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# Power Meter for Highly Distorted Three-Phase Systems

Alessio Carullo and Marco Parvis

**Abstract**—This paper describes a low-cost, three-phase power meter, which is based on a fast, specially designed acquisition board coupled to a PC via the PC parallel/printer port or by means of an AT card. The power associated with the fundamental and first harmonics is computed by software that operates in the time domain and employs a sample-weighting procedure that makes the uncertainty related to the asynchronous sampling negligible. The low-cost acquisition board features two 8-bit 1 MHz converters and a local RAM, which decouples the PC clock from the measurement requirements. Hall effect transducers are used for the current channels and fast differential amplifiers for the voltage channels. The fast sampling frequency allows simple antialiasing filters to be employed. Digital filtering is used to reduce the sample number while increasing the resolution. The power uncertainty provided by this arrangement is less than 0.1% with 2.5 measurements per second when a low-cost 486DX33-based PC is used.

**Index Terms**—Digital signal processor, electric variables measurement, power measurement, signal analysis, time domain analysis.

## I. INTRODUCTION

HIGH-SPEED motors driven by static power converters are now employed in several industrial fields. In these motors, a precise characterization in terms of circuit parameters is required in order to obtain satisfactory performance and improve the dynamic behavior of such systems. This requires testing the motors at the working frequency where sinusoidal power is not easily available. The use of an inverter as a variable frequency power supply is an attractive solution, but requires special instrumentation to perform the measurements correctly.

The total rms values of voltage and current and the total active power are not enough for a complete characterization when nonsinusoidal signals are involved, so a spectral analysis should be performed. Several manufacturers have already made special instrumentation available that is able to perform such a harmonic analysis. Unfortunately these instruments are rather costly and their powerful capabilities are often not really required.

The harmonic analysis can often be limited to the first harmonic when electromechanical devices are involved and the attention is focused on typical quantities such as efficiency and power rating. This simplification is possible because in electromechanical devices the energy that is actually converted from electrical into mechanical energy is substantially related

to the electric power that is associated with the first harmonic [1]. Moreover, energy losses and magnetic stresses in magnetic materials usually depend on the first harmonic of the voltage that is used to feed the device [2], [3].

In this paper, an inexpensive and portable instrument is described. It has been conceived to determine fundamental and low-order harmonic values with a low uncertainty. The performance is obtained using a fast though simple acquisition board and software that operates in the time domain and is responsible for the low overall uncertainty.

## II. MEASURING SOFTWARE

### A. Reasons for a Time Domain Processing

The measuring software must determine the voltage, current, and power that are associated with the different harmonics with special attention being paid to voltage. Several approaches can be followed to perform this task either operating in the frequency domain, via fast Fourier transform, or in the time domain. The frequency domain solutions are extensively employed when the complete spectrum must be analyzed, since their computational complexity is usually lower than other solutions [4]. Unfortunately the accuracy they provide on the actual amplitude of a single spectral line can be poor if special precautions are not employed. Synchronous sampling of the input signal is usually required [5]–[7] and a phase locked loop circuit is needed. This increases the instrument cost and leads to tricky problems in the presence of highly distorted signals. Commercial wattmeters based on such an approach often fail to give meaningful results when applied to pulse width modulation inverters due to an incorrect fundamental frequency determination.

Time domain procedures can be considered when only few spectral lines must be determined. This is the case of the instrument described that was mainly conceived to extract the fundamental frequency and first harmonic components. In this case, the process can be designed to give a remarkable accuracy along with a low computational complexity.

### B. Fundamental Frequency Determination

The fundamental frequency value is obtained through a minimization process that tries fitting a four parameter model to the signal samples  $S_i$

$$S_i = S_0 + A_c \cos(ki) + A_s \sin(ki) \quad (1)$$

$$k = \frac{2\pi F_f}{F_s} \quad (2)$$

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where  $F_f$  is the fundamental frequency and  $F_s$  is the sampling frequency.

The minimization problem is the well-known four parameter sine fitting [8] that is linear in three (offset  $S_0$ , cosine amplitude  $A_c$ , and sine amplitude  $A_s$ ) of the four parameters. In this paper, a least-square estimation that employs a time weighting window is used. Such a weighting window is required to reduce the effect of the nonsynchronous sampling which otherwise would not allow a meaningful estimation to be obtained. A description of the performance increase that is obtained by employing such a window is described in the following sections.

The three linear parameters are estimated for a defined fundamental frequency value  $F_f$  as

$$[S_0, A_c, A_s]^t = (\mathbf{U}^t \mathbf{U})^{-1} \mathbf{U}^t \mathbf{S}_w \quad (3)$$

where

$$\mathbf{U} = \begin{bmatrix} w_1 & w_1 \sin(k) & w_1 \cos(k) \\ w_2 & w_2 \sin(2k) & w_2 \cos(2k) \\ \vdots & \vdots & \vdots \\ w_N & w_N \sin(Nk) & w_N \cos(Nk) \end{bmatrix} \quad (4)$$

$$\mathbf{S}_w = [w_1 S_1, w_2 S_2, \dots, w_N S_N]^t \quad (5)$$

where  $[w_1, w_2, \dots, w_N]$  are weighting coefficients that implement a Hanning window on the estimation interval and  $N$  is the number of the (decimated) samples in the observation interval.

Once the parameters have been obtained, the squared difference  $R$  between the original samples and the equivalent samples obtained by applying the model is computed

$$R = [\mathbf{U} \cdot [S_0, A_c, A_s] - \mathbf{S}_w] \cdot [\mathbf{U} [S_0, A_c, A_s] - \mathbf{S}_w]^t. \quad (6)$$

The estimation is repeated by adjusting the fundamental frequency  $F_f$  until a minimum of  $R$  is found. The process usually requires only a few iterations to converge, provided a reasonable initial frequency is available.

Such a preliminary estimation is obtained by using a simple threshold detector and is performed only the first time the instrument detects a valid signal or when the subsequent minimization process ends with an unsatisfactory convergence factor.

### C. Harmonic Analysis

The samples corresponding to an integral multiple of the signal period are used to determine the power associated with the different harmonics. First, the average signal value  $S_0$  is removed from the signals, then, a two parameter estimation is performed on the voltage and the current. The estimation is the simplification of that used for the frequency and leads to the cosine and sine amplitudes

$$[A_c, A_s]^t = (\mathbf{U}'^t \mathbf{U}')^{-1} \mathbf{U}'^t \mathbf{S}_w \quad (7)$$

where

$$\mathbf{U}' = \begin{bmatrix} w_1 \sin(k) & w_1 \cos(k) \\ w_2 \sin(2k) & w_2 \cos(2k) \\ \vdots & \vdots \\ w_N \sin(Nk) & w_N \cos(Nk) \end{bmatrix} \quad (8)$$

and  $\mathbf{S}_w$  and  $w_i$  are the same as in the previous section.

The power  $P_1$  associated with the first harmonic is readily available as

$$P_1 = \frac{V_c I_c + V_s I_s}{2} \quad (9)$$

where  $V_c, V_s, I_c, I_s$  are the cosine and sine amplitudes of the voltage and current.

The sample values  $\mathbf{S}_1$  corresponding to the first harmonic are then computed as

$$\mathbf{S}_1^t = \mathbf{U}' \cdot [A_c, A_s] \quad (10)$$

and subtracted from the original signal.

The process is repeated for the subsequent harmonics the amplitude of which must be computed by changing the matrix  $\mathbf{U}'$  each time.

One should note that the computation of  $(\mathbf{U}'^t \mathbf{U}')^{-1} \mathbf{U}'^t$  does not actually require three matrix products and a matrix inversion due to the matrix symmetries that greatly simplifies the processing.

### D. Choice of Sampling Rate

The harmonic components that the wattmeter has to determine are at frequencies which barely reach some kilohertz. With these maximum frequencies, a sampling rate of around 10 kHz should be sufficient to satisfy the Nyquist limit. Unfortunately, such a choice would require complex analog antialiasing filters that are difficult to obtain with matched and well-known amplitude and phase characteristics. The authors therefore decided to use a much higher sampling frequency working with converters capable of running at 1 MHz.

At these frequencies, the energy content of both current and voltage signals is rather low and therefore a simple analog low-pass filter is sufficient to avoid aliasing errors. Higher sampling frequencies are not desirable since the number of samples required to perform the computation would become too high.

A further advantage of employing such high sampling frequencies is that converters with a reduced number of bits can be employed without affecting the final accuracy, since the quantization effect is greatly reduced by processes that operate on a large number of samples.

Once the samples are acquired, digital filtering can be employed, along with a decimation process to reduce the sampling frequency to values that allow the harmonic analysis, without flooding the algorithms with a large stream of samples.

In the wattmeter described, a cascade of two finite impulse response (FIR) low-pass filters is employed that reduce the pass-band to 1 kHz. The first FIR filter is used to allow a decimation down to 50 kHz. A rectangular window, composed of five or ten taps of unitary amplitude, is used depending on the measurement configuration (three or four wire operations). The unitary tap amplitude, though nonoptimal, is used to allow the processor to perform simple fast additions with integer mathematics, greatly speeding up the processing.

The second FIR filter is designed using an optimal window approach [9] with a low-pass frequency of 1 kHz and 101 taps and is applied to allow a decimation of the samples to a frequency of 2.5 kHz to be obtained. The combined effect

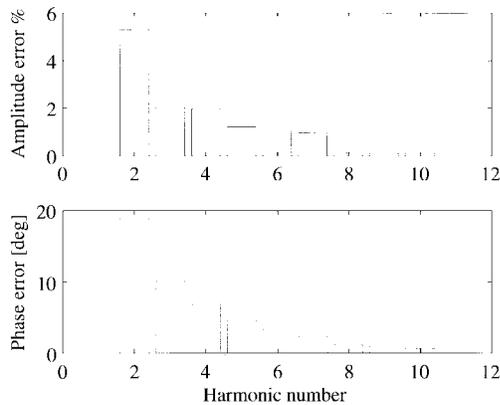


Fig. 1. Fundamental amplitude and phase uncertainties in the absence of the weighting window.

of the two filters on the amplitude of the first harmonics is limited to a fraction of a percent for frequency in the range of (20 to 100) Hz and is removed after the power is computed and the frequency is known.

#### E. Algorithm Performance

The algorithm performance, i.e., its ability to determine the actual value of one harmonic component in the presence of other harmonics and noise, cannot easily be predicted in closed form. This is due both to the iterative nature of the process and the strong nonlinearity involved in the frequency determination.

The presence of harmonic components that are different from the one whose amplitude and phase must be determined affects the estimation due to the nonsynchronous sampling which is employed. The effect depends on the over- or under-extension of the observation interval with respect to the fundamental period, on the harmonic amplitudes and reciprocal phases, and the ratio between the sampling frequency and the harmonic frequency.

The uncertainty induced by the harmonic presence can be important in the absence of the weighting function that has been employed in the processing. Figs. 1 and 2 show the maximum error that is induced on the fundamental amplitude and phase estimation by the harmonic presence whether the weighting window is used or not. The plots have been obtained by operating at a sampling rate of 2.5 kHz and estimating the amplitude and phase of a 50 Hz signal in the presence of one harmonic at a time that has the same amplitude as the fundamental. In the absence of the weighting window, the error makes the estimation meaningless. If the weighting window is employed, the uncertainty of harmonics above the second is less than  $5 \times 10^{-6}$  for the amplitude and  $5 \times 10^{-4}$  rad ( $0.03^\circ$ ) for the phase.

The noise effect on the estimated parameters can be previewed by separately analyzing the two processes the samples are subjected to. A rough but useful approximation can be derived with respect to this case.

The first process is the decimation via the finite impulse response filter. The standard deviation affecting the decimated samples depends on the actual filter that is employed, but,

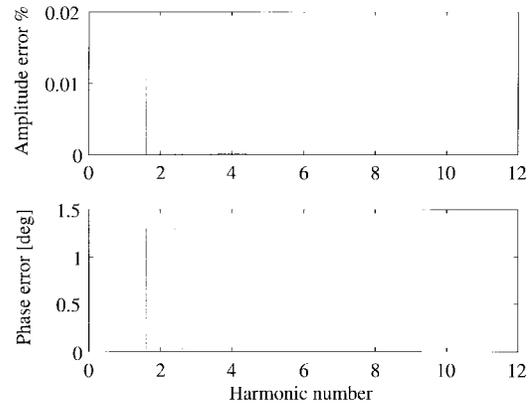


Fig. 2. Fundamental amplitude and phase uncertainties in the presence of the weighting window.

provided that a reasonable number of taps are used, a rough estimation can be computed on the filter pass-band basis. If the standard deviation of the noise affecting the undecimated samples does not depend on the sample amplitudes and is uncorrelated with respect to the signal, i.e., if the noise is additive and conventional techniques are used to get rid of the quantization deterministic phenomena, the filter effect is

$$\sigma_{S_{dec}} \approx \sigma_S \sqrt{B_r} \quad (11)$$

where  $\sigma_S$  and  $\sigma_{S_{dec}}$  are the standard deviations of undecimated and decimated samples and  $B_r$  is the pass-band of the filter normalized to the Nyquist frequency.

The standard deviation of cosine  $A_c$  and sine  $A_s$  parameters can be computed if one observes that (7) can be rewritten as

$$\begin{aligned} A_c &= \sum_i c_{ci} S_{dec i} \\ A_s &= \sum_i c_{si} S_{dec i} \end{aligned} \quad (12)$$

where  $c_{ci}$  and  $c_{si}$  are the coefficients obtained by resolving (7) and  $S_{dec i}$  are the decimated samples. The cosine and sine standard deviation can therefore be computed as

$$\sigma_{A_c} = \sqrt{\sum_N c_{ci}^2 \sigma_{S_{dec}}^2} = \sigma_{S_{dec}} \sqrt{\sum_N c_{ci}^2} \approx \sigma_{S_{dec}} \left(\frac{2}{N}\right)^{1/2} \quad (13)$$

where the factor  $\sqrt{2/N}$  that appears in the formula is a consequence of the way the coefficients  $c_{ci}$  are produced. The same standard deviation holds for  $A_s$  parameters too, as they are produced using the same processing thus

$$\sigma_{V_c} = \sigma_{V_s} = \sigma_V; \sigma_{I_c} = \sigma_{I_s} = \sigma_I. \quad (14)$$

Once the standard deviation of the sine and cosine parameters have been derived, the power uncertainty is previewed by applying the conventional uncertainty propagation rules [10] to (9)

$$\sigma_P^2 = \frac{I^2 \sigma_V^2 + V^2 \sigma_I^2}{4} \quad (15)$$

where the harmonic subscript is omitted for clarity;  $V = \sqrt{V_c^2 + V_s^2}$  and  $I = \sqrt{I_c^2 + I_s^2}$  are the modules of the

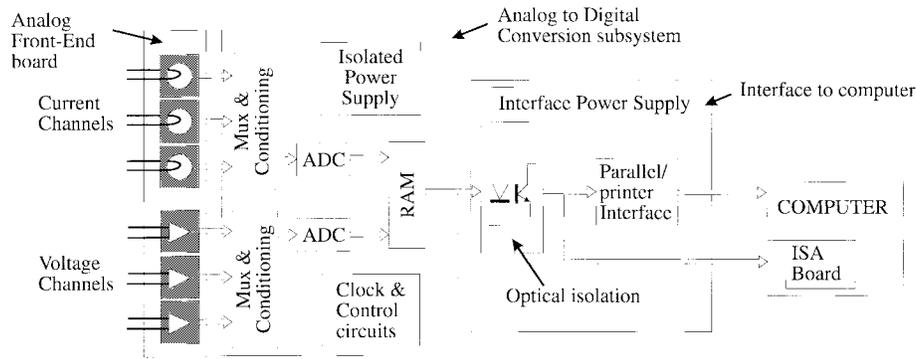


Fig. 3. Block diagram of the proposed wattmeter.

measured voltage and current and the effect of the residual parameter covariance is neglected.

Equation (15) can be usefully rearranged to show how the relative power standard deviation depends on the measuring conditions. By introducing the reduced standard deviations of voltage and current, i.e., the standard deviation  $\sigma_R$  referred to the instrument full scales  $V_f$  and  $I_f$ , one can obtain

$$\sigma_{P_R}^2 = \frac{\sigma_P^2}{P^2} = \frac{1}{\cos^2 \varphi} \left( \frac{V_f^2}{V^2} \sigma_{V_R}^2 + \frac{I_f^2}{I^2} \sigma_{I_R}^2 \right). \quad (16)$$

Equation (15) shows that, as expected, the absolute power uncertainty depends on the measured values of the current and voltage but not on their phase difference and thus the relative uncertainty (16) increases as the power factor tends to zero.

Equation (15) also allows one to preview the minimum expected uncertainty once the hardware choices have been made. With the choices described in the following section (8-bit dithered converters, 250 kHz sampling rate, and estimation over four periods), the uncertainty in the most favorable conditions (full-scale signals on both voltage and current and unitary power factor) is less than 0.01% plus the uncertainty corresponding to the input stage that is therefore expected to be the most significant uncertainty contribution.

### III. INSTRUMENT SCHEME

#### A. Hardware

The acquisition subsystem is subdivided into three sections as shown in Fig. 3: the analog front-end, the analog to digital conversion subsystem, and the computer interface.

The analog front-end is composed of three Hall effect sensors for the current channels and three differential amplifiers for the voltage channels. It is assembled on a board that is separated from the other sections for shielding and safety reasons, since the inputs of the voltage channels can be connected to dangerous voltages. The power supply required by this section is obtained from an isolated power supply that is also used for the conversion system.

The analog-to-digital conversion subsystem is composed of two 8-bit, 1-MHz sampling rate converters, two multiplexers,

which are used to scan the input channels and embed the dithering circuit, and a buffer memory. By employing such converters, a maximum sampling frequency of 500 kHz is obtained, when the Aron circuit is used on three wire power systems, and 250 kHz when a measurement is made on four wire systems.

The buffer memory has a 256 kWord capability and is required to temporary store the samples that cannot be transferred to the computer in real time. This section also contains the circuits that manage the multiplexer, ADC timing, and memory addressing.

The interface to computer allows two connection methods to be employed. The simpler but slower solution is to use the PC's parallel printer port. In this solution, no additional board is required inside the computer, which is convenient when using a notebook or subnotebook PC. The data are transferred four bits at a time to allow backward compatibility with older computers that do not employ enhanced parallel ports. The other solution is based on an additional board that fits into the ISA bus of the computer. Such a solution allows a faster transfer rate to be obtained but requires a computer with an available ISA slot.

The passage between the two solutions is automatic so that the acquisition unit can be easily used in the field, with a notebook computer and a slower performance, or in the laboratory, with a higher speed. The metrological accuracy is obviously the same in the two cases.

The interface to the computer is galvanically decoupled from the acquisition subsystem by means of a barrier of high-speed optocouplers. This precaution is required since the reference potential of the front-end can be tied to any voltage in the measured system. The interface is powered by a supply system that is isolated from the previous one.

#### B. Software

The software that manages the acquisition allows the selection of the different acquisition parameters such as the sampling rate and the type of measuring configuration. The virtual panel is written in VisualBasic since it does not requires computation-intensive operations. The measuring software that has been described in the previous section is partially written in C language for performance reasons. The C routines are integrated into the VisualBasic program via a dynamic link

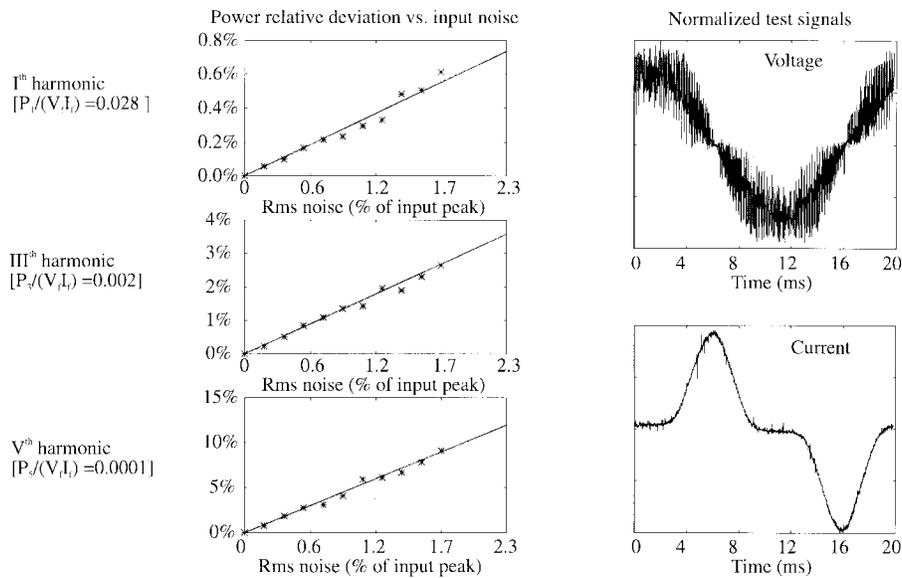


Fig. 4. Performance of the algorithm in the presence of distorted signals and noise on the samples.

library. The processing required for acquisition, frequency and first harmonic value estimations is completed within 400 ms on a 486DX33 PC, allowing 2.5 acquisitions/s.

#### IV. EXPERIMENTAL RESULTS

A complete characterization of the wattmeter is a very difficult task due to the variety of signals and situations that can be encountered [11].

Initially, the wattmeter output was compared with the output of a commercial instrument and differences of the order of 1% were obtained. Such a difference is far larger than the standard deviation obtained by the proposed wattmeter during the tests (0.1%), but the value is compatible with the stated accuracy of the commercial instrument so that no additional information can be inferred from this test.

The authors therefore employed an indirect characterization: the measurement hardware, i.e., conditioning devices and analog-to-digital subsystem, was characterized using conventional static and dynamic techniques [12], [13] obtaining scale factor uncertainties of 0.02% on the current and 0.05% on the voltage. The algorithm behavior, in the presence of the distorted signals generated by the PWM, was analyzed using a two step procedure. First a stream of samples of both the voltage and the current were acquired in different conditions by using a 4 kW PWM inverter and different loads. Resistive and highly inductive loads were employed both linear and heavily nonlinear.

The acquired data streams were analyzed in the frequency domain to obtain an estimation of their spectra. A set of signals was then generated that matches the measured spectra. Such signals were sent to the processing algorithm corrupted by uniformly distributed noise of different amplitudes.

An example of the power uncertainty consequent to the different noise levels is shown in Fig. 4 along with the previewed values obtained from (16). The example refers to

the signals that are shown in the figure and to a processing performed on four periods. The signals were obtained on a saturated power transformer and represent a rather unfavorable measuring condition due to both the high-current distortion and to the low-power factor of the load.

The traces are in good agreement and the negligible uncertainty corresponding to the no-noise condition, confirms that the harmonic effect is made negligible by the employed weighting window.

#### V. CONCLUSIONS

A low-cost wattmeter that is able to perform an harmonic analysis has been described. At a cost of approximately \$150, the wattmeter determines the first harmonic power with a typical uncertainty of 0.1%.

The wattmeter employs a time domain processing of the samples that uses a weighting window to make the uncertainty due to the nonsynchronous sampling negligible. By using such a technique, the algorithm-related uncertainty is reduced to less than 0.01% without employing smoothing techniques that would slow down the system response.

The wattmeter is made up of an acquisition subsystem coupled to a PC via either a specially designed board or the parallel/printer port. Such a solution allows a compact and simple system to be arranged by using portable computers. The fundamental values of power voltage and current are computed in less than 400 ms even though a low-cost 486DX33-based PC is used.

The metrological characterization has been performed indirectly since the uncertainty of the wattmeters which were available to the authors did not allow them to be used as standards. The authors are now working on a test set for the characterization of wattmeters in distorted environments that should be able to allow a simple and direct characterization.

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