communications systems. Combining the Galvatron absorption of Kr lines with active feedback to a multi-channel laser comb source, we propose this system to be used as a high stable reference source for the calibration of spectral analyzers or in optical WDM networks.

Keywords: Wavelength Division Multiplex systems WDM, Optical wavelength references, Optical networks, Opto galvanic spectroscopy.

1. Introduction

Optical communication systems are since the last 30 years the backbone of all international data and telecom communications. Up to now, more than 10 TBit/s can be distributed via one optical fiber in long-haul systems. This was realized by the use of highly parallel multi-channel transmission via many colors, called Wavelength Division Multiplex (WDM). Here, the International Telecommunication Union specified the channel location into a frequency space grid with 100/25 GHz distance in the infrared regime from 1300 nm to 1640 nm. To avoid interference between these densely arranged transmission channels, each transmitter output frequency has to be stabilized exactly in its center frequency to others. To supervise and stabilize the transmission comb, absolute frequency standards are very helpful to be integrated into optical cross-connect stations. Using excited state absorption lines of a Nobel gas is an easy way to generate reference frequencies, because these states are very stable and very small in linewidth. A well-known method is the detection of the absorption by the opto-galvanic effect (OGE). Here, a tube is filled with the gas Kr and the detection of the absorption will be performed by an electrical signal of the absorption cell via electrodes with high voltage supply. Many atomic transitions are nearby to the ITU grid frequencies of ITU G 652 Recommendation.

2. Wavelength Standards

Nowadays, atomic and molecular absorption lines are used in wide areas of optics as wavelength calibration references, since these form highly stable

frequency standards even under fluctuating environmental conditions. Molecules such as acetylene and hydrogen cyanide have very narrow-band absorptions in the range of 1500 nm as rotation-vibration transitions. These transitions can easily be detected directly by absorption spectroscopy. Atomic transitions of noble gases in the range of 1300-1700 nm are generally not directly detectable, since they are transitions between excited states of the atom. These excitation states are usually formed with the aid of laser excitation or by means of an electrical gas discharge. A single gas laser reference line from the HeNe laser at 1523 nm can be used. Further atomic or molecular references can only be achieved by doubling the frequency of the light from e.g. Rubidium from 1560 nm to 780 nm can be achieved, which can only be achieved with very high expenditure on equipment [1-3].

To describe the accuracy of a frequency reference, one refers to the ratio of the relative frequency deviation to the absolute transmission frequency of the atomic or molecular transition used. With the Cesium standard at 10 MHz, a source with an accuracy of 1.5×10^{-14} is available internationally. The second is 9,192,631,770 times the period of the radiation corresponding to the transition between the two hyperfine structure levels of the ground state of atoms of the nuclide ^{133}Cs [4].

Additional standards are available in most developed industrial countries from their respective metrology organizations. Since such a source is only portable to a limited extent, a further secondary source based on the Cs reference is sought. On the one hand, the accuracy of the reference at 10 MHz must be extended to other frequency ranges eg. to realize a frequency comb [5, 6].

Starting from the Cs standard with an accuracy of better 10^{-14} , there should be a class of reference lasers in the field of optical communication technology derived directly from the Cs in the overlay experiment. These lasers should have an accuracy of better than 10^{-10} . Building on this, there should be portable laboratory standards that should achieve a frequency accuracy of 10^{-9} ($\Delta f_R = 0.2$ MHz at $f = 200$ THz or $\lambda = 1500$ nm). This type of laboratory standard represents an important link between the non-portable reference lasers and the less precise field equipment. The distance between two reference frequencies should not exceed $\Delta \lambda_R = 20$ nm in order to allow a sufficiently small distance between two calibration points of a spectrometer. However, these laboratory standards require complex electronics and beam guidance in order to avoid the Doppler broadening of the natural line width of the absorptions used [7].

Atomic and molecular absorption lines are used in a wide range of optics as a wavelength calibration references, since these form even under fluctuating ambient conditions highly stable frequency standards. Molecules such as acetylene and hydrogen cyanide have very narrow-band absorptions in the range of 1500 nm [1, 8-10]. These transitions can be easily detected directly by absorption spectroscopy. Atomic transitions of rare gases in the range of 1300-1700 nm are generally not directly detectable because they are transitions between excited states of the atom. Usually, these excited states are produced by means of laser excitation or by an electrical gas discharge. A single gas laser reference line from the HeNe laser at 1523 nm is available.

These portable laboratory standards [11] with a frequency accuracy of 10^{-9} (0.2 MHz at 200 THz, 1500 nm) represent an important link between the non-portable reference laser and the less accurate field equipment. These laboratory standards make a very demanding electronics and beam guidance necessary to avoid the Doppler broadening of the natural linewidth of the absorptions used [12-14].

3. Experimental Setup

Figs. 1 and 2 show the test site with the equipment used. From the tunable laser source (0) the laser beam is modulated by the chopper wheel (1) intensity. The control unit of the chopper is connected to a phase sensitive lock-in amplifier which detects the change in electrical voltage in the Galvatron tube (3). A high voltage supply was used to start the discharge current. With the wavemeter (4) the wavelength is determined with an accuracy of 3×10^{-13} . In the Galvatron-tube, the isotopes of Krypton are included in the following proportions (see Table 1). It also contains as a filling gas Neon, the cathode is made of iron.

In examining the temperature dependency the Galvatron tube was placed in a climate chamber, shown Table 2.

Table 1. Isotopic proportions of natural Krypton gas [14].

Isotope	⁷⁸ Kr	⁸⁰ Kr	⁸² Kr	⁸³ Kr	⁸⁴ Kr	⁸⁶ Kr
Part [%]	0.35	2.28	11.59	11.49	57.00	17.30

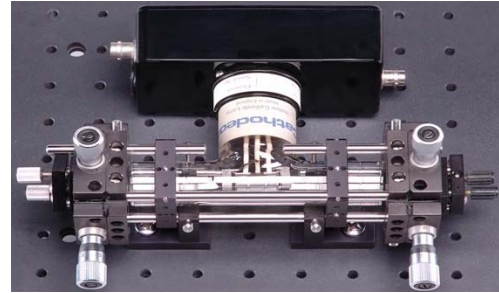


Fig. 1. Experimental setup Galvatron tube with optical coupling setup.

Fig. 2 shows the schematic setup of and the connections between the devices used. All units were controlled with the help of the GPIB interface of a LabVIEW program that increases the wavelength of the laser source from a start value to an end value with a fixed step size. The intensity of the lock-in amplifier and the wavelength is read at the wavemeter and stored in a file. Since the wavelength range could not be covered by a laser source, 3 different sources with the following areas a) were 1270-1375 nm b) 1370-1480 nm, and c) used from 1440 to 1640 nm.

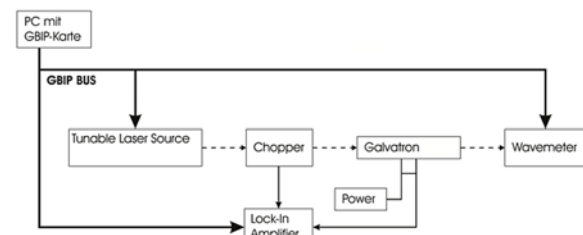


Fig. 2. Electrical and optical connections of experimental setup.

In Fig. 3, the measure section is depicted beginning at the laser source via the chopper, the Galvatron-tube to the wavemeter. In the experiments, the laser source is operated with an output power of 10 dBm, which is lowered by the loss of connectors and especially by the chopper and the Galvatron-tube down to -21 dBm at the wavemeter.

Based on the LabVIEW program (see Fig. 4) created by van Doornik [15], some modifications and extensions were made. The entire communication with regard to setting parameters and reading out the measured values from the laser source and the wavemeter has already been implemented. The previous program includes a part in which the internal modulation of the laser source was programmed. The internal intensity modulation was not acceptable for our experiment because the distance between the main wavelength and the sidebands was too small. The

modulation is implemented externally via a chopper. This reduced the intensity or the spacing of the sidebands that occurred.

Previously, the measured values of the lock-in amplifier were read out via a DAQ card. By converting the GPIB interface of the lock-in amplifier, the measured values could now be read out digitally, which leads to a further reduction in measurement

inaccuracies, since the quantization noise of the analog-digital conversion of the card is eliminated. For this purpose, the initialization and the measured value recorded at the lock-in amplifier were integrated into the program. Furthermore, the clarity of the program was increased by using local variables and the user interface was restructured.

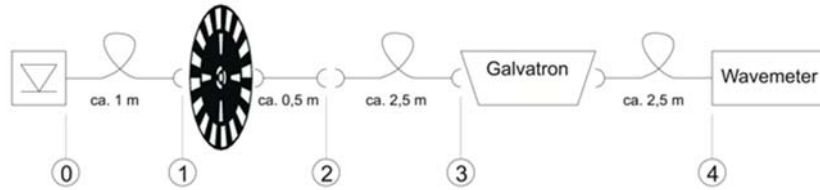


Fig. 3. Optical setup: 0: tunable laser source, 1: optical chopper wheel, 2: opt. connector, 3: Galvatron tube, 4: optical high resolution wavemeter.

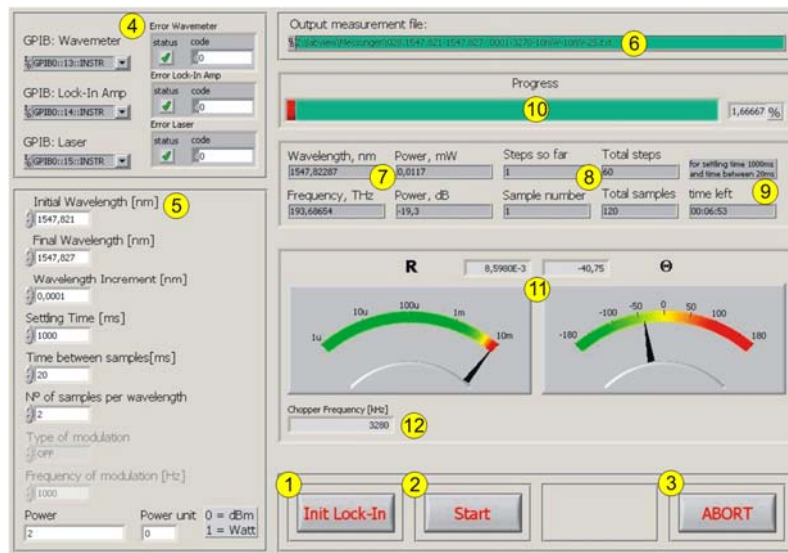


Fig. 4. LabVIEW user interface.

4. Results and Discussion

From Kaufmann et al. an energy level diagram from Kr I is available which is depicted in Fig. 5.

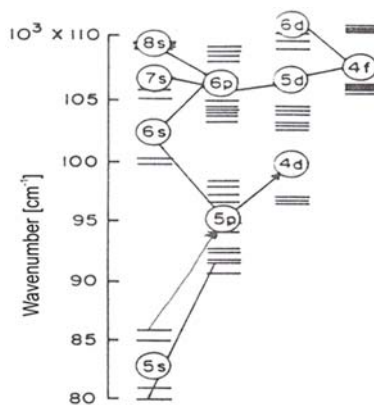


Fig. 5. Excerpt from the energy level diagram of Kr I from [17].

Several absorption lines can be noticed between 5S and 8S level. Here, the absorption lines of Krypton and Neon in the range of 1270-1640 nm were measured. To our knowledge this is the first continuous measurement and presentation of opto galvanic absorption lines in the whole range of the ITU-T G.652 [16] for optical communication systems. In Fig. 6, there are the consistently numbered absorption lines of Galvatron tube in the areas of laser sources.

It should be noted, is that lines of neon (eg line 01), iron (eg line 27) and carbon (eg line 15) were measured. These were using the database of the NIST determined [2]. Afterwards, at selected lines ({1286.54123 nm line 03}, {1374.26111 nm line 10} and {1547.82593 nm line 41}) the influence of temperature on the drift of the intensity maximum is investigated. As an example the results of the krypton absorption line shown at 1286.54 nm (Fig. 5d) is depicted. The picture shows the superposition of three isotopes of krypton and the Gaussian fit of the three

individual lines. The resulting deviations of the wavelengths from those described in the literature [17, 18] are due to the measurement accuracy of the wavemeter.

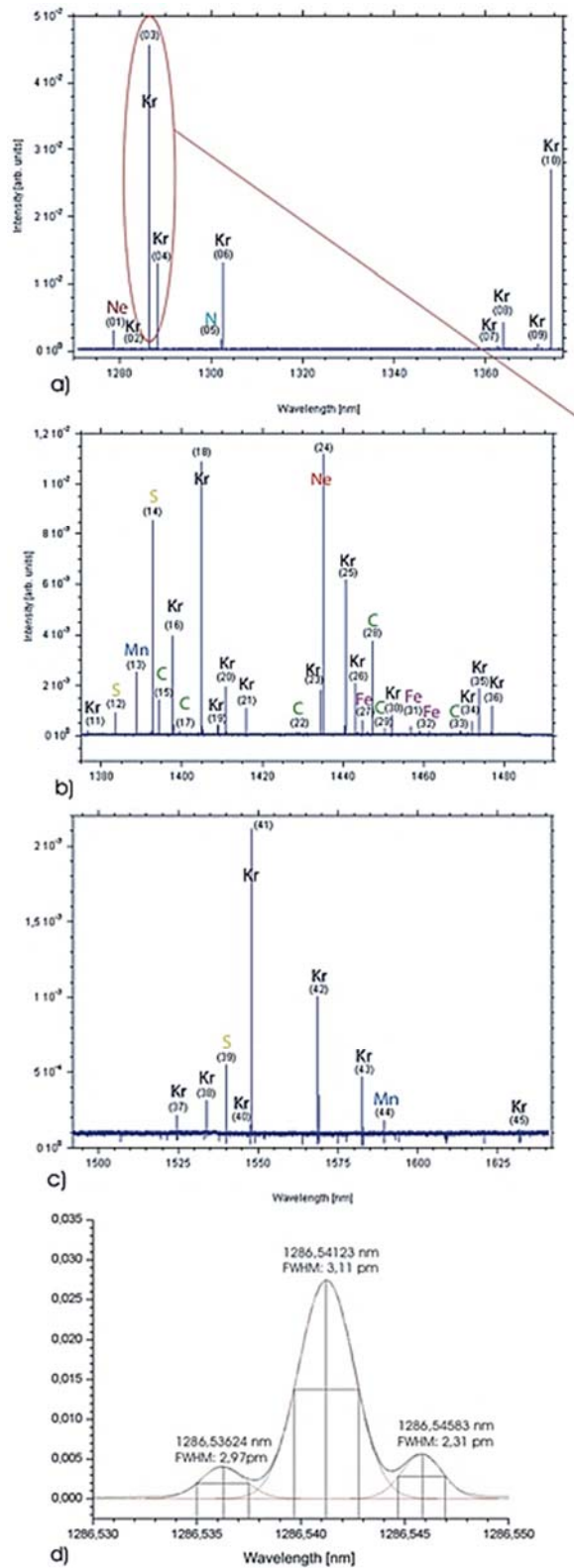


Fig. 6. Absorption lines of Kr line 1286.54123 nm at 10° C, a) Kr-absorption lines between 1200 nm – 1380 nm, b) Kr-absorption lines between 1380 nm – 1490 nm, c) Kr-absorption lines between 1490 nm – 1640 nm, d) high resolution absorption spectrum Kr 1286nm-line.

In the tests no drift of the main peak as a function of temperature could be determined (see Table 2). The half-maximum-value width of the main peak remains almost constant. In the secondary maxima a slight broadening was observed with increasing temperature.

Table 2. Values of the lines at 1286.54106 nm at various temperatures.

T/ °C	λ [nm]	FW-HM [pm]	λ [nm]	FW-HM [pm]	λ [nm]	FW-HM [pm]
10	1286.53624	2.52	1286.54123	3.0	1286.54583	2.31
20	1286.53619	2.56	1286.54121	3.24	1286.54584	2.33
30	1286.53621	2.58	1286.54120	3.11	1286.54581	2.34
40	1286.53621	2.63	1286.54121	3.12	1286.54582	2.34
50	1286.53623	2.73	1286.54122	3.12	1286.54583	2.45

5. Conclusions

A complete overview of all opto-galvanic absorption lines of Krypton and Neon mixed gas in the range of ITU-T G.652 were measured and reported. The measured absorption lines show little deviations on the dates given in the literature. The largest deviations were ± 0.5 pm, which can be explained already by the measurement accuracy of the used Agilent wavemeter.

The stronger lines are separated between 5 nm and 30 nm. Hence the creation of a grid of excellent reference lines can be used as a very high precise wavelength standard for optical communications or in spectrum analyzers for internal calibration. Since the absorption lines do not depend on the temperature, a long-term stable wavelength standard is technical applicable.

Acknowledgements

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(003)

Hardware Implementation and Test of a Verification System for Time Signals and -telegrams

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Summary: A correct time is basis for all time-triggered actions in information and communication systems. There are several methods for time synchronization between remote clocks, but system errors or manipulations like delays and attacks on transmitted time stamps can cause wrong time information at receivers and endanger the security and safety of devices, environment and people. This paper describes a verification method that physically measures the time between two received time information messages and compares the measured time with the difference of time information contained in succeeding time information messages. The method of time verification is based on the relation between time and frequency of the modulated signal used in wireless communication. The receiver counts the periods of the modulated signal used to transmit the time information. This method is applicable for all time distribution systems that use a continuous carrier for modulation. This paper describes the method of this time verification method, a hardware implementation and results of extensive tests.

Keywords: Broadcast time distribution, GNSS, GPS, Navstar, Radio ripple control, Time synchronization, Time verification, Wireless time distribution, WWVx.

1. Introduction

Each device that requires the current time for its actions must have an available and reliable time source. Incorrect time information can lead to situations, in which actions are started or stopped at wrong times, clocks are not synchronized correctly, or communication is no longer possible. This can lead to a loss of control over processes. Cryptographic mechanisms often use an actual timestamp to ensure uniqueness and temporal correctness for key management, data integrity and authentication protocols. Synchronized clocks are an important requirement for distributed systems, but it is also essential that the synchronized time is physically correct. Therefore, it is necessary to verify a received time information. Trustworthy time information should only be used for time synchronization and scheduled process control.

Many wireless time distribution systems broadcast time information on various frequency bands with different coding and modulation schemes. Furthermore, actual time information is distributed by systems like *global navigation satellite systems* (GNSS) and data communications systems such as *ripple control* technology. Millions of receivers rely on the time information and use it to synchronize local clocks. The time verification scheme that is described and tested in this paper is suitable to be used by entities within such broadcast systems before they apply they provided time information for synchronization or process control.

2. State of the Art

2.1. Time Distribution Services

There is a network of time code transmitters that is distributed all over the world. These time code transmitters broadcast current time information on different frequency bands with various coding and modulation schemes like ASK, FSK, PSK, BPSK or combinations of them. Table 1 lists some popular time code transmitters, which continuously send time information.

Table 1. Time code transmitter [1].

Call Sign	Location	Carrier Frequency	Modulation
WWV	USA	2.5/5/10/15/20 MHz	ASK
WWVB	USA	60 kHz	ASK/BPSK
WWVH	USA	2.5/5/10/15 MHz	ASK
MSF	GB	60 kHz	ASK
DCF77	GER	77.5 kHz	ASK/PSK
JJY	JP	40/60 kHz	ASK

In contrast to time code transmitters, which transmit time information continuously, there are services, which broadcast time telegrams. These time telegrams that contain current time information are transmitted between customer oriented data telegrams. One example is the radio ripple control service of the EFR GmbH in Germany, which operates three distributed broadcast stations at a carrier frequency of $f_c = 129.1$ kHz, $f_c = 139$ kHz and $f_c = 135$ kHz. These three stations cover most parts of Europe.

Other examples for time distribution service are navigation systems like Navstar-GPS (USA), Glonass (Russia), BeiDou (China) and Galileo (Europe).

None of the described time distribution services provides a backchannel, which means that communication is always unidirectional and entities are not able to send information back to the broadcasting systems.

2.2. Time Synchronization

The simplest method of time synchronization is the use of the power line frequency, e.g. 50 Hz or 60 Hz. However, this is not really a time synchronization, because no time information is exchanged. In this scenario, the local clock is not synchronized with another (master) clock. Instead, a user sets the initial time and the frequency of the alternating current drives the clock.

There are patents, e.g. [2] that describe a method of time synchronization by transmission of the time information of a master clock or clock of a higher hierarchical level plus adding the propagation time of a message containing the master clock's time information. The propagation time is calculated by the difference between both local times.

The *Network Time Protocol (NTP)* is the most common protocol for time synchronization in computer networks [3]. It uses the local time of a higher hierarchy level and considers the propagation time, which is calculated by the roundtrip time divided by 2.

The Simple Network Time Protocol (SNTP) is a less complex version of NTP.

Another time distribution protocol is *Precision Time Protocol (PTP)*. PTP is suitable for scenarios, which require high precision of time synchronization. PTP is defined in IEEE 1588-2008. The focus of PTP is the synchronization of clocks inside a LAN. A new application of PTP is specified in IEEE 802.1AS for Time Sensitive Networks (TSN), where it is used for scheduling of traffic shaping (IEEE 802.1Qbv). However IEEE 802.1AS does not consider techniques to obtain accurate time from an external source.

All described time synchronization protocols consider the roundtrip times during synchronization.

Therefore, NTP, SNTP and PTP cannot be applied in time distribution systems such as the ones described in Section 2.1.

2.3. Risks and Attacks

Wrong time synchronization may be caused by the system behavior or intentional manipulations.

In March 2018, the clocks, which were driven by the frequency of power lines, were delayed up to 6 minutes in great parts of Europe. The reason was a reduced frequency of the electric power distribution system [4, 5]. Incidents such as a power outage within

or outside a building require all affected clocks to be manually set to the current time.

All time synchronization systems that rely on received time information are potentially vulnerable to attacks that aim to manipulate an internal real time clock (RTC). These attacks include active attacks like Denial of Service attacks (DoS), Man-In-The-Middle attacks (MITM) or replay attacks. Attackers can manipulate transmitted timestamps as well as propagation times of messages. Cryptographic methods such as Message Authentication Codes or digital signatures only offer a partial protection, e.g. against manipulation of time telegrams. However, to cause a delay of time stamps is always possible.

Local clocks can be used to verify the plausibility of received time information, but only in certain limits. They do not replace the necessity to synchronize clocks with the correct time.

In wireless communication it is difficult to distinguish between DoS attacks and atmospheric noise. Both cannot be prevented, and it is not easy to detect a DoS by a recipient.

To initiate a MITM attack, an attacker must know the transmission protocol structure of the wireless communication system. If the transmission power of the jamming transmitter used for the MITM attack is high enough, an attacker can transmit its own time information to manipulate the clock of the receiver in a range accepted by the synchronization process.

A replay attack is a special kind of MITM attack. First, an attacker records transmitted time data and performs a DoS attack simultaneously. In the next step, the attacker transmits the recorded time data with a certain delay. These re-played messages may have contain message authentication codes, but an attacker does not change any contents of recorded messages. Both [6, 7] give detailed descriptions of aforementioned attacks.

Time synchronization via GPS becomes more and more popular and important. Therefore, GPS was checked continuously for security gaps since the year 2000. A distinction is made between jamming and spoofing attacks. A jammer transmits noise in the periphery of a recipient during a jamming attack. In this case, the field intensity of the disturbance is much higher than the received GPS signal and the recipient is not able to lock the current GPS signal. During a spoofing attack, an attacker transmits modified GPS data telegrams.

In the year 2013, researchers of the University of Texas were able to manipulate the GPS navigation system of a yacht in the Mediterranean Sea using a spoofing attack [8]. Detailed descriptions about more attacks on GPS recipients are published in [9, 10].

Active attacks on a wireless radio require attackers to operate a portable jamming transmitter with an external antenna in close proximity to the attacked target. [11] presents results of tests that have been performed with the German time signal long wave transmitter DCF 77.

3. Time Information Verification

A new time verification method was presented at the *IEEE IFCS/EFTF 2015 (International Frequency Control Symposium and European Frequency and Time Forum)* in Denver and the *EFTF 2016* in York (UK) [1, 7].

The general idea of this verification method is based on the comparison of a physical and a logic time interval between two succeeding events. The physical difference time td_{cd_i} ("reference time interval") is measured between two reference times.

A reference time is the moment, in which the time information $time_i$ is valid. This can be a fixed position within the time signal or time telegram tt_i .

The physical difference time td_{cd_i} is measured by counting the periods of the received modulated signal between two succeeding successfully received time

signals resp. time telegrams (tt_i and tt_{i-1}). The index i refers to one arbitrary element of all sent time signals or telegrams.

The logical difference time Δtt_i (target time interval) is the difference between the time information ($time_i$ and $time_{i-1}$) contained in tt_i and tt_{i-1} . Fig. 1 shows the timing of the verification sequence for time signals and the measurement of td_{cd_i} .

The verification process compares the reference time interval and the target time interval for every received time signal or time telegram relative to its predecessor. If the distance of both intervals is smaller than a predefined threshold t_{th} , the latest received time information is verified and considered trustworthy. Otherwise, the verification process failed. Fig. 2 illustrates the complete verification scheme as flow chart. Detailed information on the described verification method can be obtained in [1, 7, 12-15].

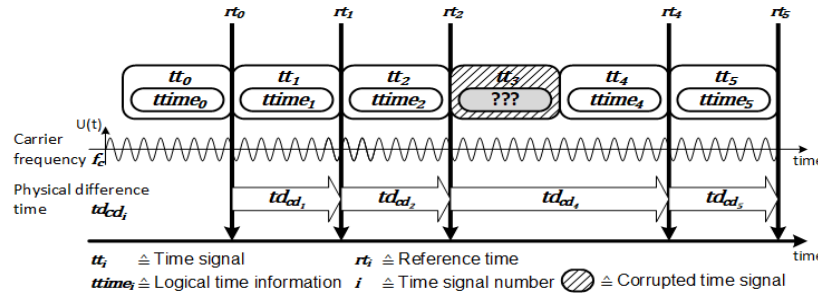


Fig. 1. Measurement of the physical difference time for time signals.

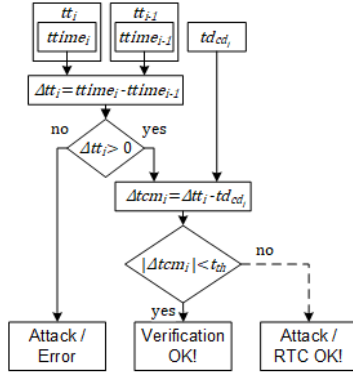


Fig. 2. Measurement of the physical difference time for time signals or telegrams.

At the beginning of the verification mechanism, the received first time information ($time_0$) is set by a trusted local time source.

4. Implementation

The presented verification method can be adapted to all time transmission systems that use a continuous carrier for modulation. If multiple modulated time data streams are received by a recipient (e.g. in the case of GPS), every time data stream is verified in a separate verification module (Fig. 3).

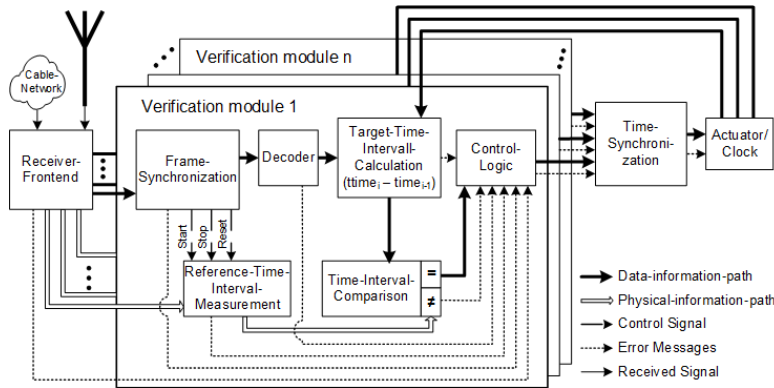


Fig. 3. Block diagram of the verification process with n verification modules.

All verified received time information are synchronized in the Time Synchronization module.

One or more digital counters measure the physical difference time td_{cd_i} . The clock signal for the digital counters is recovered directly from the received modulated signal. The number of clock oscillations cd_i is converted into the reference time interval td_{cd_i} .

Equation (1) shows this conversion for an ASK modulated received signal that is transmitted with a carrier frequency f_c .

$$td_{cd_i} = \frac{cd_i}{f_c} [s] \quad (1)$$

Equation (1) must be adapted for other modulation schemes. Further information about the realization of the counter logic for other modulation schemes can be found in [1, 7, 15].

The target time interval Δtt_i is calculated by subtracting the time information $ttime_{i-1}$ from the time information $ttime_i$. Both time information are obtained from two succeeding successfully received time signals or time telegrams tt_{i-1} and tt_i .

$$\Delta tt_i = ttime_i - ttime_{i-1} [s] \quad (2)$$

In this case, $ttime_i$ is the currently received time data and $ttime_{i-1}$ represents the last correct verified time data.

If Δtt_i (2) is not greater than 0, a replay attack is detected, because the currently received time information $ttime_i$ can be not older than the last verified time information $ttime_{i-1}$. In this case the verification of $ttime_i$ fails, an error flag is set, the counting of periods is continued and it will be waited for the next correctly received time signal or time telegram, which will be handled as tt_i .

In the next step, the difference Δtcm_i between the reference time interval td_{cd_i} and the target time interval Δtt_i is calculated

$$\Delta tcm_i = \Delta tt_i - td_{cd_i} [s] \quad (3)$$

If $|\Delta tcm_i|$ is smaller than a predefined threshold value t_{th} ($|\Delta tcm_i| < t_{th}$), the verification check for the last received time information is successful and the received time information can be used as new trusted $ttime_{i-1}$. Otherwise, the verification process fails. A value of t_{th} will be proposed later.

Patent US 10,165,530 B2 [14] provides a detailed description of the block diagram shown in Fig. 3.

The presented verification method has been tested for the German time code transmitter DCF77 and the radio ripple control service of EFR GmbH. In this paper, all presented measurements used DCF77 hardware. This hardware consists of an external HF-Frontend, which is part of the *Meinberg receiver UA24V* [16], a FPGA-based measurement logic and a PC software for evaluation.

5. Field Tests

The precision of the Reference-Time-Interval-Measurement logic was tested with the German time distribution system DCF77. This transmitter operates continuously and transmits one complete time signal every minute. The system uses a combination of ASK and PSK modulation [17].

The difference time td_{cd_i} and the counter values cd_i of 19933 received time signals were evaluated during the measurement interval from May 14, 2020 12:48 am until May 28, 2020 09:00 am (CEST).

Fig. 4 shows the measured deviation of the digital counter values cd_i from the calculated value cd_{calc} (4) in the measurement interval. The normal error free value of cd_{calc} per correct received time signal is calculated by the multiplication of the transmission time of a time signal (a complete time signal is transmitted every minute over the carrier frequency of 77.5 kHz).

$$cd_{calc} = 60 s \cdot 77.5 \text{ kHz} = 4650000 \quad (4)$$

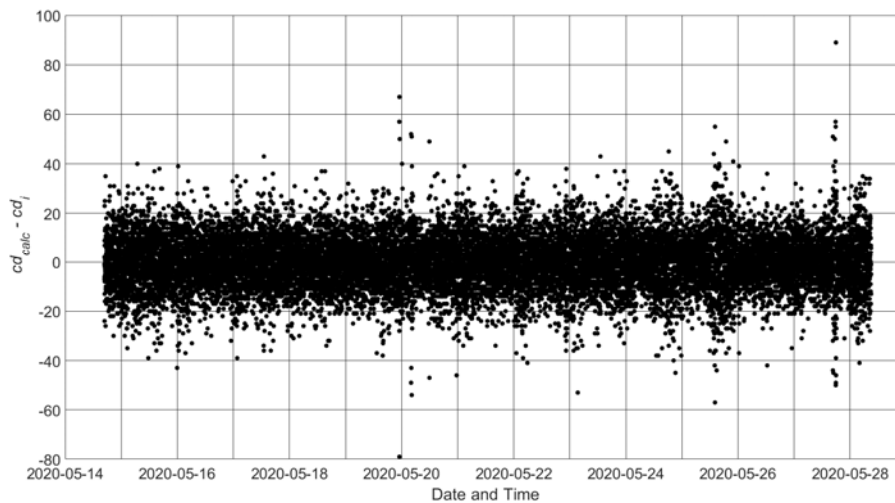


Fig. 4. Deviation of the DCF77 counter values from cd_{calc} .

The deviation values ($cd_{calc} - cd_i$) can be converted directly into the distance time Δtcm_i between two correct received time signals as shown in (5).

$$\Delta tcm_i = \frac{cd_{calc} - cd_i}{f_c} [s] \quad (5)$$

Within the measurement interval, the maximum deviation of the counter values are $\Delta cd_{max} = +89$ and $\Delta cd_{min} = -79$ oscillations, corresponding to a

max time deviation of 1.15 ms, and the arithmetic mean value amounts to $cd_{mean} = 0.01485$, corresponding to 0.2 μ s.

Fig. 5 shows the normal frequency distribution calculated deviation values ($cd_{calc} - cd_i$) with the standard deviation of $\sigma_{cd} = 10.61$ and the variance of $\sigma_{cd}^2 = 112.54$. The σ_{cd} corresponds to $cd_{\sigma_{cd}} \approx 11$ oscillations of the carrier frequency f_c , corresponding to 142 μ s.

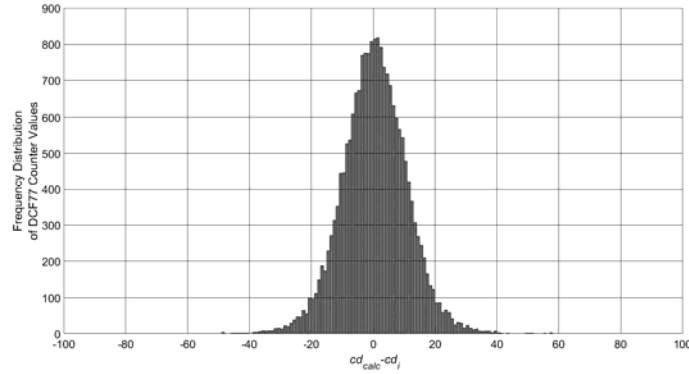


Fig. 5. Frequency distribution of deviation of counter values.

Within the total measurement period of 13 days, 20 hours and 12 Minutes, only one received time signal had a parity error, which was detected. The counting was continued, the error does not have any impact on the verification process. The verification process continues with the reception of the next correct time signal.

This implies that the reference time td_{cd_i} for the corrupted and unused time signal has to be added to the reference time $td_{cd_{i+1}}$ of the next received time signal.

It is possible to optimize this addition in the hardware implementation: A 64 bit counter is clocked by the carrier frequency f_c . This counter once starts at the reference time $ttime_0$. From this moment on, the counter runs free until the verification logic is stopped or shut down. The counter value is stored at each reference time $ttime_i$. The difference of two correct succeeding counter values is the value cd_i . If the logic

detects a corrupted time signal at $ttime_i$, the actual counter value is discarded and the verification logic waits for the next correct received time signal resp. telegram. The new counter value is used to calculate the difference value cd_i .

In the DCF77 Time transmission system, the 64 bit counter can run $t_{max} = \frac{2^{64}}{4650000} [s]$ without an overflow.

If no parity or data errors are detected, the internal verification time interval for the measurement of td_{cd_i} and the calculation of Δtti is one minute.

This interval increases up to two or more minutes in case of errors, when parity errors are detected or the verification process fails. This is caused by wrong results inside the “Time-Interval-Comparison”. Fig. 6 shows the normal frequency distribution for several verification time intervals ti in the range from 1 minute up to 240 minutes.

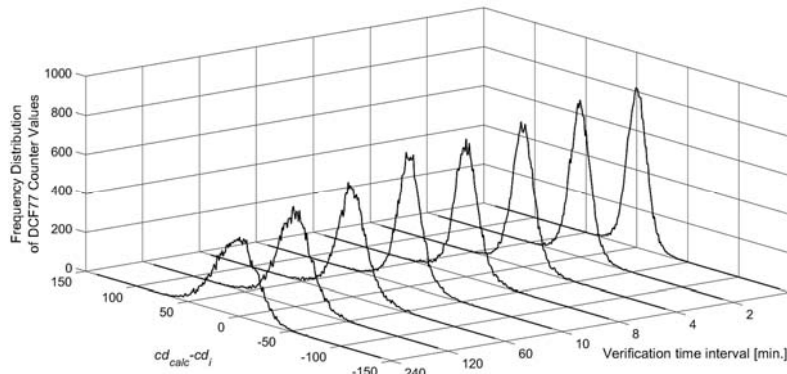


Fig. 6. Frequency Distribution for different Verification time intervals.

If the verification time intervals increases, the standard deviation increases as well (Fig. 6). Table 2 shows the maximum deviation of the counter values $\Delta cd_{max_{ti}}$, $\Delta cd_{min_{ti}}$, the standard deviation $\sigma_{cd_{ti}}$ and the corresponding times $t_{\Delta cd_{max}}$, $t_{\Delta cd_{min}}$ and $t_{\sigma_{cd}}$.

Table 2. Counter and standard deviation in the Verification Interval ti .

	Verification time interval (ti [min])							
ti	1	2	4	8	10	60	120	240
$\Delta cd_{max_{ti}}$	89	81	85	105	95	130	103	110
$t_{\Delta cd_{max}}$ [ms]	1.15	1.05	1.10	1.35	1.23	1.68	1.33	1.42
$\Delta cd_{min_{ti}}$	-79	-107	-74	-76	-76	-83	-88	-81
$t_{\Delta cd_{min}}$ [ms]	1.02	1.38	0.95	0.98	0.98	1.07	1.14	1.05
σ_{cd}	10.61	11.29	12.43	13.60	14.07	16.89	18.16	20.45
$t_{\sigma_{cd}}$ [μs]	136.9	145.6	160.3	175.5	181.5	217.9	234.3	263.9

6. Determination of the Threshold Value t_{th}

The threshold value t_{th} is used in the “Time-Interval-Comparison” (Fig 3). It determines the required accuracy of the deviation between reference time interval td_{cd_i} and target time interval Δt_{ti} . This value is dependent on following requirements:

Modulation method: Available modulations methods show different robustness against environmental influences;

Distance: The field strength of the received signal at the receiver depends on the distance between time signal transmitter and recipient;

Precision and stability of carrier frequency: As higher the precision and stability of the carrier frequency as smaller the threshold can be chosen.

Each recipient in a time transmission system is subject to individual conditions and parameters. Therefore, the threshold value should be determined empirically. For the test environment used for the result of this paper the threshold value was fixed to $t_{th} \approx \pm 2 \text{ ms/min}$. This value ensures that even after 10 minutes of missing time information a time verification is successful. This threshold time was confirmed over several tests in an interval over more than 3 weeks with different environmental influences of rain, wind, thunderstorm, etc.

The precision and stability of the carrier frequency has to be monitored by the recipient, because undetected manipulations of the frequency would allow slow manipulations of the time information caused by a couple of periods, which could sum up over many successful verifications.

7. Summary, Conclusion and Future Work

The main topic of the paper is the presentation and test of a universal verification method for time signals

and time telegrams for different modulation schemes in wireless time distribution systems.

The paper shows the implementation and measurement results for the German DCF77 time distribution service.

It was shown that the presented verification method could be adapted directly for time distribution services which use different kinds of modulation. With the help of this verification method, MITM and Replay attacks can be detected.

In the next step, the usability of the present method should be tested in detail for GNSS like GPS, Galileo, etc. Therefore it is required to develop a GPS or Galileo receiver prototype.

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(006)

Precise Frequency and Time Transfer for the Square Kilometer Array

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Summary: The Square Kilometer Array (SKA) is a next-generation radio astronomy facility that will revolutionize our understanding of the Universe and the laws of fundamental physics. This paper describes a Synchronization and Timing (SAT) system used in the Square Kilometer Array (SKA).

Keywords: Synchronization and Timing (SAT), Timescale, Co-ordinated Universal Time (UTC), One Pulse per Second (1PPS), H-Maser, UTC(SKA), White rabbit.

1. Introduction

The Square Kilometer Array (SKA) is a next-generation radio astronomy facility that will revolutionize our understanding of the Universe and the laws of fundamental physics. It will have a uniquely distributed character: one observatory, operating two telescopes, on three continents for the global scientific community. The SKA's science goals are broad and ambitious, looking back into the history of the Universe as far as the Cosmic Dawn, when the very first stars and galaxies formed, to seek answers to some of the biggest remaining questions in astrophysics. The SKA will be constructed in two phases: SKA1 (10 %) and SKA2 (90 %) [1].

Construction for SKA1 is expected to begin in 2021 on two sites, Australia (SKA1-Low) and South Africa (SKA1-Mid). The two SKA telescope sites are radio-quiet zones. In comparison to existing state-of-the-art telescopes, the SKA provides a significant leap

in sensitivity, resolution and survey speed, the three principal measures by which astronomers assess performance. The SKA1 radio telescopes will provide continuous frequency coverage from 50 MHz (6 m wavelength) to 15.4 GHz (1.99 cm wavelength). This paper focuses on the performance requirements for synchronization and timing (SAT) systems used in the SKA1 telescope. It also describes how these systems achieve stringent performance requirements in practice [2, 8].

2. The Architecture of SKA1 Telescope

The SKA1 is a distributed observatory with radio telescopes at two different sites. SKA1-Low is located in Australia, SKA1-Mid in South Africa and SKA Global Headquarter (SKA-HQ) is in the United Kingdom (UK) as shown in Fig. 1.

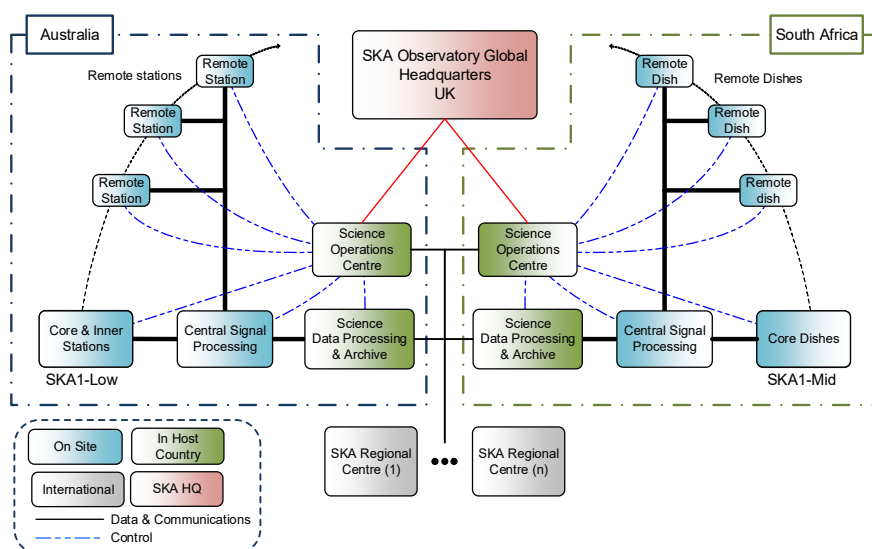


Fig. 1. A schematic diagram of the SKA Observatory, showing the site entities (telescopes), the entities at Host Country centers (Host Country Operations, Science Data Processing & Archiving), and the globally located SKA1 Regional Centers. (courtesy: Peter Dewdney, SKAO Architect).

For both the telescopes there is individual Central Processing Facility (CPF), Science Data Processing facility (SDP) and Science Operation Centre (SOC) located in the host country.

The SKA1-Low telescope will be located on a remote site at Boolardy Station in the Murchison Shire of Western Australia and the SKA1-Mid telescope will be located in the Karoo district in the Northern Cape Province of South Africa. SKA1-Low will operate from 50 MHz to 350 MHz whereas SKA1-Mid will operate from 350 MHz to 15.4 GHz. The SKA1-Mid telescope will consist of an array of reflector antennas (197 fully steerable dishes including 64 dishes of currently functioning MerrKAT) with a maximum baseline of 150 km. The SKA1-Low telescope consists of 512 stations (steerable electronically) with a maximum baseline of 70 km. Each station includes 256 dual-polarized log-periodic antennas. All the sky signals captured by both SKA1-Mid and SKA1-Low are digitized and timestamped on the field before being transmitted to the respective central processing facility (CPF) [2].

2.1. Synchronisation and Timing (SAT) System Architecture

The sampling clocks for the digitization process at remote receptors will be generated from a highly stable reference frequency source maintained at the CPF for each telescope. Similarly, the timing signals at each remote receptors will be generated from timescale maintained at the CPF for each telescope. The SAT system in each telescope is responsible for generation and distribution of reference frequency and timing signals across the receptors. Ideally, the reference frequency stability at each receptor should be equivalent to that of the stability provided by Hydrogen Maser (H-Maser). Installing an H-Maser at each receptor for clock generation is a costly and impractical solution. The more convenient and cost-effective option is to maintain a single source of the reference frequency and timing signal at a common location and distribute them to every receptor in the field to maintain a station time and reference frequency (STFR). This is common practice in radio telescope arrays.

The common overview of the SAT system for each telescope is shown in Fig. 2. The SAT system consists of the following subsystems.

The *SAT.CLOCK* subsystem is responsible for generating and maintaining time and frequency standards for the telescope. This sub-system will be located at the central processing facility (CPF).

The *SAT.STFR.FRQ* subsystem is responsible for the distribution of reference frequency signal generated by *SAT.CLOCK* to the remote receptors. The *SAT.STFR.FRQ* subsystem for SKA1-Mid is developed by the University of Western Australia (UWA) [4] whereas for SKA1-Low it is developed by the Tsinghua University [5]. Both these subsystems are different from each other and the composition of them

solely depends on the engineering requirements for the respective telescope.

The *SAT.STFR.UTC* subsystem is responsible for the distribution of the accurate time signal generated by *SAT.CLOCK* to each receptor in the array. The *SAT.STFR.UTC* subsystem used in SKA1-Low and SKA1-Mid telescope is similar and developed by the Joint Institute for VLBI (JIVE) [7].

The *SAT.LMC* subsystem is responsible for local monitoring and control (LMC) of the health of the SAT system and it reports to the central telescope manager system. The scope of this subsystem is outside of this paper.

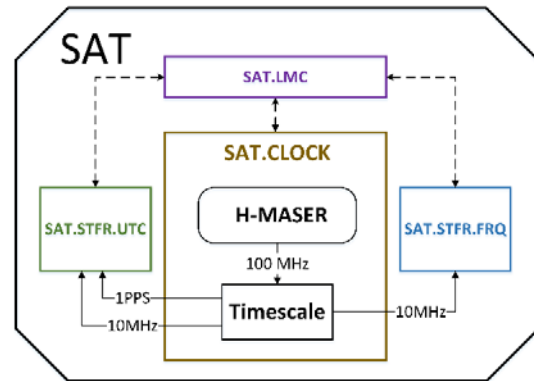


Fig. 2. Synchronization and Timing system (SAT) overview.

3. Performance Requirements for a Synchronization and Timing (SAT) System

In general, the performance of a radio telescope is specified by parameters such as sensitivity, angular resolution, a field of view, frequency coverage, bandwidth, etc. The unprecedented sensitivity of SKA1 is an important aspect to achieve various science objectives. The capacity of the interferometric radio telescope (sensitivity) to capture the maximum possible power from an astronomical signal depends, among other factors, on the frequency stability of the reference clock at the antenna location. In the same way, timing accuracy will affect the capability of the telescope to do precise time stamping of the astronomical signal (needed for Pulsar science) and to keep the synchronization of the system. The following system performance requirements are derived from the science requirements.

The *SAT.STFR.CLOCK* subsystem must provide and maintain a stable frequency standard for Local Oscillators (LOs) and digitizer clocks to support VLBI science. It is also responsible for maintaining time standards for the telescope (SKA-Timescale) with a maximum rms error of 5 ns and 9 ns from UTC for SKA1-Mid and SKA1-Low respectively.

The *SAT.STFR.FRQ* subsystem must distribute the reference frequency to the remote receptors with coherence loss less than 2 % for every single baseline

and with an output linear phase drift less than a radian over 10 minutes at the maximum operating frequency. The coherence loss requirements can be represented in terms of Allan deviation for measurement purposes [3].

The SAT.STFR.UTC subsystem must distribute time maintained by SAT.CLOCK subsystem to each receptor location with rms error less than 2 ns and with the pulse-to-pulse scatter (jitters) less than the inverse of the Nyquist frequency of the largest sampled bandwidth.

The synchronization and Timing (SAT) system developed to be used in the SKA1 is unique in terms of the performance it achieves over a very large area and in the presence of harsh environmental conditions.

4. Fibre Cable Network Architecture

The SAT distribution subsystems for both the telescopes distribute a reference frequency and time signal using optical fiber cables. For SKA1-Mid all the receptors (dishes) are connected to the Central Processing Facility using a combination of buried and overhead aerial fibers. Because of exposure to various environmental conditions such as temperature gradients, wind gusts and humidity, these fibers produce independent and random phase fluctuations in the propagating signals. SKA1-Low telescope stations (consisting of 256 log-periodic dipole elements) beyond the central region are connected to a remote processing facility (RPF) and all the RPFs are connected to CPF using buried fiber cables. These fiber cables produce temperature-dependent random phase fluctuations in the signal propagating through it. Because of the large area, and diverse temperature

profiles throughout the day, each baseline may be exposed to a different temperature at a given time and could cause severe coherence loss and timing inaccuracy. For this reason, the SAT system includes round-trip control loops to compensate for these effects (see the next section).

5. System for Synchronization and Timing (SAT)

The SKA1 synchronization and timing system consists of a clock ensemble, an actively compensated frequency transfer network, and a White Rabbit based timing network. Various options have been evaluated for synchronization and timing across the array in terms of performance, cost, and feasibility of implementation.

The designs of the SAT.CLOCK and SAT.STFR.UTC subsystems are similar for both telescopes whereas the designs of the SAT.STFR.FRQ subsystems are quite different.

5.1. SAT.CLOCK Subsystem

The SAT.CLOCK subsystem is the backbone for the station time and frequency reference (STFR) for the telescope. The reference inputs for SAT.STFR.FRQ and SAT.STFR.UTC subsystems are from SAT.CLOCK. The design of the clock subsystem for each telescope is similar to that shown in Fig. 3. It uses clock ensembles of three Hydrogen Masers (H-Maser) connected in a three-cornered hat configuration [8], which provides the realization of the main and backup clock source.

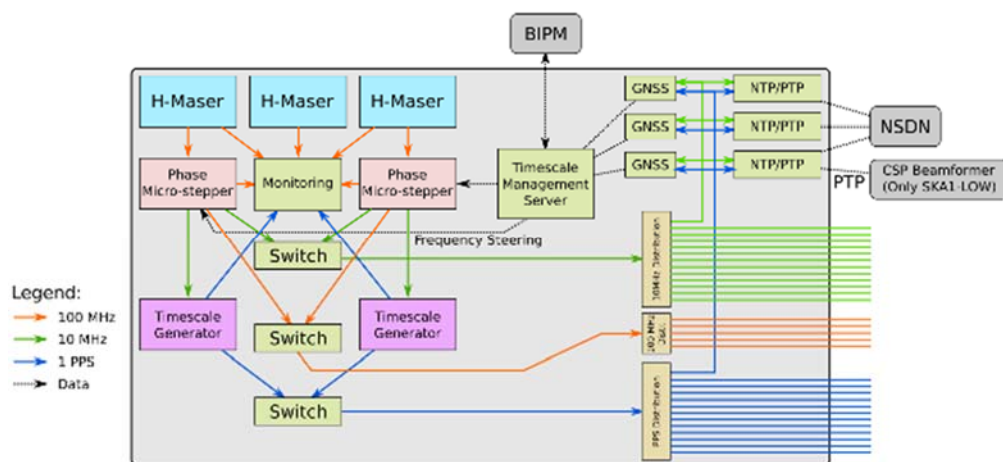


Fig. 3. Block diagram of SAT.CLOCK for both SKA1-Mid and SKA1-Low telescopes. GNSS global navigation satellite system; NTP network time protocol; PTP precision time protocol; BIPM Bureau International des Poids et Mesures.

The SAT.CLOCK subsystem provides a reference frequency of 10 MHz and 100 MHz along with a UTC aligned 1-pulse-per-second (1PPS), which is generated by timescale generator. SAT.CLOCK uses a unique component known as phase micro stepper to steer the

output clock of H-Maser. The phase micro stepper avoids the frequent steering of H-Maser output frequency directly.

The outputs of SAT.CLOCK are used by the distribution subsystems and also used by the Correlator

Beam Former (CBF). SAT.CLOCK also provides a network time service for all other auxiliary devices connected in the network using the Network Time Protocol (NTP). For lower time uncertainties in the traceability to UTC, the SKA-Timescale will participate in the Bureau International des Poids et Mesures (BIPM) group by implementing UTC instances, UTC(SKA) at each of the two telescope sites. BIPM is the internationally recognized body for coordinating timescales.

5.2. SAT.STFR.FRQ Subsystem

To produce a stable and accurate reference frequency signal at all the receptors, a unique closed-loop feedback control distribution system is needed which can keep a track of phase fluctuations in all the fiber links over the time and act to mitigate their effects to maintain stability and accuracy of the reference frequency signal at the receptor. Both the reference frequency distribution subsystems for SKA1-Low and SKA1-Mid are closed-loop feedback control systems [4, 5].

5.2.1. SAT.STFR.FRQ Subsystem for SKA1-Mid

For the SKA1-Mid telescope, a round-trip stabilized microwave frequency transfer using an optical phase sensing and actuation system is used which mitigates any noise-like fluctuations in electrical length between the transmitter and the receptor [4]. The frequency distribution subsystem for SKA1-Mid delivers a slightly frequency-offset signal from a nominal of 3.96 GHz reference frequency for each dish to mitigate self-generated Radio Frequency Interference (RFI). The block diagram for the frequency distribution subsystem used for SKA1-Mid is shown in Fig. 4.

It consists of a pair of transmitter and receiver modules for each dish receptor. All the transmitter modules will be located in the Central Processing Facility and there will be one receiver module located at every remote receptor. All the receiver modules are connected to respective transmitter modules using the fiber cable network described in Section 4. All the transmitter modules get input signals from a common rack cabinet which accommodate the optical source, frequency synthesizer, signal generator, a microwave shift and rack distribution.

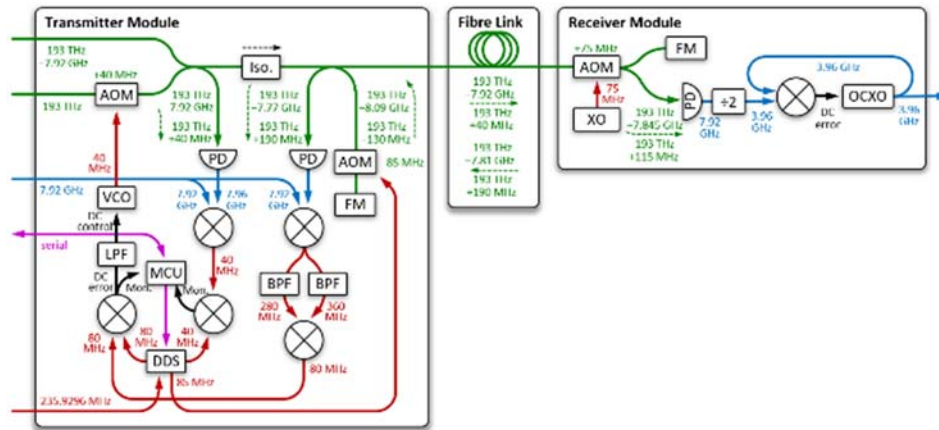


Fig. 4. Schematic diagram of the stabilized microwave-frequency transfer technique. Optical-frequency signals are shown in green; microwave-frequency signals in blue; radio-frequency signals in red; and the error signal in black. DPM dual-parallel Mach-Zehnder modulator; AOM acousto-optic modulator; Osc. Oscillator; Iso. optical isolator; FM Faraday mirror; PD photodetector; VCO voltage-controlled oscillator; BPF band-pass filter; and RF ref. radio-frequency reference (courtesy: Dr Sascha Schediwi, University of Western Australia).

The common rack cabinet gets the reference input 10 MHz from the SAT.CLOCK subsystem. The receiver module at each receptor sends a partially reflected signal back to the transmitter module. The transmitter module uses the partially reflected signal to perform all the measurements and actuates the corrections at the transmitter side to mitigate the phase noise over the link and deliver the stable reference frequency at the remote receptor. The receiver module also houses an oven-controlled crystal oscillator (OCXO) in phase-locked loop (PLL) configuration which acts as a clean-up oscillator. This clean-up oscillator removes the short term (time less than the

round trip delay of the optical signal) phase fluctuations which are also known as jitter.

Fig. 5 shows the performance of the SAT.STFR.FRQ subsystem in terms of Allan deviation tested on a 166 km metropolitan fiber link. These results are close to the reference frequency stability delivered by SAT.CLOCK. The SAT.STFR.FRQ subsystem for SKA1-Mid has been tested rigorously for various cable lengths, and various environmental conditions such as temperature gradient, wind gusts and humidity. The subsystem was also tested on a 186 km overhead fiber cable at SKA South Africa with a temperature range of 7.1 °C to

10.1 °C, a temperature gradient of ± 0.8 °C/10 min and relative humidity of 21 % to 97 %. The testing shows a maximum coherence loss of 0.07 % for 60s integration time [6]. The subsystem outperforms the 2 % requirement detailed in Section 3.

The phase drift at the output signal of SAT.STFR.FRQ is 0.08 radians for 10 min which is again very promising as compared to the 1-radian requirement mentioned in Section 3. The SAT.STFR.FRQ subsystem for SKA1-Mid is also compliant with the SKA's stringent electromagnetic interference (EMI) requirements.

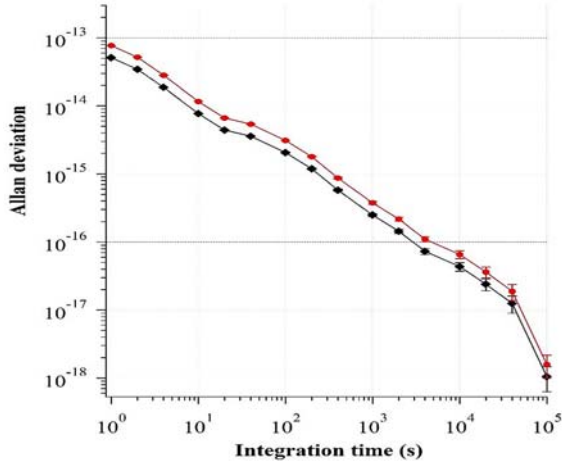


Fig. 5. Performance of the SAT.STFR.FRQ subsystem for SKA1-Mid in terms of Allan deviation as a function of integration time, measured over a 166 km metropolitan fiber link (black line) and extrapolated for two independent antennas with maximum fiber distance of 175 km (red line).

5.2.2. SAT.STFR.FRQ Subsystem for SKA1-Low

For SKA1-Low, a non-harmonic precision reference frequency synchronization scheme is used for the distribution of the reference frequency [5]. It was developed by Tsinghua University and is quite simple compared to the subsystem used in SKA1-Mid. It has a transmitter and receiver pair for every remote processing facility (RPF). All the transmitters are located at the central processing facility (CPF) and there is one receiver per RPF. Transmitters use a 100 MHz reference frequency signal from the SAT.CLOCK subsystem. SAT.STFR.FRQ subsystem for SKA1-Low distributes 10 MHz signals to the RPF using RF modulation at 1 GHz and 2 GHz as shown in Fig. 6. On the receiver side, a 100 MHz reference frequency signal is recovered and used to synthesize a 10 MHz signal.

The basic principle for SAT.STFR.FRQ for SKA1-Low is based on transmitting a 2-GHz signal modulated over a 1547.72 nm laser that will travel from the transmitter to the receiver. The receiver sends a 1-GHz signal modulated over a 1548.53 nm laser to the transmitter. At the transmitter, the signal will be bounced back over an optical circulator. On the receiver side, an array of mixers will combine the

single trip 2-GHz signal and the round trip 1-GHz signal and thus measure the phase fluctuations (in the form of an error signal) introduced by the fiber link. This error will drive a PLL which controls the output OCO delivering a 100 MHz signal which is divided to deliver required 10 MHz reference frequency. This configuration act as a clean-up oscillator. The performance of SAT.STFR.FRQ subsystem for the SKA1-Low shown in Fig. 7 has been tested in the lab using a fiber spool of 50 km kept in a temperature varying chamber. The tested coherence loss was less than 0.0001 % for 1 min integration time and phase drifts 0.0015 radians for the 10 minutes.

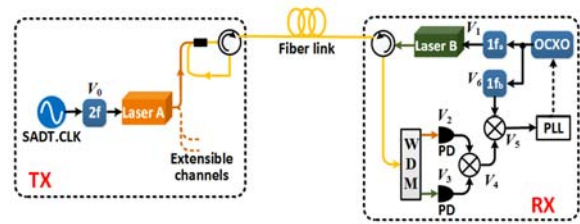


Fig. 6. Schematic diagram of SAT.STFR.FRQ subsystem for SKA1-Low. WDM wavelength division multiplexing; PLL phased lock loop; OCO Oven control crystal oscillator; PD photodiode; Tx transmitter; Rx receiver. (courtesy: Bo Wang, Tsinghua University).

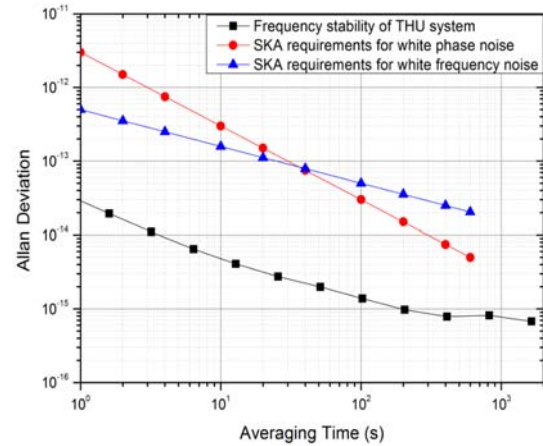


Fig. 7. Performance of SAT.STFR.FRQ subsystem for SKA1-Low in terms of output frequency stability measured in Allan deviation (black line and black squares), as compared to SKA requirements represented in terms of white phase noise (red line with red circles) and white frequency noise (blue line with blue triangles). The test was carried out in the lab with a 50 km fiber spool kept in a temperature varying chamber.

In addition, the SAT.STFR.FRQ subsystem for SKA1-Low has also been tested on 76.8 km overhead fiber at the SKA South Africa site. It has also undergone astronomical tests in the UK, operational environment tests and EMC tests in China. This subsystem meets the SKA requirements mentioned in Section 3 for coherence loss and linear phase drift.

5.3. SAT.STFR.UTC Subsystem

Both telescopes SKA1-Mid and SKA1-Low use the same subsystem for distribution of time signals from respective SKA-Timescales to every receptor. It shares the same fiber network which is used by SAT.STFR.FRQ subsystems; hence it encounters similar instabilities encountered by SAT.STFR.FRQ subsystems.

This subsystem carries out two functions. Firstly, it allows each recorded sample at each receptor to be tied to the SKA1 timescale with high precision, enabling e.g. pulsar timing. Secondly, it helps in the phasing of an array (or sub-array). A survey of various systems for time distribution has been done. Out of all the systems available, the White Rabbit (WR) system was found to be most appropriate for the SKA requirements. WR is a protocol for distribution of frequency and time over an Ethernet network with sub-nanosecond accuracy level. Originally developed by CERN, the WR protocol uses several commercial

off the shelf (COTS) components and hence it is more reliable and flexible. Fig. 8 shows a very basic WR link consisting a single WR master and single WR slave. WR works by synchronizing the clock of the slave device to the received timestamps of the master. It measures the round-trip delay between the WR master and the WR slave and retards the slave clock by the one way delay approximated from the round-trip delay. Originally WR was designed to achieve an accuracy of 1 ns over distances of up to 10 km. The WR design for SKA1 is augmented by the using longer fiber distance (up to 100 km), wavelength stabilized laser optics with external wavelength filters to meet accuracy over desired distance. WR repeaters are considered in the design to achieve a longer distance of 173 km [7]. It uses BiDi (bi-directional) Dense Wavelength Division Multiplexing (DWDM) small form factors (SFPs) to use only a single strand of bra for both-way signal transmission, which also helps to avoid delay asymmetries.

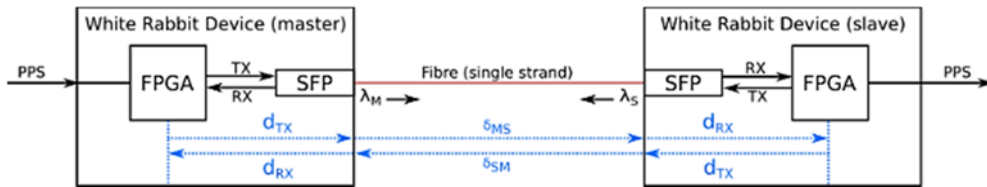


Fig. 8. Very basic white rabbit link connecting a single WR master to WR slave. d_{TX} , d_{RX} : Fixed, calibrated delays in a WR system, respectively in the transmitting and receiving path; δ_{MS} , δ_{SM} round trip delay or delay between master and slave, and vice versa; SFP Small form factor; λ_S , λ_M slave and master wavelengths. (courtesy: Paul Boven, JIVE).

The WR based SAT.STFR.UTC distribution subsystem for the SKA1 mitigates temperature dependent inaccuracies imposed by the optical fiber. It also mitigates the effects of dispersion, wavelength uncertainty inside the fiber and the Sagnac effect. The WR switches (WR-masters) are located at the CPF and get reference frequency and time (1PPS) signal from SAT.CLOCK. The WR endpoint (WR-slave) is located at each dish and each RPF for SKA1-Mid and SKA1-Low respectively. The WR-slave synchronizes its clock with WR-master clock and produces the 1PPS which will be used for timestamping data samples.

The SAT.STFR.UTC subsystem for SKA1 has been tested rigorously in the lab as well as on the field

using both buried and overhead aerial fiber. It has been tested for 140 km distance by cascading a calibrated WR link.

Fig. 9 shows the performance of WR for a 140 km long cascaded link. The plot shows (purple dots) the scatter between the rising edge of the transmitted 1 PPS and delivered 1 PPS with the spread of ± 100 ps, which is superimposed on the slower systematic variations (green curve), which are more clearly seen by averaging the measurement points by 100s. The WR based design of SAT.STFR.UTC for both the telescopes meets timing accuracy requirements.

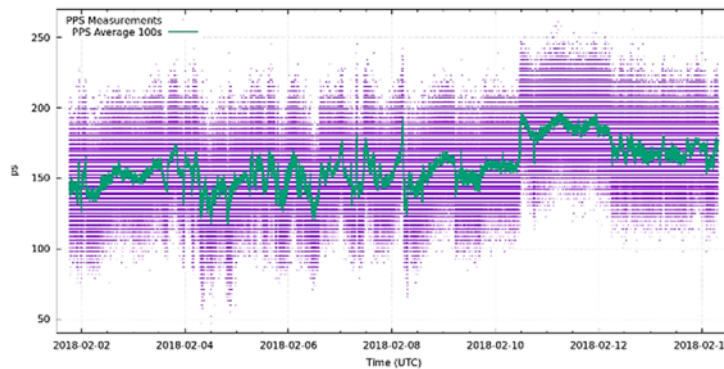


Fig. 9. Time jitters recorded on time interval counter during two weeks of continuous operation for WR performance on 140 km cascaded link.

Acknowledgement

The Square Kilometer Array (SKA) is an ambitious project involving many institutions and efforts of many people over the world. The systems presented in this paper are the results from the collaborative efforts of the international SKA SADT (Signal and Data Transport) consortium, one of 12 engineering consortia responsible for designing the look and functionality of the SKA's different elements and ensuring they all work together. The SADT consortium was led by the University of Manchester and supported by the University of Western Australia, Tsinghua University and the Joint Institute for VLBI (JIVE). The consortia brought together hundreds of experts from research institutions and industry around the world, whose talents and inputs were invaluable throughout the SKA's design phase and in preparation for SKA construction.

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Attenuation of Acoustic Waves in LiTaO₃ Crystals

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Summary: Anisotropy of the attenuation of acoustic waves in LiTaO₃ crystals have been studied by the Bragg light diffraction method in the frequency range from 0.4 to 1.6 GHz. The results were used to construct the acoustic attenuation surfaces. It is shown that the dielectric losses can influence on the magnitude and anisotropy of attenuation coefficient of piezoelectric longitudinal and transverse waves in the plane (010).

Keywords: Acoustic wave, Anisotropy of attenuation, LiTaO₃ crystals, Real and imaginary elastic constants, Bragg diffraction method.

1. Introduction

Lithium tantalate crystals (LiTaO₃), possess a rare combination of acoustic and optical properties thanks to which they are widely used in acousto-optic devices [1]. Nevertheless, the anisotropy of acoustic attenuation in these crystals is studied insufficiently.

It is known, the attenuation of acoustic waves in crystals due to the interaction of the wave with thermal phonons can be described phenomenologically by taking a complex elastic tensor of the form [2-4].

$$c_{iklm} + j\omega\eta_{iklm}, \quad (1)$$

where c_{iklm} is the lossless elastic tensor, η_{iklm} is the viscosity tensor for crystal, ω is the angular frequency of acoustic wave. Such consideration (Akhiezer mechanism of acoustic attenuation) is assumed that the frequency of acoustic wave is below that of any possible relaxations in the crystal. If the thermal phonon relaxation times are short with the period of acoustic waves, then the complex elastic constant can be used to describe the loss of acoustic wave's energy.

The real components of elastic tensor are known for great number crystals [1]). At the same time the imaginary components, describing attenuation of acoustic waves, were not practically investigated. The most detailed researches of imaginary components of elastic tensor have been made in [5, 6], but in a number of cubic crystals. The anisotropy of the attenuation of acoustic waves in LiNbO₃ crystals was studied in [3], however, were not taken into account the dielectric losses.

2. Samples and Experimental Technique

In the present work, the propagation velocities and attenuation coefficients of acoustic waves were

determined in LiTaO₃ crystals. Samples were prepared in the form of a cube with length of edges 12 mm. The faces of the cube have been oriented with accuracy no worse than 1° along the crystallographic axes [001], [100] and [010] or [001], [110] and [1-10], respectively.

The velocity and attenuation of acoustic waves were investigated with the help of Bragg diffraction of light on acoustic waves in frequency range 0.4-1.6 GHz at room temperature. Longitudinal and transverse acoustic waves were excited by piezoelectric transducers of quartz, accordingly of the X-cut or Y-cut. The helium-neon laser with a wavelength of 632.8 nm was used as a light source. The intensities of diffracted light were measured with the help of photoelectric multiplier.

The acoustic wave velocity V was determined from the Bragg diffraction angle with the accuracy of 0.2 % [4]:

$$V = \lambda v / 2 \sin \theta_B, \quad (2)$$

where v is the linear frequency of acoustic wave, θ_B is the external Bragg angle.

The anisotropy of the attenuation has been investigated for longitudinal and transverse acoustic waves propagating in the plane of symmetry (001) of these crystals. The attenuation coefficient was calculated with the accuracy ~5 % by the formula:

$$\alpha = 10 \lg(I_1/I_2) / 2L_0 \quad (3)$$

Here L_0 is the length of the sample, I_1 and I_2 are the intensity of the light diffracted on the direct acoustic wave and the intensity of the light diffracted on the wave, reflected from the free end of the sample, respectively.

3. Results and Discussion

Results of measurements of the velocity and attenuation coefficient of acoustics waves along the crystallographic directions are shown at Table 1.

Table 1. Results of measurements.

q	γ	V , $10^3 \text{ m}\cdot\text{s}^{-1}$	α , $\text{dB}\cdot\mu\text{s}^{-1}$ (1 GHz)
[100]	[100]	5,55	0.67
[100]	[001]	4,21	0.47
[100]	[010]	3,37	0.54
[010]	$\sim$ [010]	5,69	0.62
[010]	$\sim$ [001]	3,88	
[010]	[100]	3,53	0.52
[001]	[001]	6,16	0.36
[001]	[010]	3,60	0.62
[011]	$\sim$ [011]	5.90	0.50
[011]	[100]	3.55	0.73

Based on the received results and data from [7] all the real and imaginary components of the complex tensor of elastic constants c_{iklm} have been calculated. The obtained values were used to describe graphically the anisotropy of acoustic attenuation in the LiTaO_3 crystals. The cross-section of the surface of the attenuation of longitudinal and transverse acoustic waves by the plane ZOY is presented at Fig. 1.

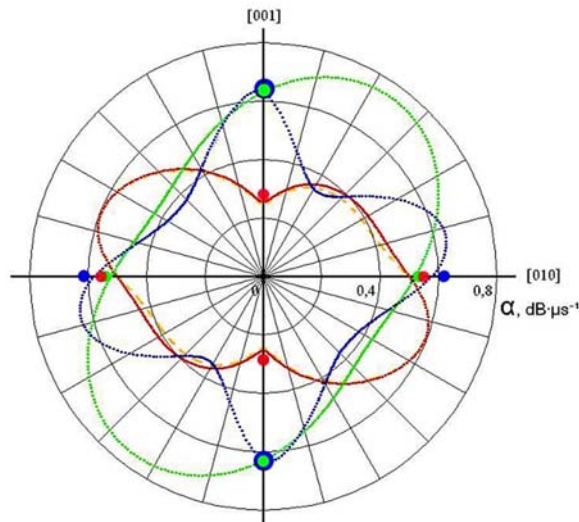


Fig. 1. Cross-section of the attenuation surface of quasilongitudinal (red), quasi-transverse (blue) and transverse acoustic waves (green) by the plane ZOY in LiTaO_3 crystals. Dots are experimental values of acoustic attenuation.

It should be noted that lithium tantalate crystals are characterized by relatively low anisotropy of elastic properties. These results can apparently be explained

by the fact that transverse and longitudinal acoustic waves in crystals interact mainly with the same branches of the thermal phonon spectrum. The only exception is slow shear waves along the X axis.

The analysis shows that the most noticeable change in the acoustic attenuation for all types of waves occurs when the direction of propagation of acoustic waves turns from the axis [111] to the [001].

Thus the anisotropy of acoustic attenuation in crystals can be clearly represented by the dependence of the attenuation coefficient of longitudinal and transverse acoustic waves on the direction of propagation in a certain crystallographic plane, for example, in the (001) plane.

Our results also show that there is a relationship between the velocity anisotropy and the attenuation anisotropy of acoustic waves, which depends on the anisotropy of the real and imaginary components of the effective elastic constant, respectively.

4. Conclusions

It is shown, that in full accordance with the expressions 1 [2] there is an interrelationship between the orientation dependences of the velocity and attenuation of acoustic waves. With a decrease in the velocity of acoustic waves, the attenuation coefficient of these waves is increased. In the same way, the opposite dependence is observed.

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(009)

Design of a 4.2 GHz SMR Filter for Space Applications: Methodology and Practical Implementation

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Summary: The FABRICS project answers the European Space Agency demand for the development of RF filters based on Miniature Bulk Acoustic Wave Resonators. Filters based on integrated Bulk Acoustic Wave technology at C-band are designed, manufactured and tested in that context. The ambition of the project is to dramatically reduce the size and mass of the corresponding devices with improved electrical performances compared to the present filter technology deployed for space applications. As a final target, the project must enable the development of future miniaturized microwave modules for Space mission of the next future. The exploited technology is based on Aluminum-Nitride films deposited by sputtering onto Silicon, a solution well adapted to address the ESA specifications. In this presentation, we show elementary parts of the development and more particularly the demonstration of building blocks required to manufacture one of the targeted filters operating at 4.2 GHz with a 72 MHz bandwidth. Elementary material stack is presented with some design results. The manufacturing technology is also reported and resonator responses are shown, enabling the extraction of resonator figures of merit. As a conclusion, further steps for completing the filter manufacture are discussed.

Keywords: BAW, SMR, Resonator, C-band, Filter, Aluminum Nitride.

1. Introduction

Film Bulk Acoustic wave Resonators (FBARs) [1] and more particularly the Solidly-Mounted-Resonator (SMR) based version are now well adopted as representing an essential part of front-end module filters of modern telecommunication systems and as representing a very favorable solution for C-band filtering demands [2, 3], particularly for the coming 5G standard but also for Space applications. The SMR structure was also selected because of its relative simplicity, compactness and robustness compared to FBAR based on air-gap reflection [4]. Although Surface Acoustic Wave filter technology has proven to be very successful in many telecoms applications, the resolution of SAW industry standard tool for lithography yields ultimate acoustic wavelength of

1.4 μm for which, considering wave propagation velocity of about 4200 m.s^{-1} , a maximum frequency of 3 GHz can be reached. BAW devices actually allow for accessing S and C frequency bands (2-4 and 4-6 GHz) using Aluminum Nitride or derivative materials developed as sub-micron films on Si wafers according the micro-electronics manufacturing standards. They provide advantageous solutions for very compact and high quality filters, capable to exhibit low losses (below 3 dB) and incident signal power robustness with acceptable thermal drift (less than -30 ppm.K^{-1}). With the aim of replacing payload input cavity channel filters which have significant size and mass by new technologies, BAW filters are very promising for optimal integration together with high spectral selectivity (see Table 1).

Table 1. Performance comparison between cavity filter, planar filter and BAW filter (++/+ : Top performance; 0: Median performance; -/-: Low performance).

Filter type	Losses	Power	Footprint	SMT	Quality factor Q	Rejection	Large Band	Narrow Band
Cavity	++	++	--	--	++	++	+	++
Planar	+	+	+	+	0	+	++	0
BAW	0	0	++	++	+	+	+	+

In this contribution, we discuss in detail the methodology for developing an AlN-based SMR filter operating at 4.2 GHz with 1.5 % bandwidth ($\sim 70 \text{ MHz}$) corresponding to an effective application for Space telecommunication satellites (see Fig. 1). As designing

such a filter requires an accurate knowledge of the constitutive material properties, we first present the corresponding characterization effort and results. Technological building blocks then are detailed showing intermediate resonator response and

parameter extraction to consolidate the design. Before starting the manufacturing process, a final design consolidation is achieved combining electromagnetic and acoustic simulations showing the capability of the device to address the requirements. Several aspects of the device manufacturing are detailed to highlight the critical steps to provide the final product.

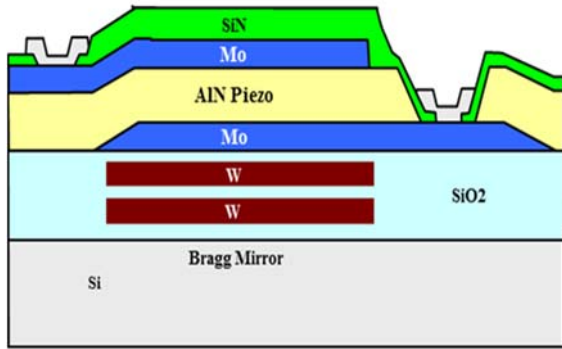


Fig. 1. Typical structure of one SMR resonator.

2. BAW Resonator Structure and Design

In the case of SMR structure, the resonator is deposited on a Bragg mirror, which consists in an alternation of materials with high and low acoustic impedances, in order to confine the waves and improve the quality factor of the resonator. These resonators will be assembled to constitute classical ladder filter structures in which series and parallel branches are combined to achieve the filter function along well-known design rules [5]. The design phase of the project is leaded by CEA-Leti with *freqn|sys* assessing in parallel its design toolbox, initially dedicated to SAW filters.

Firstly, the Bragg mirror uses a combination of 2.5 pairs of Silicon Dioxide/Tungsten layers, taking advantage of the acoustic contrast between this two materials (see Fig. 2). The layer thicknesses are adjusted to superimposed the shear and longitudinal wave-related reflection coefficients at the central frequency of the filter, 4.2 GHz. As we can see in Fig. 3, the maximum reflection of these two waves must actually coincides with the filter center frequency to yield the best energy confinement within the BAW resonator.

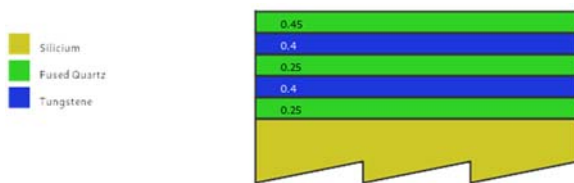


Fig. 2. Final layer thicknesses of Bragg mirror:
SiO₂ = 250 nm, top SiO₂ = 450 nm and W = 400 nm.

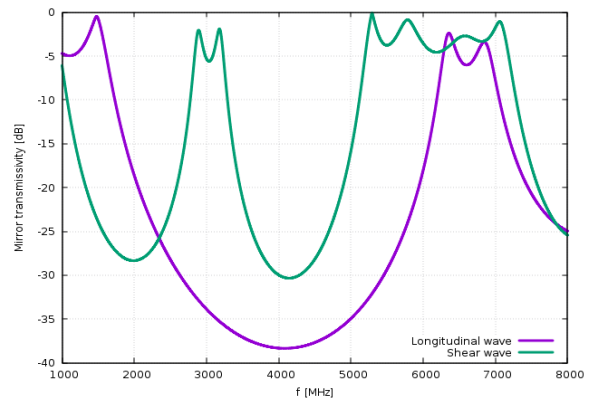


Fig. 3. Reflection coefficients of both longitudinal and shear waves.

Then, the next design step consists in optimizing the other layers of the stack: the electrodes and the Aluminum Nitride layer. Obviously, each new layer added in the stack affects the previous ones which implies a new optimization of the whole thicknesses at each step. For the case of the electrodes, Molybdenum was chosen for both seed and top electrode as complying with the need for an optimal growth of the AlN layer to obtain the best piezoelectric properties possible. Finally, after converging to an optimal stack, this one is again perturbed by introducing a top passivation layer used for both protection and frequency trimming. Therefore, the frequency shift between series and parallel resonators will only be done by this layer. The remaining layers stay identical. On each of these design steps, we try to achieve the largest possible coupling factor and keep the desired resonance frequency. The series and parallel stacks are illustrated in Fig. **Error! Reference source not found..**

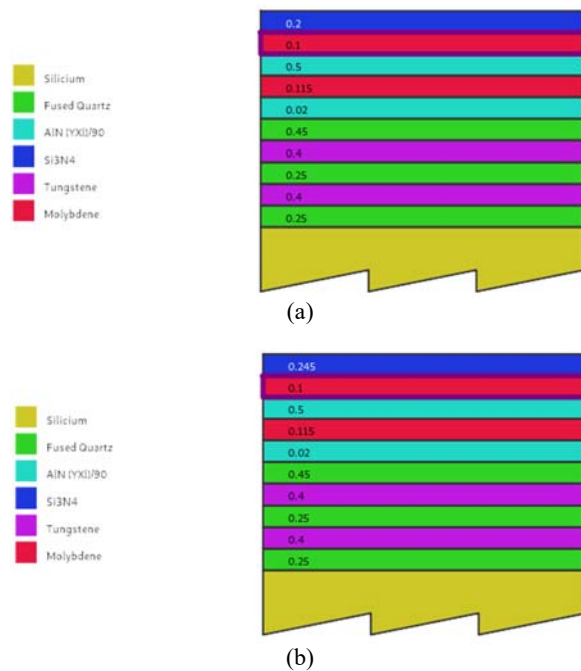


Fig. 4. Configurations of the two type of resonators composing the filter. (a) Series resonator, (b) parallel resonator.

To address the specifications, a 8-stage BAW ladder filter is considered. An optimization tool was developed based on the Particle Swarm Optimization (PSO) method to determine the surfaces of all resonators in the ladder filter. Using this method, we are able to quickly meet the theoretical response of the filter with the requirements (see Fig. **Error! Reference source not found.**).

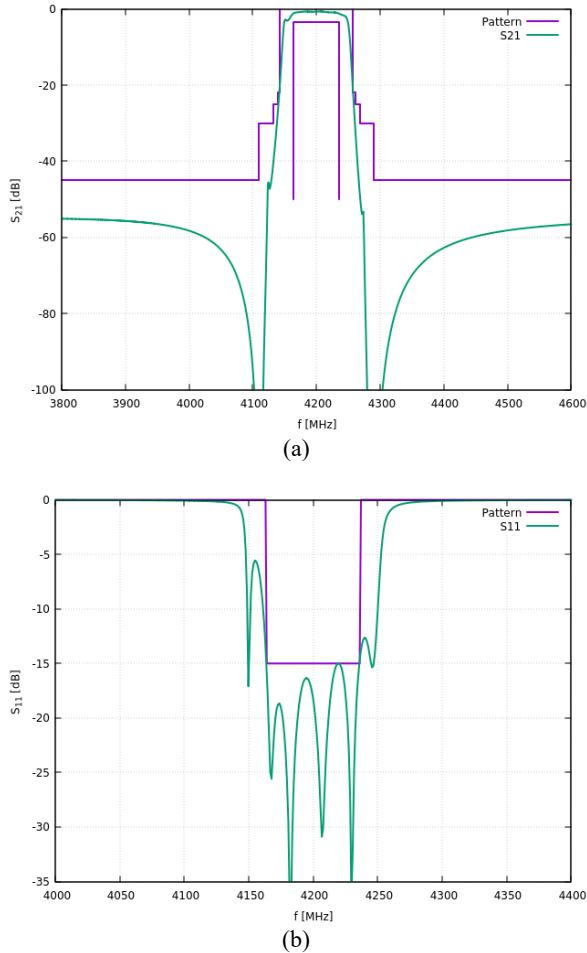


Fig. 5. Results after optimization process of surfaces. (a) S₂₁ transfer function modulus, (b) S₁₁ reflection coefficient modulus.

The purple curves depict the templates of transfer function (S₂₁ parameter) and reflection function (S₁₁ parameter) while the green curves are for theoretical predictions.

3. Device Fabrication

CEA-Leti has already developed a strong know-how of such devices and associated manufacturing [6]. Therefore, CEA-Leti mainly drives this activity, but freq|n|sys develops in parallel the same structure, excepted the Bragg mirror, to validate the company capability to manufacture BAW filters using its manufacturing suite.

Before starting the effective manufacturing of the test vehicle, a first iteration of deposition layers is performed to characterize the materials using pico-second microscopy. Then, these data are used to update the above-presented design.

Ten 200 mm Bragg mirrors provided by CEA-Leti have been downsized to be processable on freq|n|sys pilot line. Unlike CEA-Leti, we do not yet intend to pattern the Bragg mirror nor the bottom electrode although it is clear that it will be required to avoid parasitic capacitance which could notably affect the filter operation. The deposition of Mo was not available in our facilities at the beginning of this part, so the stack of the resonator was slightly modified to achieve preliminary tests. A Ti/Pt (5 nm/150 nm) bottom layer was deposited on the SiO₂ top layer using a sputtering process to allow for providing a textured metal layer compatible with best AlN growing conditions. A 450 nm thick Aluminum Nitride was then grown onto this Pt layer. Finally, an Aluminum top electrode was pattern using a lift-off process to limit the piezoelectric layer under-etching. The whole new material stack is illustrated in Fig. **Error! Reference source not found.**

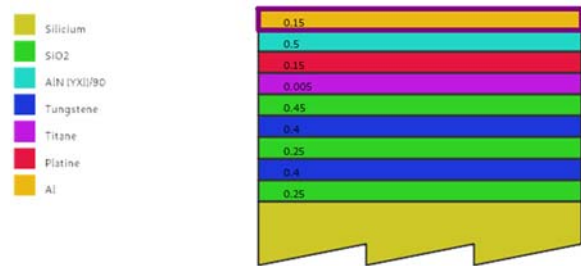


Fig. 6. Scheme of freq|n|sys material stack configuration for single resonator manufacturing.

Several test vehicles have been selected and manufactured to extract resonator and stack figures of merit (FoM). These different types are shown in Fig. 4.

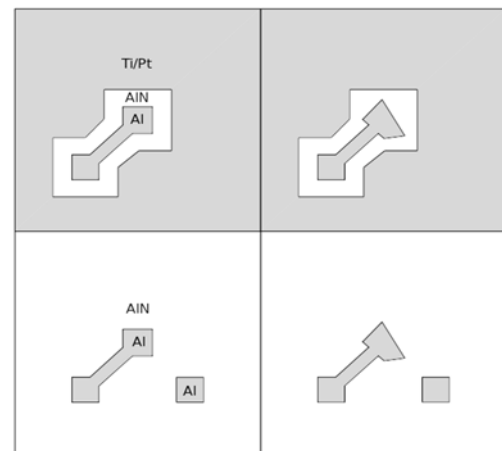


Fig. 4: Some patterns of the new test vehicles used for single resonator manufacturing and characterization.

An electrode with non parallel sides must avoid the presence of parasitic modes in the response of the resonator. Etching AlN layer was estimated mandatory to provide electrical contribution large enough for characterization purpose when exciting the modes compared to the use of two electrodes on AlN top surface (operating as two resonators in series). It turned out this assumption was not effective and etching the AlN layer was not required at this development stage (see. Fig. 6). Various top electrode surfaces have also been designed ($4000 \mu\text{m}^2$, $25000 \mu\text{m}^2$ and $62175 \mu\text{m}^2$) for impedance matching according to various thicknesses of AlN layer. Fig. 5 shows an optical microscope image of one of these test vehicles without patterning of the AlN layer.

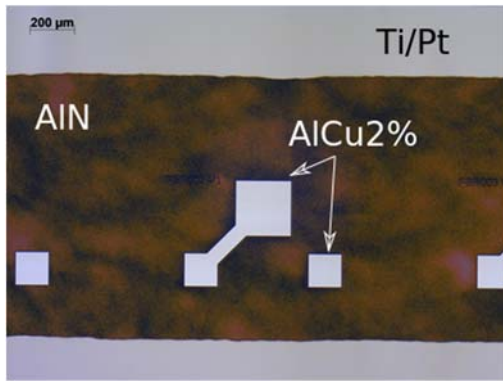


Fig. 5. Optical microscope image of a single port resonator used as test vehicles for demonstrating the effective operation of AlN thin films grown in freqn/sys cluster, without AlN etching near the top electrode, excitation is achieved by electrostatic influence in the ground pad.

4. Measurements

Characterization results have been obtained for all kinds of structures. An example of conductance measurement obtained for a $4000 \mu\text{m}^2$ resonator

(reference FBR006) with non parallel sides and no access to the bottom electrode is reported in Fig. 6. Each type of devices is measured at 5 spots on the wafer: top (purple curve), left (green curve), center (blue curve), right (orange curve) and bottom (yellow curve).

A frequency shift can be observed in Fig. 6 between the measurement in the center of the wafer (blue curve) and elsewhere on it. This is due to a non uniform thickness of Ti/Pt in this region. A 20 % increase, at least, of this thickness is measurable here compared to the remainder. In all the cases, the signal was resolved enough to allow for deriving principal FoM of the resonator, i.e. resonance and anti-resonance frequencies providing an estimate of the electromechanical coupling factor k_t^2 and resonance and anti-resonance quality factors Q_r and Q_a , respectively. These two coefficients can be deduced from the full width at half maximum of the conductance and resistance, but also from the calculation proposed in [7]. The results obtained along this later approach are presented in Table 2.

These results give a representative overview of the resonators operation. All devices were found to operate correctly, although measured quality factors were found still modest compared to those obtained on CEA-Leti side. It is however important to mention that we did not make an effective effort to de-embed the resonator response and the contribution due to the pads which yields parasitic capacitance and resonance dispersion that can also partly explain that the coupling factor does not reach values higher than 4.78 %. Whatever, this coupling is large enough to enable the fabrication of the project filter although the measured quality factors would not allow yet for meeting all the filter requirements (transition bands, minimum losses and ripples...). We also note that despite the use of specific electrode shapes in that purpose, spurious modes were not totally suppressed and contribute at frequency 5.5 GHz to degrade the filter spectral purity.

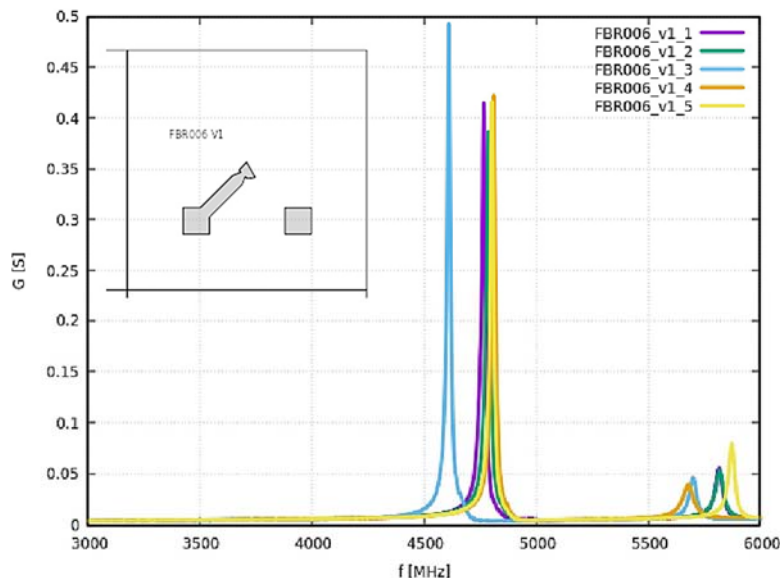
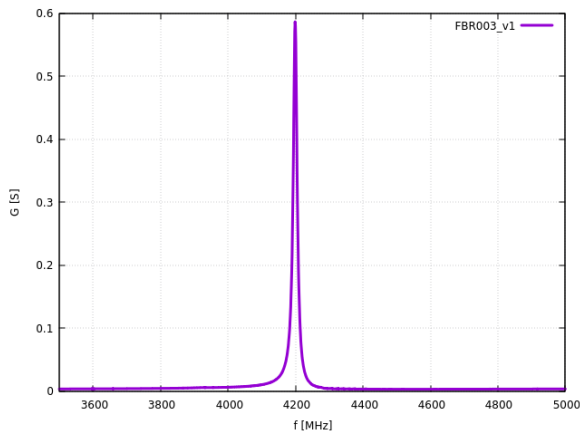


Fig. 6. Characterization of a 4000 μm^2 resonator with non parallel sides, no access to bottom electrode. 5 resonators measured: cardinal points + center.**Table 2.** Resonator FoM for the FBR006 test vehicle type.

	f_r [MHz]	f_a [MHz]	k_t^2 [%]	Q_r	Q_a
FBR006_v1_1	4766	4840	3.77	249	119
FBR006_v1_2	4788	4863	3.81	208	135
FBR006_v1_3	4611	4702	4.78	250	176
FBR006_v1_4	4812	4888	3.84	239	161
FBR006_v1_5	4804	4877	3.69	242	127

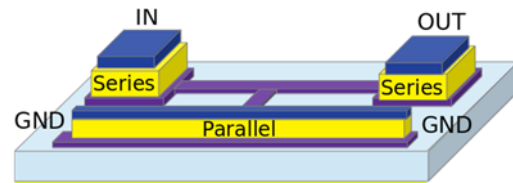
Whatever, before working on the manufacture of the filter, we need to modify the thickness of the AlN layer to fit the considered center frequency of the project (as a remainder, 4.2 GHz). After calculations, the new thickness is determined and a new deposition is done. The AlN layer is now 630 nm thick. The same devices as presented in Fig. 4 are manufactured on this new stack. Measures have been obtained for various resonators. An example of the response measured for a FBR006 resonator with parallel sides and no access to the bottom electrode is illustrated in Fig. **Error! Reference source not found..** As we can see, the resonance frequency is now around 4.2 GHz. These results allow us to move on to the next step of the project. Note that for this new test sequence, all the resonator FoM were improved with k_s^2 equal to 5 % and Q_r and Q_a above 300. We assume that these improvements are mainly due to a better control of the backside electrode and in general a strong effort to control stack layer uniformity.

**Fig. 10.** Characterization of a 4000 μm^2 resonator with parallel sides and no access to bottom electrode.

At this time, the overall structure of the filter can not be manufactured in the freq|n|sys facilities due to a lack of tools. Consequently, the work is concentrated on the development of filter building blocks. A first design will consists in a 1-stage ladder architecture with 2 series resonators and 1 parallel resonator (see. Fig. 7).

This design is optimized to lower interconnections and for probing. The shape and the surface of the

parallel resonator are especially stretched compared to series resonators to use 2-point probing. The goal is here validate the manufacturing process and check if the figures of merit meet the freq|n|sys computations. This flowchart consists in “only” 14 steps with one mask, one lift-off and two wet etches. This step corresponds to the next goal of the project.

**Fig. 7.** Scheme of the 1-stage ladder filter with 2 series resonators and 1 parallel resonator.

5. Conclusions and Perspectives

As a conclusion, we discuss here the road-map of the current development of RF filters based on miniature Bulk Acoustic Wave resonators for Space application and new communication standards. The exploited technology is based on a SMR structure with an Aluminum Nitride films deposited onto a Bragg mirror. The targeted filter operates at 4.2 GHz with a 72 MHz bandwidth. The first step of development consisted of the optimization of layers thicknesses to achieve the requirements in terms of center frequency, coupling factor, quality factors, etc. freq|n|sys design process was actually capable to converge toward a material combination providing the filter function looked up.

Then, the fabrication of single port resonator as test vehicles on CEA-Leti's Bragg mirror substrate has been achieved, allowing for characterizing the main figures of merit required to assess the capability to deposit thin AlN layers capable to meet the needed specifications to complete the filter. Without de-embedding, coupling factors were founded in the range 3-4 % whereas resonance and anti-resonance Q were lying in the 200-250 and 120-180 ranges, respectively, near 4.7 GHz. The AlN thickness was then corrected to reach the target frequency, with notable improvements of the coupling and quality

factors. Manufacture of new sets of resonators, firstly to demonstrate the capability to reach the targeted frequency and secondly, to implement filter elements as a new step to a completed filter manufacturing are planned to continue the development.

Acknowledgment

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