

New Construction of Conductive Fluids Velocity Measurement in Point - Using Measurement Over Interval Method

Feasibility Study

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Abstract— In this paper the method for measuring and reconstructing a signal from an electromagnetic probe, dedicated to the measurement of fluid velocity in the point, is proposed. The emphasis is on eliminating thermal noise and other interferences, consequently increasing the accuracy and precision of the (desired) measurements. The method is based on the principles of recently developed measurement over the interval method. The simulation results of the measurements completely support the theory upon which this method relies. It is concluded that the interferences are successfully digitally removed, and the measurement signal is completely reconstructed.

Keywords—fluids; measurement; noise; simulation; stochastic; velocity;

I. INTRODUCTION

Measurements of a fluid velocity represents a significant theoretical and practical problem nowadays. There are several different approaches and techniques for the measurement of a fluid velocity. These include measurements using Pitot tubes, Acoustic Doppler method, Laser Doppler method, Electromagnetic method and so on. In this paper we discuss the measurement of a fluid velocity “in a point”. Electromagnetic method represents the best approach for this type of measurement.

This method is based on the principle that a conducting fluid will generate an electromotive force (EMF) proportional to the flow velocity as it passes through the magnetic field created by the sensor (probe), [2]. For clarification, the EMF will actually be induced on the end of the electrodes which are in contact with fluid. This principle basically represents Faraday's law in its simplified form and it is graphically shown in Fig. 1.

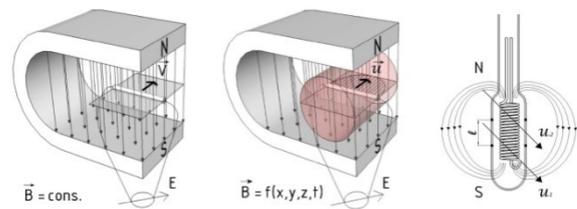


Figure 1. Electromagnetic induction principle (left), conductive fluid flow measurement using the electromagnetic induction principle (middle), conductive fluid velocity measurement “in point” (right) [4]

In the real life scenario the magnetic field is not homogeneous, yet it is variable in space and time. The velocity of the flow through a pipe of a certain diameter is not constant, yet the velocity depends on the spatial velocity field and time. So induced EMF will be integral :

$$E = \int_D \vec{B} \times \vec{V} \cdot d\vec{L} \quad (1)$$

In (1), E represents induced EMF, \vec{B} represents magnetic field, \vec{V} stands for fluid velocity and \vec{L} is conductor length (in this case distance between electrodes submerged into a fluid).

Induced electromotive force E involves a number of parasitic components which are superimposed to the measured voltage [3].

$$E = E_V + E_C + E_L + E_{DC} + E_{DIST} \quad (2)$$

In (2), E_V represents the signal due to measured velocity (within a target range of 0.1 microvolt to few microvolts), E_C and E_L are respectively capacitive and inductive disturbances, due to excitation signal. E_{DC} is a DC signal due to variable electrochemical water potential. The values of E_{DC} are often in range of a few volts. In the end E_{DIST} represents

remaining disturbances (leakage currents, sudden changes in electric conductivity, etc.).

In reference [1] it is explained that the electromagnetic principle can be used for a fluid velocity measurement “in point”. It also how it can be used for the average fluid velocity measurement in a relatively small control volume. The new method for velocity measurement “in point” is proposed by using *measurement over an interval method* [4-6].

The aim of this paper is a proof of that concept (proposal). As a basic tool for the verification of the proposed method, a computer simulation program has been specifically written for this purpose.

Chapter 2 gives review on »state of the art« solutions for electromagnetic measuring equipment. Chapter 3 discusses advantages of the Measurement over an interval method compared to Measurement in a point method. Chapter 4 deals with the simulation and the simulation results. And chapter 5 provides a discussion and conclusion.

II. STATE OF THE ART

For magnetic field formation, the first series of electromagnetic measurement devices were used 50 Hz sine waveform excitation signals. This method caused a number of serious problems. The problems were :

- the inability to separate the useful signal from the power supply noise (because they operate on the same frequency),
- the losses were too big,
- the interferences from slow electrolytic processes and so-called “zero shifting” were occurring [7].

Modern electromagnetic measurement devices use pulse excitation with square waveform signal of 8.33Hz frequency or 12.5Hz frequency (power grid frequency – 50Hz is an integer multiple of those frequencies). The reason for the use of pulsed excitation is reflected in the fact that the measuring probe is powered by the same excitation signal. Fig. 2 (to the left) shows a typical signal generated on the electrodes of the probe.

The measured signal carries, beside the useful signal, all the other disturbances. These are often an order of magnitude larger than the useful signal. Fig. 2 (to the right) shows individual components of a signal, with typical values that can be obtained from measurements in the water.

For velocity measurements it is necessary to wait for some time (usually 3/4 of half-period of the signal) initially to allow transients to settle down. This is followed by measurement of voltage E_1 in the first half-period, and E_2 in second. The mean value of measured voltages represents a DC component that should be subtracted from individual measured values to obtain the velocity value. Noise suppression is achieved by filtering. The order of magnitude of the time constant is typically a few seconds or greater.

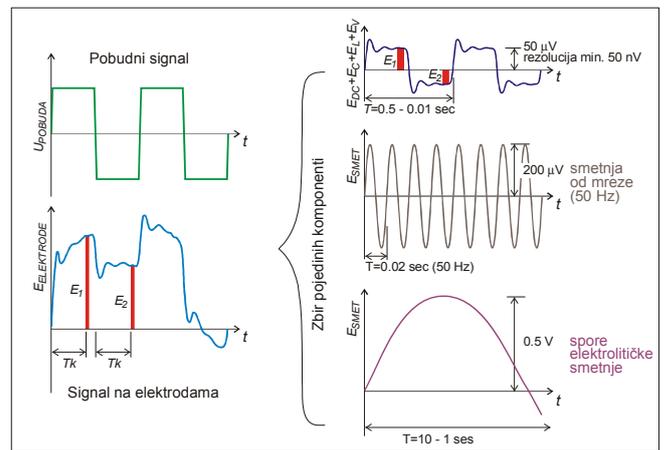


Figure 2. Excitation signal and signal generated on the electrodes (left), signal splitted into components (right) [1]

Fig. 3 shows the two-component electromagnetic probe commonly used. It is manufactured by a Serbian company, Svet Instrumentata (Belgrade). The head of the probe has a diameter of 18mm and it contains an excitation coil, four pairs of electrodes and a precise, low-noise amplifier which provides probe placement approximately ten meters from the rest of the electronics [1].

It can be concluded that the signal is periodic and only in part contains useful information about the velocity. It is customary to measure voltages E_1 and E_2 with a slow analog to digital converter (ADC) of great precision. ADC is synchronised with excitation signal and with a latency of T_k (Fig. 2) allowing satisfactory precision measurement despite the great interferences.

In [1] a slightly altered method is suggested. It was proposed to make a record of one period of the measured signal in the memory of the measuring device. Then, the signal is being processed, i.e. the method of harmonics measurement during one period of the useful signal is applied. When we have harmonics of one period of signal, simply by applying Inverse Fast Fourier Transform (IFFT) we can completely reconstruct the signal in that period.



Figure 3. Two-component electromagnetic probe [1]

III. MEASUREMENT IN A POINT VERSUS MEASUREMENT OVER AN INTERVAL

A. Measurement in the point

The term »measurement« almost always stands for discrete digital measurement, in other words measurement in point. In metrological slang, measurement in point is also known as sampling measurement method. There are two causes of

systematic error in a sampling method: discretization in time and discretization by value. If sampling theorem conditions are satisfied, discretization in time can be eliminated as cause of systematic error. Discretization by value always causes systematic error, and it is impossible to eliminate it, but under certain conditions it is possible to reduce it to an acceptable level.

The essence of the sampling method is: theoretically in infinitely short time interval, practically in a moment, the sample of analog measured value is taken, and in a time interval Δt , using ADC is converted into a number.

It is quite clear that equation :

$$\frac{1}{\Delta t} = f_s = 2f_h, \quad (3)$$

must be satisfied. In (4), f_h stands for the highest frequency of measured signal, or in the other words, upper limit of the signal's frequency range. It is therefore important for Δt to be as small as possible. The fastest are flash ADC's which have $\Delta t \approx 1ns$. The problem that occurs is flash ADC's small resolution. The resolution of ten bits is the maximum, in accordance with it, measurement uncertainty is large. It is well known that each additional bit of resolution doubles the hardware of flash ADCs, therefore the number of systematic error sources also doubles. From that point of view, it is much better that flash ADC has lower resolution. The problem with low resolution ADCs (less than seven bits) is reflected in the fact that Bennett's quantisation error model no longer applies, so quantization error cannot be treated as uniform white noise and it becomes serious theoretic and practical issue. To summarize: precise and accurate ADCs are slow, and fast ADCs are imprecise and inaccurate. This is central problem of the measurement in a point method – the extreme weakness (inaccuracy) on high frequencies.

Second problem of the measurement in point method is measurement of noisy signals. In the theory of discrete signals this is known as signal estimation among the noise, and theoretical approach does not take into account the quantization error (discretization by value). It is shown that a signal can be estimated better if sampling frequency is higher, and in this case fast ADCs become crucial.

The trend, not only in measurement in the point method development, but also in: telecommunications, management (control), power electronics etc. is development of fast ADCs with high resolution. The mathematics which describes the measurement in point is discrete mathematics and the theory of discrete signals and systems, and crucial mathematical tool is algebra. In case of noisy signals, the theory of random processes is applied.

For development of measurement in the point method, is not enough just to make a good AD converter. In order to obtain various parameters of the signal, it is necessary to process discrete signal values. The technological component, which provides rapid and efficient processing of discrete values is a digital signal processor (DSP).

Everything mentioned so far indicates that in the discrete digital measurement, methods and hardware are becoming standard. Regarding the signal parameters measurement, development of a narrow field of optimal measurement algorithms and signal processing is yet to be started. The general impression is, that in terms of research, there is not much to be discovered but just to standardize and apply what has already been found.

B. Measurement over an interval

Measurement over an interval represents complement of the measurement in the point. The most important feature of the measurement over an interval method is the ability to eliminate the limitations of measurement in point method and at the same time to retain almost all of its benefits.

Advantages of measurement over an interval method:

- measurement at high frequencies
- noisy signal measurement,
- high linearity and accuracy of measurements.

All three characteristics could be combined, so more accurate results could be obtained even in areas where it has not been the case.

In this method, ADCs with small resolution are used – flash ADCs, so the sampling frequency is practically maximal allowed by technology. For quantization error influence elimination, which is in this case significant, uniform random noise with mean value 0 is added to an input signal, in range of one quantum of applied flash ADC. It is shown that the quantization error then satisfies the conditions of Central limit theorem and the Sampling theory when the mean value of the signal is measured over an interval. Its standard deviation decreases with the square root of the number of samples in an interval.

Over an interval we may be interested in the effective value of the signal, so the measurement unit, in this case, is extending with: another flash ADC, additional uniform random noise generator and the block named multiplier-accumulator. If uniform noises on both ADCs are not correlated and the same signal is on both channel inputs, the mean of accumulator's content is equal to the square of effective value over an interval, and standard deviation of quantization error satisfies terms of Central Limit theorem and the Sampling theory. It is clear that standard deviation of quantization error is reverse proportional to the square root of the number of samples over an interval. For a very large number of samples effective value measurement error can be very small.

If, however, at the second channel of multiplier-accumulator's input is basis function from an orthonormalized set of functions, such as Fourier's, accumulator's mean represents the value of corresponding coefficient of signal expansion in orthonormalized set, so it is clear that the coefficients can be measured very accurately in this way.

The problem is that the correct representation of the signal over an interval requires (i.e. in the case of Fourier series expansion) a large number of coefficients (defined by

Sampling theorem). If understanding of the ideas is literal, measurement hardware can be very complicated and impractical.

Since a function of the second channel is known, and noise is also known, ADC can be easily simulated, so it is possible to prepare samples of basis functions over an interval and store them in memory. The memory output is then connected to a second input port of multiplier-accumulator, and multiplier-accumulator itself can then become a more complex structure (so there are several motives for its parallelization) so the greater speed of orthonormalized coefficient measurement can be achieved, significantly higher compared to the speed of modern DSPs. This structure has already been developed and named *Stochastic digital processor of orthonormalized transformation*.

It is noted that if the resolution of noised basis function samples is for at least two more bits higher than applied flash ADC resolution, then the upper limit of the standard deviation of orthonormalized coefficient measurement (arbitrarily orthonormalized transformation), depends only on quantum of applied flash ADC and applied transformation norm, and it doesn't depend on input signal waveform nor coefficient's magnitude and it is same for all coefficients. The consequence of this fact is that the coefficients of orthonormalized function can be obtained accurately and without the use of floating point arithmetic. This fact allows a drastic simplification of hardware for processing. It is shown that low resolution integer arithmetics is sufficient. In a prototype device for measuring the Fourier coefficients (harmonic) multiplier-accumulator of 6x8 bit integer was used, and in the second instrument for the same purpose 8x10 integer multiplier-accumulator.

There are an interesting facts about the mathematics that describe this measurement. The mathematical model has been deliberately omitted, because it is a much broader thing - almost a philosophy. We will try to briefly describe it: The mean value over an interval is the integral, in other words, the sum, and for addition commutation law applies, so it is absolute regardless of order in which we summarize elements of Darboux sum (deterministic or by an arbitrary sequence of uniform distribution). Thus, the time over an interval can be, with respect to mean values, or in other words, to an arbitrary point of distribution, treated as a random variable of uniform distribution. In this manner we placed the problem completely inside the field of Probability theory and field of Statistical theory of samples.

For the purpose of explaining the experimental results, especially ones achieved by simulations, because those made the majority of results, we use the Central limit theorem in a slightly more general form. This is something we have not seen in the literature, but thousands upon thousands of simulation experiments confirm the validity of central limit theorem. High-precision calibration equipment confirmed the accuracy of the derived formula.

C. Instrument for noisy signal harmonics measurement

Fig. 4 is showing block diagram of measurement instrument.

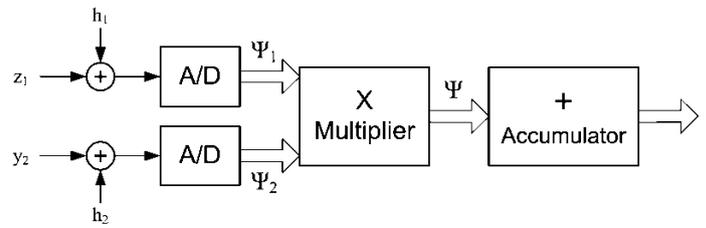


Figure 4. Block-diagram of the instrument for measurement of one harmonic of the noisy signal

Dither signals h_1 and h_2 are random, uniform, mutually uncorrelated and satisfy:

$$\begin{aligned} |h_i| &\leq \frac{\Delta_i}{2} \\ p(h_i) &= \frac{1}{\Delta_i}, \quad (i=1,2) \end{aligned} \quad (4)$$

where Δ_1 and Δ_2 represent quanta of uniform quantizers, and $p(h_i)$ is probability density function of random voltages h_i .

The inputs are $z_1 = y_1 + n$, $y_1 = f_1(t)$ and $y_2 = f_2(t)$. If $y_1 = f_1(t)$ and $y_2 = f_2(t)$ are integrable functions of distribution density $p(n)$, then the mathematical expectation of output, $\bar{\Psi}$, for time $t \in [t_1, t_2]$, can be written like:

$$\begin{aligned} \bar{\Psi} &= \frac{1}{t_2 - t_1} \cdot \int_{t_1}^{t_2} f_1(t) \cdot f_2(t) dt \\ &+ \int_{-R}^R n \cdot p(n) dn \cdot \frac{1}{t_2 - t_1} \int_{t_1}^{t_2} f_2(t) dt \end{aligned} \quad (5)$$

and measurement error is limited with:

$$\sigma_e^2 \leq \frac{\sigma_s^2}{N}, \quad (6)$$

$$\sigma_s^2 \approx \frac{\Delta_1^2}{4} \frac{1}{t_2 - t_1} \int_{t_1}^{t_2} f_2^2(t) dt, \text{ for } \Delta_1^2 \gg \Delta_2^2, \quad (7)$$

where N is number of samples in time interval $(t_2 - t_1)$, and Δ_1 is quantum of ADC on channel one.

For $\bar{n} = \int_{-R}^R n \cdot p(n) dn = 0$, which is mostly the case, upper limit of relative error Γ_u in measurement

$\bar{\Psi} = \frac{1}{t_2 - t_1} \int_{t_1}^{t_2} f_1(t) \cdot f_2(t) dt$ is given with :

$$\Gamma_t = \frac{\sqrt{\sigma_e^2 + \sigma_r^2}}{\bar{\Psi}} \leq \frac{\sigma_s}{\sqrt{N}} + \frac{\sqrt{\frac{1}{t_2 - t_1} \int_{t_1}^{t_2} f_2^2(t) dt} \cdot \frac{\sigma_n}{\sqrt{N}}}{\bar{\Psi}} = \Gamma_u \quad (8)$$

It should be noted that in addition to dither which is a pseudo-random process, AD conversion as a deterministic process dominates in the scheme.

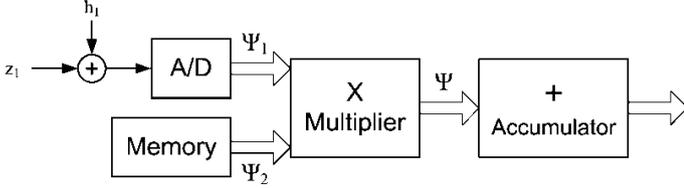


Figure 5. Modified block-diagram of the instrument

Problem solving by this method in literature has a name – estimation of signal parameters, and it is resolved in an idealized hardware environment. Most often used limit in a literature is Cramer-Rao Lower Bound (CRLB), [5]. So there is question asked about ratio of CRLB and limit (8).

Let the:

$$y_1 = f_1(t) = \frac{a_0}{2} + \sum_{i=1}^{\frac{N}{2}-1} (a_i \cos i\omega t + b_i \sin i\omega t) \quad (9)$$

be deterministic function on channel 1. It is obviously trigonometric polynomial. Since Weierstrass and Bernstein it is well known that continuous function can be approximated (with arbitrary accuracy) by trigonometric polynomial on the interval. If interval is given by $T = t_2 - t_1$. Then the next applies:

$$a_i = \frac{2}{T} \int_0^T f_1(t) \cos i\omega t dt = \frac{2}{N} \sum_{k=0}^{N-1} Y_1(k) \cos \frac{2\pi i k}{N} \quad (10)$$

$$b_i = \frac{2}{T} \int_0^T f_1(t) \sin i\omega t dt = \frac{2}{N} \sum_{k=0}^{N-1} Y_1(k) \sin \frac{2\pi i k}{N} \quad (11)$$

$$(i = 0, 1, 2, \dots, \left(\frac{N}{2} - 1\right))$$

where $Y_1(k)$ represents samples of $y_1 = f_1(t)$ function, and N is even and satisfies Nyquist's criterion.

In the literature it is explicitly assumed that the ADC is ideal ($\Delta_1 = 0$) $\Rightarrow (\sigma_s^2 = 0) \Rightarrow (\sigma_e^2 = 0) \Rightarrow (\sigma_r^2 = 0)$, so the limit (8) in that case becomes:

$$\Gamma_u = \frac{\sqrt{\frac{1}{t_2 - t_1} \int_{t_1}^{t_2} f_2^2(t) dt} \cdot \frac{\sigma_n}{\sqrt{N}}}{\bar{\Psi}} \quad (12)$$

Let the $y_2 = f_2(t) = R \sin i\omega t$, or $y_2 = f_2(t) = R \cos i\omega t$, then:

$$\bar{\Psi} = \frac{1}{T} \int_0^T f_1(t) R \cos i\omega t dt = \frac{R \cdot a_i}{2} \quad (13)$$

$$\bar{\Psi} = \frac{R \cdot b_i}{2}, \text{ applies.} \quad (14)$$

In these cases the:

$$\frac{1}{t_2 - t_1} \int_{t_1}^{t_2} f_2^2(t) dt = \frac{R^2}{2}, \text{ applies.} \quad (15)$$

If the noise is Gaussian noise, then $\sigma_n = \sigma$, so:

$$\Gamma_u = \frac{\frac{R}{\sqrt{2}} \cdot \frac{\sigma}{\sqrt{N}}}{\frac{R}{2} \cdot a_i} = \frac{\sqrt{2}\sigma}{\sqrt{N}a_i}, \text{ or,} \quad (16)$$

$$\Gamma_u = \frac{\frac{R}{\sqrt{2}} \cdot \frac{\sigma}{\sqrt{N}}}{\frac{R}{2} \cdot b_i} = \frac{\sqrt{2}\sigma}{\sqrt{N}b_i} \quad (17)$$

which is slightly altered form of CRLB presentation, and it is achieved by suggested instrument. The variance of measurement (estimation) a_i and b_i obviously does not depend from them, but it is constant and is given by:

$$\text{var}(\bar{a}_i) = \text{var}(\bar{b}_i) = \frac{2\sigma^2}{N} \quad (18)$$

Note that the limit of accuracy (8) is more general because it takes the real ADC into account and does not introduce any assumption about the nature of the noise. On the other hand, n , h_1 and h_2 have to be uncorrelated, N has to be great and condition $|f_1(t) + n| \leq R$ has to be satisfied.

Note that the limit (8) applies in case when „word“ of ADC (Ψ_1), „word“ from memory (Ψ_2) and „word“ from multiplier (Ψ) have small resolution, and can be implemented

to easily control the accuracy and parallel processing in the FPGA chip.

IV. SIMULATION AND RESULTS

As a key part of the feasibility study, the computer program is written to simulate the conductive fluid velocity measurements using the method of measuring over an interval.

A. The task of the simulation

Before the simulation of velocity measurement of the conductive fluid even starts, it is necessary to create faithful mathematical representation of the measured values. This primarily refers to the excitation signal, but also to noises which occur during the measurement. As it was said earlier in this paper, the excitation signal has pulse nature (square waveform) with a frequency of $(50/6)Hz$ (approximately $8.33Hz$) and phase of 0 radians. For simplicity, the amplitude of the excitation signal has the value of 1 V, and rest of the signals – noises which are simulated, are adjusted by the amplitude of the simulated excitation signal.

Interference that exists in the measurement are:

- interference from the power supply (50Hz sinusoidal signal)
- slow electrolytic disturbance (0.8Hz sinusoidal signal)
- white noise with Gaussian distribution (mean 0 and standard deviation 0.18).

Measurement signal is the sum of responses to the excitation signal, and these disturbances (of which the slow electrolytic disturbance has by far the largest amplitude). As for the response to the excitation signal, it should be noted that the ideal case of transmission characteristics of the measuring system is considered. In other words, in the presence of interference, considering that the environment does not distort the signal, the response to excitation is the same as the excitation itself.

It should be noted that the slow electrolytic disturbance is only used to display the total measurement signals, and it is not included in the digital processing, because of its nature (primarily large amplitude) must be filtered before entering amplifier and ADC block, using HP filter. Presentation of individual signals and the total measured signal are given in Fig. 6.

B. The relevant parameters of the simulation

In addition to simulated signals that participate in the formation of the measuring signal, it is necessary to generate two dither signals, which (as already mentioned) are mutually uncorrelated. Dithers have a uniform distribution, and their values are random and are in the range $[-2.5, 2.5]V$. Dithering frequency of the measurement signal in this case is $100kHz$.

What more should be taken into account is the total measurement time, which is directly related to the period of the measured signal.

For clarification, it is necessary to cover integer number of periods of the measured signal. In this particular case, the total measurement time is $6s$, which corresponds to the number of exactly 50 cycles of the measured signal. Sampling rate of the 2 bit flash ADC is $100kHz$, which corresponds to time increment of $\Delta t = 10\mu s$.

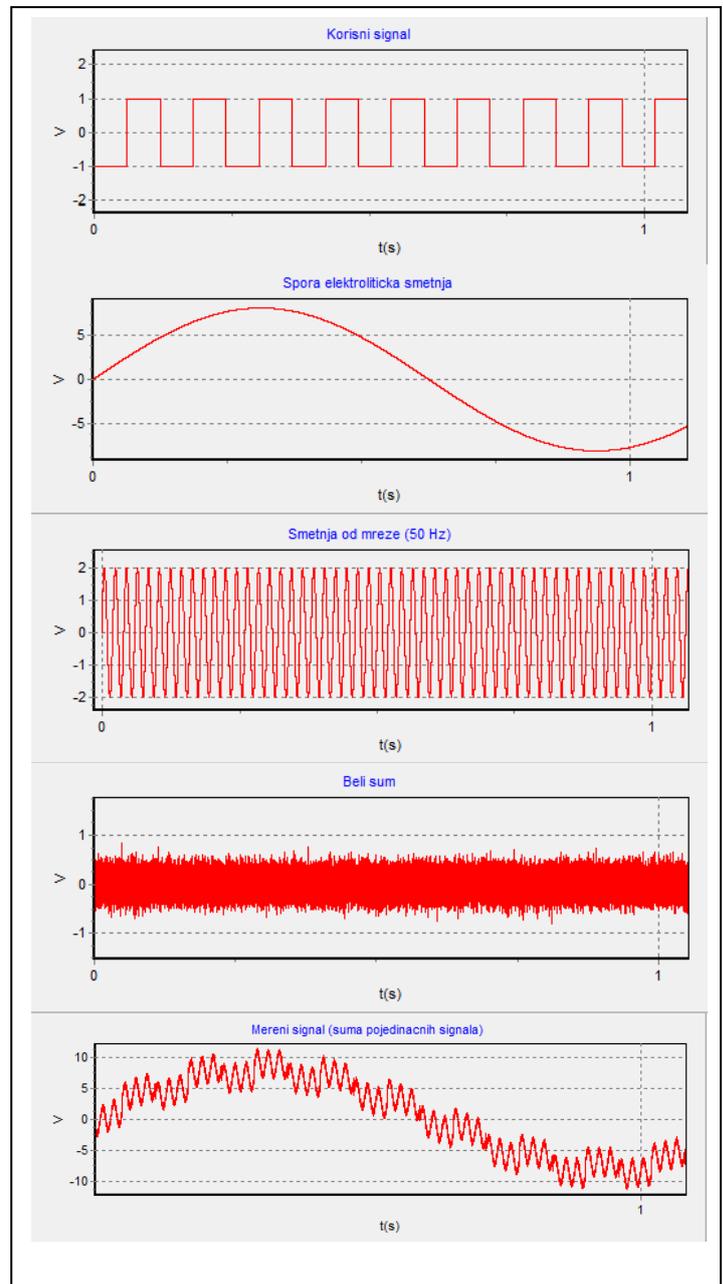


Figure 6. From top to bottom – excitation signal, 50 Hz disturbance, slow electrolytic disturbance, white noise and at the end measurement signal as sum of previous signals

Increase of the total measurement time, or increase of sampling frequency of the flash ADCs results in increased accuracy of measurement. Therefore it is of interest for these two parameters to be as large as possible, but within the

boundaries of realistic possibilities of the measurement hardware which needs to be simulated.

Two base signals (sine and cosine) are also simulated for each of 50 relevant harmonics, whose Fourier coefficients we want to calculate. The amplitudes of these basic signals are 5V and frequency in accordance with the harmonic whose coefficients are calculated.

C. The simulation results

At the end of 6 seconds of simulated measurements, inside the accumulators are values which are used for calculation of Fourier coefficients (sine and cosine). After all 50 sine and 50 cosine coefficients (harmonics) are calculated, it is possible to do so-called harmonic analysis or better put spectral analysis of the measured signal, shown in Fig. 7.

Sine coefficients are represented in red color on the figure, with blue cosine coefficients. The sinus nature of the signal comes to the fore, which is a consequence of the phase (of 0 radians) of the excitation signal.

Something that sticks out at the first glance is emphasized 6th harmonic of the signal, specifically its sinusoidal component. If you look closely, the value of this coefficient is 2V, and 6 times higher frequency than the frequency of excitation signal, which tells us that this is a 50Hz sinus interference that comes from the system's power supply. In order to obtain correct measurement results this coefficient should be excluded from the calculations. Ideally cosine coefficients (blue color in the figure) should not be there at all, though they are present because of the negligibly small errors caused by the method itself.

The next step of the simulation is the reconstruction of the initial signal. By applying inverse fast Fourier Transform (IFFT) on the obtained signal coefficients (leaving out the sixth harmonic) signal can be fully reconstructed in a given period of time - Fig. 8.

Statistical analysis of the measurement results of effective value of the reconstructed signal is made. Analysis was performed on a sample of thirty consecutive repeated simulations, and the results are shown in Table I.

The mean of these thirty measurements is 991.8 mV, the variance is 47.1 μV^2 , and standard deviation is 6.86 mV or 0.69%, (0.276% with regard to full scale of the ADC).

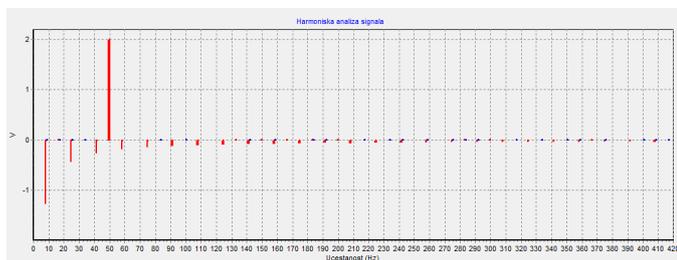


Figure 7. Spectral analysis of the measured signal – all 50 harmonics

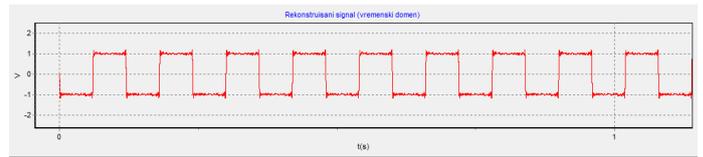


Figure 8. Reconstructed measured signal

TABLE I. RESULTS OF THE SIMULATED MEASUREMENTS

Results of the simulated measurements (values are in Volts)			
Ordinal number of simulation	Measured RMS value	Ordinal number of simulatios	Measured RMS value
1.	0.99035	16.	0.97837
2.	0.98619	17.	0.99343
3.	0.99638	18.	0.98779
4.	0.98761	19.	0.99357
5.	0.99809	20.	0.99875
6.	0.9936	21.	0.99108
7.	0.99325	22.	0.99958
8.	1.00265	23.	0.98429
9.	0.99308	24.	0.99801
10.	0.97558	25.	0.9812
11.	0.9931	26.	0.99702
12.	0.98602	27.	1.00216
13.	0.99602	28.	0.98896
14.	0.99609	29.	0.98304
15.	0.99906	30.	0.98996

V. DISCUSSION AND CONCLUSION

Based on the simulation results conclusions can be drawn that can greatly improve the overall system for measuring the velocity of a conductive fluid. We have seen in the prior discussion, the principles of operation of modern electromagnetic meters. In current practice - for measurement in the point it is important that the frequency of excitation is small, because in that case a period of response-signal is greater, ie. provides enough time for transients to settle down, and then the value of the signal is measured.

From the construction and performance standpoint, this approach has several disadvantages. The first is reflected in the large excitation current in the coil of the electromagnet. For clarification, a high current occurs as a result of low-frequency of the excitation (only 8.33Hz) and low resistivity of the excitation coil. In the case of reactive resistance, as is the case here, the lower the excitation frequency, the less the resistance and vice versa, in other words, resistance is directly

proportional to the excitation frequency. If we stick to the low frequency excitation, we can see that this is very undesirable, from the standpoint of noise originating from the supply voltage, especially if the power grid frequency is an integer multiple of the excitation frequency (as is the case here).

The proposal is to be imposed on the use of excitation which frequency is not an integer multiple of the power grid frequency (eg $(7 * 50 / 43) Hz$), because in this case yet 350th harmonic represents the disturbance from the power supply. And the 350th harmonic is completely irrelevant, given that only the first 50 harmonics are important for the signal reconstruction. Or, using the excitation frequency which is higher than the power grid frequency, and preferably sinusoidal excitation.

The use of sinusoidal excitation would bring a huge improvement to the measurement system, and especially from the harmonics measurement point of view. In that case only 1 harmonic would be measured, instead of 50 harmonic in the case of excitation with square waveform signal.

For the measurements over an interval method, the low frequency excitation becomes irrelevant, and therefore a number of advantages over the measurement in point are achievable, which are reflected in:

- a lower current in the driving circuit,
- possibilities of analog filtering of the slow electrolytic interference and interference from the power supply using high-frequency bandpass filter (only useful signal and the white noise remain).

This solution is not ideal, and without its own flaws, which are reflected in the complexity of the block for excitation of the

electromagnetic circuit. In fact, this proposal represents a compromise between, slightly more complex electronics for generating excitation signals (sinusoidal), on one side, and easier and faster (more efficient) analysis of the measured signal on the other side. From the performance and complexity viewpoint of the whole measuring system (hardware and software for measurement analysis and digital signal processing), this compromise is more than acceptable.

On the basis of the aforesaid, we can conclude that the interferences (50Hz disturbance, slow electrolytic disturbance and white noise) are successfully digitally removed and quality of reconstructed signal is satisfactory.

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