Sampling wattmeter at INTI

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Abstract: We developed an automated sampling reference system to measure electrical power. The system allows calibration of instruments within an uncertainty of 20 μW/VA at power frequencies. It can also measure harmonic power.

Calibration is fully traceable to national standards. All the components of the system can be separately evaluated and by this means its total uncertainty estimated.

Keywords: Dual- channel sampling, phase measurements, power standards.

1. INTRODUCTION

Electrical Power Laboratory at INTI developed in 1996 a thermal power comparator. Since that time, the primary power standard of Argentina has been based on thermal converters working with the well known sum and difference method [1]. It provides high accuracy, typically 20 $\mu W \cdot W^{-1}$ at power factor one [2]. With this system, good results were obtained in the corresponding CIPM key comparison [3]. The main disadvantage of the thermal power comparator is the time needed for one measurement, typically 180 s to allow the thermal converter to thermally stabilize in each of the three measurements [2]. Another disadvantage is that the thermal wattmeter cannot measure power of individual harmonics, because the thermal converter measures the heat dissipated by the sum of all the frequencies. To face these problems we developed a sampling wattmeter. It uses two digital multimeters (DMM) sampling in DC asynchronously with the signal [4,5].

2. DESCRIPTION OF THE INSTRUMENT

Fig. 1 depicts a diagram of the system use at INTI as a sampling wattmeter. As power supply we use a two channel source like Fluke 6100A or Zera VCS320. These sources can drive voltage and current in each channel at different power factors. Current and voltage are applied directly to the Unit Under Test (UUT). At the current input of the sampling wattmeter there is a multirange current transformer (CT) (10..0.1 A/0.1 A). A resistor of 10 Ω is connected to the secondary of the CT. Thus, we get a voltage of 1 V under nominal conditions. In the voltage channel a multirange (240,120,60/6 V) voltage transformer (VT) is used. For sampling the signals, two digital multimeters (DMM) Agilent 3458A in the master-slave configuration in the 1 V and 10 V range are used.

A computer controls the multimeters and the CT and VT ranges.

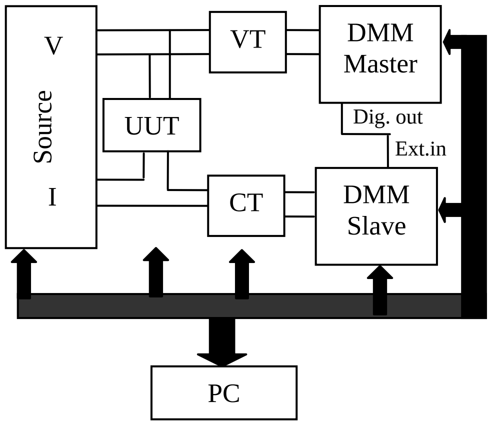

Fig. 1. Scheme of the sampling wattmeter.

Both signals are sampled by the DMM using Swerlein´s algorithm [6] and the samples are stored in a PC. To calculate harmonic power, we used an algorithm developed by Pogliano [7].

2.1 SAMPLING SYSTEM

The core of the wattmeter is the sampling system. We use asynchronous sampling because the sampling frequency is not locked to the source frequency by hardware. We use two digital multimeter (DMM) Agilent 3458A in a master-slave configuration. Therefore, each time one DMM (Master) takes a measurement, it send a pulse to the second one (Slave) to keep phase timing. This can be done using a cable connection between the “trig out” terminal of the Master DMM and the “trig in” terminal in the slave DMM.

For the measurement and processing the signal we combined two methods. To obtain the samples parameters for the current and voltage signals we use the well-know Swerlein´s algorithm [6]. The advantage of use Swelein´s algorithm is that it reduces the error in the root mean square (RMS) value by choosing the adequate sampling parameters. In our case, we are sampling two signals; voltage signal has an rms value of 6 V (corresponding to the voltage output of the VT) and the other one has an rms value of 1 V (corresponding to the voltage drop in the resistor of 10 Ω). In order to avoid phase errors, the delay between the two DMM must be measure and then corrected. In our case the time delay is aprox. 500 ns and it depends on the cable length used to synchronize the two DMMs.

Once the samples are stored, Swerlein´s algorithm calculates the RMS value of each signal. Another approach to calculate RMS values of each harmonic is the extension of Swerlein´s algorithm developed by Kyriazis [8]. In this work, the algorithm adjusts the parameters of the samples in a Fourier series in the frequency domain using least square adjustment. The samples are fitted to the following equation

$$u_{ik}^{'} = A_0 + \sum_{j=1}^{Nharm} \left(A_{jk}^{'} \cdot \sin\left(2\pi j f t_i\right)\right) + \left(B_{jk}^{'} \cdot \cos\left(2\pi j f t_i\right)\right) \quad (1)$$

where i denotes the sample taken al time t_i and k denotes de burst.

All the sampling data for each burst are store in the memory of a PC. Once the samples are store in the PC, we process them with an algorithm developed by Pogliano in [7]. In this algorithm, the signals are assumed to be described as Fourier functions of time as,

$$u(t) = U_0 + \sum_{h=1}^{M} U_{ch} \cdot \sqrt{2} \cdot \cos\left(2h\pi\, ft\right) + U_{sh} \cdot \sqrt{2} \cdot \sin\left(2h\pi\, ft\right) \quad (2)$$

$$i(t) = I_0 + \sum_{h=1}^{M} I_{ch} \cdot \sqrt{2} \cdot \cos\left(2h\pi\, ft\right) + I_{sh} \cdot \sqrt{2} \cdot \sin\left(2h\pi\, ft\right) \quad (3)$$

where $U_0, U_{ch}, U_{sh}, I_0, I_{ch}$ and I_{sh} are, respectively, the dc component and the rms value of the cosine and sine h_{th} harmonic components of both voltage and current, and f is the frequency of the fundamental.

The first step in Pogliano´s approach is determine the parameters U_0, U_{ch}, U_{sh} by least square adjustment, by minimizing the sum of the squares of the residuals

$$R_v = \sum_{n=1}^{N} \left(u\left(t_n\right) - Su_n \right)^2 \quad (4)$$

where Su_n is the N voltage samples taken at times t_n while the values $u\left(t_n\right)$ are the estimated coefficients. This is the first estimation for the parameters. Then, expression (2) is expanded in series as a function of the frequency and truncated to the linear terms,

$$u^{'}(n,f) \approx U_0 + \sum_{h=1}^{M} \left[U_{ch} \cdot \sqrt{2} \cdot \cos\left(2h\pi ft\right) + U_{sh} \cdot \sqrt{2} \cdot \sin\left(2h\pi ft\right) \right] + \quad (5)$$

$$+ \left[\sum_{h=1}^{M} 2h\pi t \left[-U_{ch} \cdot \sqrt{2} \cdot \sin\left(2h\pi ft\right) + U_{sh} \cdot \sqrt{2} \cdot \cos\left(2h\pi ft\right) \right] + \right] df$$

this new expression is used to evaluated, by means of least square adjustment, a new set of parameters U_0, U_{ch} and U_{sh}, in addition to the frequency correction df. In this case, the algorithm minimizes the sum of squares residuals

$$R_v = \sum_{n=1}^{N} \left(u`\left(t_n, df\right) - Su_n \right)^2 \quad (6)$$

The whole process is done again until df is equal to zero. The same process is done for the current samples, estimating the parameters I_0, I_{ch} and I_{sh}

To calculate power, each parameter is corrected by the correction factor due to the finite time integration of the sampling DAC [6]

$$c = \frac{\pi h f T_i}{\sin\left(\pi h f T_i\right)} \quad (7)$$

which depends on the frequency and the harmonic component. After that, the power is computed in three steps

a) Using the parameters calculated by least squares, the value of power is calculated per burst and per harmonic

b) Then, an average per burst is done. As result we obtained the value of power per harmonic

c) The sum over all the harmonics is done

2.2 SOFTWARE VALIDATION

The software validation is an important task because it links the sampling system to national standards.

In the case of Swerlein´s algorithm, we use thermal converters (TC) to validate the rms value calculated by the algorithm. Fig.2 shows a scheme of the system used, where HP3245 is a two channel source, RM.1 is a computer controlled switch, HP3458 is the DMM taking the samples, K182 is a nanovoltmeter and Tcn is a thermalconverter.

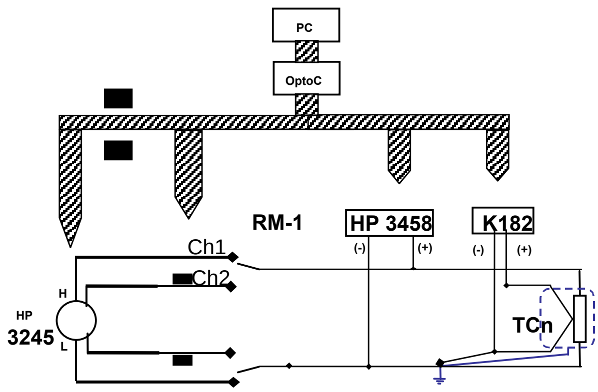

Fig. 2. Connection diagram to validate Swerlein´s algorithm

Using this procedure, the same voltage is applied to the TC and the DMM. Knowing the ac-dc difference of the TC it is possible to calculate the error in the RMS value given by Swerlein´s algorithm.

Table 1. AC-DC differences (in μV/V) measures by Swerlein´s algorithm and thermal converters at different frequencies

	11 Hz	21 Hz	33 Hz	40.4 Hz
Swerlein	-3.82	-2.33	-1.67	-1.47
Thermal converters	-5.03	-2.03	-1.53	-1.67

The validation of the measurement of phase was done using the procedure described in [9] by Stenbakken.

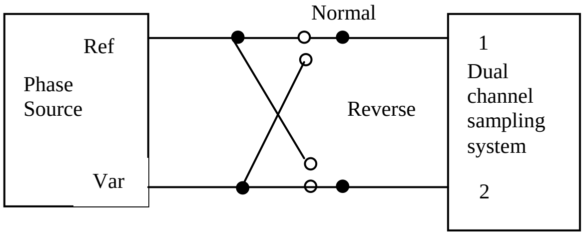

Fig. 2. Connection diagram for the phase measurement calculations

When measuring power, any different between the time delay of the two channels changes the measurement of power factor. Lets τ be the time delay of the voltage channel and $(\tau + \tau_d)$ the time delay of the current channel. Thus,

$$U_{kd} = U \cdot \sin\left(w\left(kT+\tau\right)+\alpha\right) \quad (8)$$

$$I_{kd} = I \cdot \sin\left(w\left(kT+\tau+\tau_d\right)+\alpha+\beta\right) \quad (9)$$

and the power is

$$W_d = \frac{UI}{2}\cos\left(\beta+w\tau_d\right)-\frac{UI}{2}\sum_{k=0}^{n-1}\cos\left[2wk\left(T+\tau+\frac{\tau_d}{2}\right)\right]+2\alpha+\beta \quad (10)$$

If the truncation error part of (10) is not significant, then the error due the differential time delay, E_d, can be determined expanding the cosine term. Moreover, if $w \cdot \tau_d$ is small, we can take $\cos(w \cdot \tau_d)=1$ and $\sin(w \cdot \tau_d)= w.\tau_d$. Thus, E_d can be written as

$$E_d = \frac{UI}{2} w\tau_d \sin(\beta) \quad (11)$$

It can be seen that the error depends on the phase angle. Taking four measurements with different angles β, the time delay can be calculated. Fig 2 shows the connections for the four measurements. P denote "power", "N" denotes that the switch is in Normal position, and "R" in reverse position.

Channel A of the source is used as reference, and channel B as V_A or V_B

1) PN90 the reference is connected to channel 1 and β=90º
2) PR90 the reference is connected to channel 2 and the variable to channel 1 and β=90º
3) PN270 the reference is connected to channel 1 and β=270º
4) PR270 the reference is connected to channel 2 and the variable to channel 1 and β=270º

With this procedure, the time delay between the two DMM can be calculated. Table 2 shows the values for the four measurements and the calculated delay between the DMMs using a two channel source HP3245 and the output of each channel was setting in 1 V.

Table 2. Power measurements for determining the time delay between the two DMM

PN90	PR90	PN270	PR270	τ_d (s)
-1.51e-4	1.65e-4	1.52e-4	-1.66e-4	4.75e-7

The value obtained for the time delay between the two DMM using the method described above was 475 ns. It confirms to the value obtained with the software.

2.3 VOLTAGE CHANNEL

In the voltage channel a two stage voltage transformer (VT) is used. It has 3 ranges for 240 V, 120 V and 60 V with a secondary of 6 V. We choose 6 V at the output to use the DMM in the 10 V range. To measure the error of the transformer in module and phase, a lock-in was used to compare the VT with a two stage standard voltage transformer calibrated at PTB. Fig. 3 shows the connection diagram of the set up, where PSV110 is a Lock-in, ITV-1 is the standard transformer and X is the transformer under test,

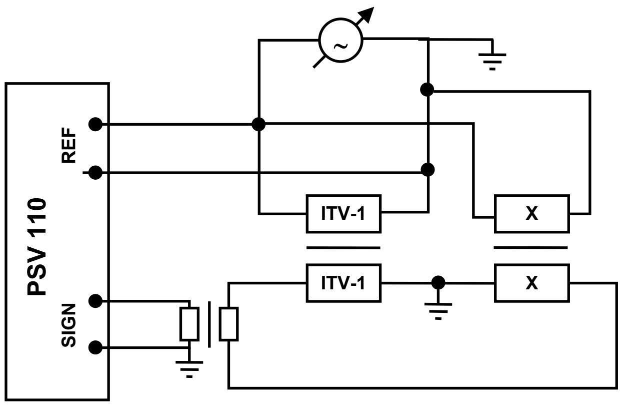

Fig. 3. Connection diagram for calibration of the voltage transformer

Measured error in the VT at nominal input voltages of 6 μV/V and 8 μrad are corrected by software. Due to the errors of the VT, we do not use a compensation transformer.

2.4 CURRENT CHANNEL

At the current input of the sampling wattmeter there is a multirange current transformer (CT) (10..0.1/0.1 A). To measure the error of the current transformer in module and phase angle, the same scheme as in the VT calibration was used. The CT was calibrated with a current transformer calibrated at PTB. Fig.4 shows the connection diagram of the calibration system, ITV-1 is the standard transformer and X is the transformer under test,

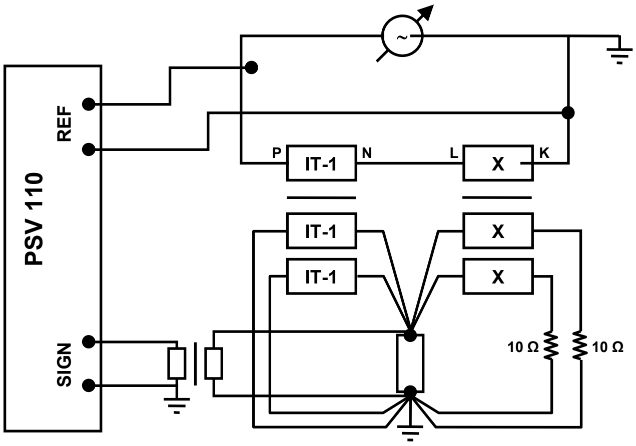

Fig. 3. Connection diagram for calibration of the current transformer

Errors measured at nominal input currents of 2 μA/A and 2 μrad were measure are corrected by software. A resistor of 10 Ω is connected to the secondary of the CT, thus we get a voltage of 1 V under nominal conditions. The resistor is built with 10 Vishay resistors of 100 Ω. We use 10 resistors instead of one to run each resistor at one quarter of its nominal power. Temperature coefficient (TCR) of the resistor is 0.5º/ºK. Despite of this low TCR, the resistor is immerse in an oil bath and the internal temperature of the bath is monitored by a PT100 thermal resistance. The value of the 10 Ω resistor is corrected by the temperature measured by the PT100.

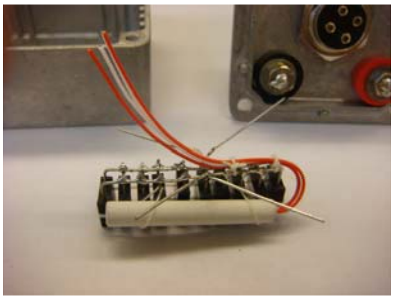

Fig. 4. Photograph of the 10 Ω resistor and PT100 resistor to monitor the oil bath temperature

Current transformer can be replacing by ac-dc current shunts. In our case, we used 1A and 5 A shunts. This shunts were calibrated with a Fluke 5700 calibrator in 1 A and 2 A respectively and the output voltage was measure with a DMM Agilent 3458.

3 MEASUREMENT RESULTS

To know the performance of the new sampling wattmeter, we compared it with the thermal power comparator at 120 V and 1 A at power factors or 1, 0.5 inductive and 0.5 capacitive. Fig 4 shows a diagram of the connections used for this purpose.

Table 3 shows the results and the uncertainties of the measurements. The components in the uncertainty budget of the sampling wattmeter are (expanded uncertainty):

- Sampling algorithm: 5 μV/V
- Standard deviation of the measurements: 10 μW/VA
- Current shunt : 20 μΩ/Ω
- Voltage transformer 5 μV/V

The total uncertainty of the sampling wattmeter is 25 μW/VA. The main component is the thermal dependence of the ac-dc shunt.

Table 3. Differences between power thermal converter and sampling wattmeter in μW/VA.

	Differences between power thermal converter and sampling wattmeter in μW/VA.
Cos φ = 1	23
Cos φ = 0.5 lag	-17
Cos φ = 0.5 lead	-23

4 FUTURE WORK

In order to reduce the uncertainty due to the thermal dependence of the shunt a new series of shunts of 5 A and 1 A are under development. Together with the shunts a resistive divider will be used to measure power harmonics. Also, new measurements at 50 Hz will be done with the current transformer.

5 CONCLUSIONS

A new sampling wattmeter was built at INTI. Comparisons between the new system and the thermal power comparator show a promising agreement within the uncertainties.

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