Easing the problem of traceability for complex waveforms by optimizing instrument design

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1 Abstract

Instruments providing previously unavailable facilities often pose challenges for those responsible for providing measurement traceability. The main challenge in the case discussed in this paper was to measure the phase of up to eight channels of complex voltage and current waveforms with five to ten times better accuracy than available from commercially available systems. The resulting solution had to be cost effective for third party calibration facilities and those who maintain their own traceability. As is often the case, a simple solution turned out to be extremely effective but the implementation would not have been possible without early collaboration between metrology and instrument design functions. The paper describes how novel instrument system design allowed phase and amplitude of various harmonically distorted and amplitude-modulated waveforms to be determined with a single digitising DMM and a few voltage dividers and shunts. Other potential measurement system architectures are reviewed and system traceability issues are discussed.

2 Introduction

Plans for the design of a new instrument cannot be said to be complete without consideration for production processes, including calibration and measurement traceability. When the parameters to be measured are new to an organization, or a significant improvement in accuracy is needed, the importance of this activity becomes relatively higher. The case discussed in this paper is of the development of commercially available instruments to provide power line frequency voltage, current and phantom power with user controllable 'distortions'. Each instrument would provide:

- Voltage output channel 1V to 1000V,
- Current output channel 10mA to 20A
- Sinusoidal waveforms 16Hz to 450Hz
- Sinusoidal and non-sinusoidal 'phantom' power with accuracy of the order of 200ppm of output.
- Harmonics up to 100th, maximum frequency 6kHz
- Other phenomenon including Flicker (amplitude modulation) of sinusoidal or non-sinusoidal waveforms, Fluctuating Harmonics, Interharmonics (harmonics of non-integer order) and Dips and Swells in the outputs.

The new instruments would provide up to four ‘phases’ of voltage/current channels, eight channels in all. Figure 1 depicts the outputs from a three-phase system. All voltage and current channels were required to be completely independent except for phase relationships, but controlled from a single user interface. The requirement specification of the new instrument included amplitude accuracy targets well within the measurement capabilities of Fluke given that sourcing accurate low frequency sinusoidal voltage and current has been a corner stone of Fluke’s calibration business for decades. More of a challenge was the need to measure non-sinusoidal, amplitude-modulated waveforms and phase measurement with accuracy previously unavailable outside turnkey solutions in highly capable laboratories. Cost of calibration equipment and traceability was also an important issue as service facilities must be provided in several locations around the world. In fact, the viability of the whole design project hung to a large extent on finding a cost effective calibration solution for phase.

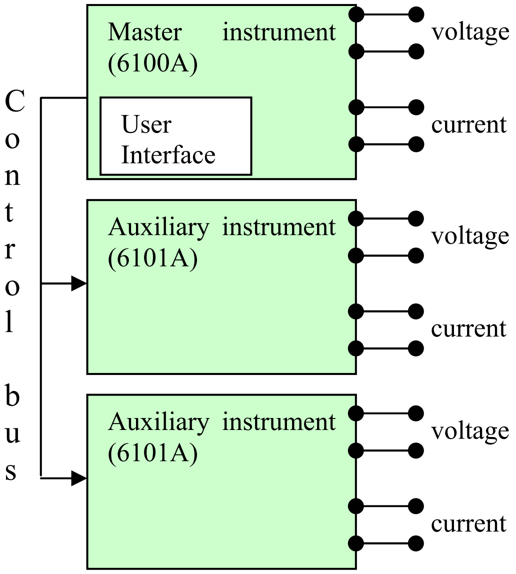

Figure 1. A three phase system

3 Phase accuracy requirement

The general definition of electrical power is given by: $P = \frac{1}{T}\int_T u.idt$

where P is power, u is the instantaneous voltage and i the instantaneous current.

Very accurate measurement of power can be achieved using this relationship from sampled voltage and current. Unfortunately there are potential ambiguities when resolving phase angle from voltage, current and power and this is exacerbated when outputs are non-sinusoidal. Consequently, to generate power accurately, accurate phase control of all components in waveforms is essential. Given Fluke’s expertise in generation of accurate voltage and current, the phase accuracy requirement dominates the power accuracy attainable. The contribution of phase accuracy to power accuracy $u(P)$ is given by:

$$u(P) = (1 - \frac{\cos(\Phi \;\; + u(\phi)}{\cos(\Phi)}) \times 10^6 \, ppm, \qquad (1)$$

where Φ is the set phase angle and $u(\phi)$ is the phase accuracy.

The target for the 6100A specification was that phase accuracy should contribute less than 100 ppm to power accuracy at power line frequencies and Power Factor = 0.5. From (1), phase accuracy needed to be less than:

$$u(\phi) = \arccos\left(\left(1 - 100^{-6}\right) \times \cos(60)\right) - 60, \;\; i.e., \leq 0.0033°$$

Therefore, for a Test Accuracy Ratio (TAR) of 4:1, the current to voltage phase calibration uncertainty needed to be better than 0.000825°. Similarly, calibration uncertainty for a harmonic at 6kHz needed to be better than 0.1°.

Many instrument design/calibration regime options were considered but the eventual solution turned out to be one of the simplest.

4 Signal generation

A brief description of the signal generation system will aid discussion of the calibration system. To provide a common Phase Reference point for all voltage and current output channels a 'Phase Reference' signal is provided by the Master instrument to all channels including those in the Auxiliary instruments. Each voltage or current output waveform is generated using a form of DDS (Direct Digital Synthesis). A digital representation of the desired waveform is stored in memory. A master clock steps through the memory locations and reconstructs the waveform via digital to analog (D to A) conversion and amplification. The (D to A) conversion process and subsequent amplification introduces phase shifts such that without correction, the output at the binding posts lags the digitally generated waveform. Figure 2 is a stylized representation of the relationship between the 'Phase Reference', the digitally sampled waveform and the analog output signal.

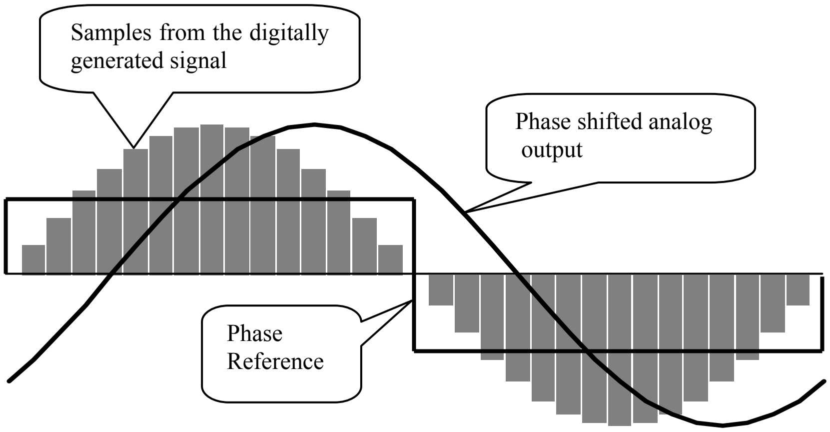

Figure 2. Before phase adjustment

Unlike most DDS systems, the 6100A uses a digital feedback system that continuously analyses the output waveform and modifies the values stored in the digital memory to remove amplitude and phase errors. Nevertheless, the analog to digital (A to D) process in the feedback path introduces phase errors that cannot be controlled by the feedback. These errors must be minimized by calibration adjustment. Figure 3 shows the relationship after phase adjustment has shifted the digitally sampled waveform to align the analog output to the Phase Reference. In practice there will be a small residual phase error determined by the accuracy of the measurement.

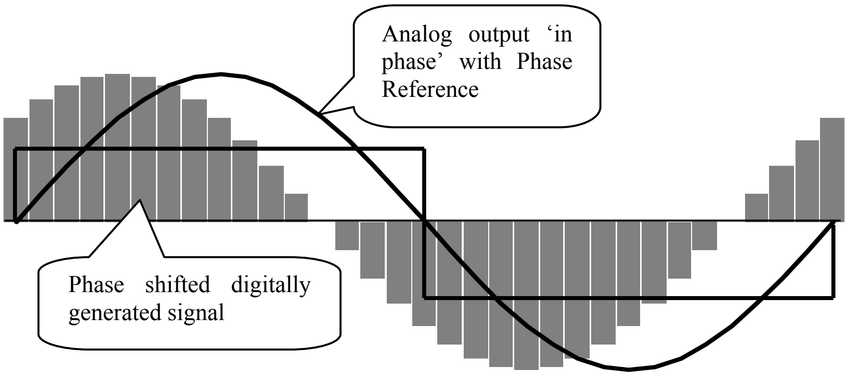

Figure 3. After phase adjustment

Because the phase of all signals is derived from the same, common Phase Reference, the calibration of an Auxiliary instrument can be completely independent of the Master instrument controlling it.

5 Phase calibration methods

There are many ways that amplitude could be measured, the big challenge was to measure the phase angle between two 'channels' i.e., current of phase one to voltage of phase one or voltage of phase two to voltage of phase one etc..

5.1 Two phase measurement techniques that were rejected are:

5.1.1 Zero crossing phase meter

- Phase is computed from a very small portion of the signal and accuracy is degraded by noise on the signals at the zero crossing.
- Cannot provide phase information on the individual components of a harmonically distorted signal.
- The zero crossing point of a complex signal varies with even harmonic content.
- Accuracy required not available commercially.

5.1.2 Simultaneous sampling of two output channels

- Requires either a two 'channel' device or two identical channels. Some devices compute sinusoidal power accurately but phase angle is a by-product of that process and as a consequence has lower accuracy.
- A turnkey solution using two sampling digital multimeters (DMM) considered was not viable (some phase errors would add).
- Accuracy required not available commercially

5.2 The method adopted

A DDS control system with some fairly minor design modifications provided the method to measure phase angles between channels. As each output channel is locked to a common Phase Reference, comparing all signals against that reference effectively provides the relative phase angle of every channel. It is immediately obvious there are two uncertainty contributions to the current to voltage phase relationship (or voltage to voltage relationship in a multiphase system) but as will be shown, these uncertainties are small and other relatively large uncertainties effectively cancel. Essentially, the phase measurement system uses the 6100A's own DDS sampling pulses to provide trigger pulses to a sampling DMM as depicted in Figure 4.

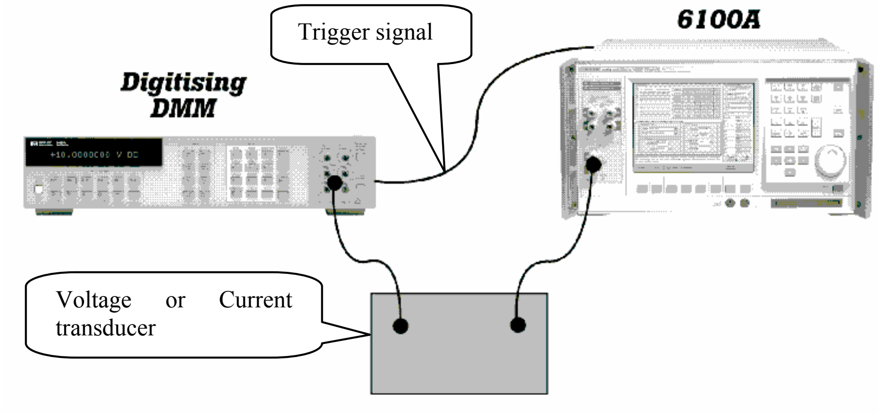

Figure 4. Phase calibration set up

6 The calibration system

Figure 5 shows how the instrument provides a sample reference which is used to determine the difference between the Phase Reference signal and the analog outputs. In theory, as all analog signals in a system are referenced to the same signal, the actual error does not matter as long as it is constant. The requirement for independance of Master and Auxiliary instruments and for calibration facilities world-wide means in practice however that phase errors must be eliminated or corrected.

With the exception of connecting transducers, the calibration system is completely automated. The sample reference signal is turned on and off via the GPIB. When turned on, the trigger signal is gated internally by the 6100A master instrument such that it remains high until the positive going edge of the master phase signal, thus synchronizing the sample taken by the calibration system DMM with the master phase signal. It is this synchronization that provides the phase information.

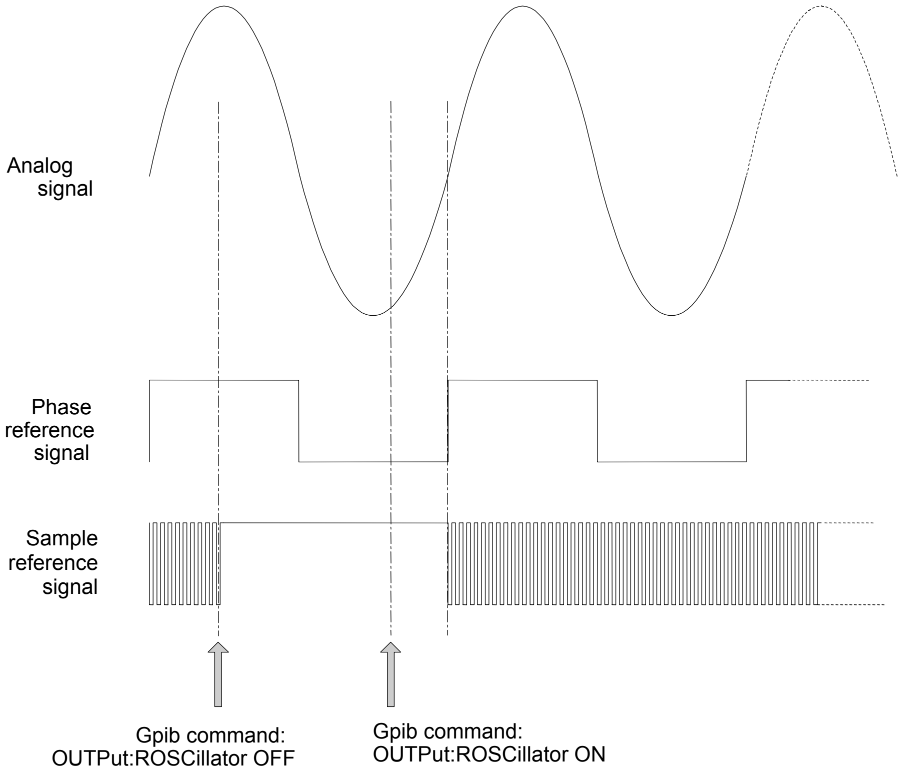

Figure 5. Phase measurement

The system DMM is programmed to take the required number of samples, so as long as the minimum required time is allowed to elapse, the timing of the sample reference ‘off’ command is not critical. The sample reference frequency is always a binary multiple of the analog fundamental frequency thereby simplifying the task of analysing the sampled data.

6.1 Fourier analysis of the sample

As the sample taken by the DMM is always a binary multiple of the analog signal fundamental frequency simple FFT analysis can be employed. Instrument design considerations required that the range of fundamental frequencies be split into bands with varying sample rates. The harmonic content of waveforms from the 6100A and 6101A instrument is limited by the smaller of 6kHz or 100th harmonic. The number of samples per cycle of fundamental frequency varies between 2048 samples at 16Hz and 128 samples at 450Hz ensuring the highest harmonic frequency is always well below the Nyquist frequency.

6.2 Transducers

Transducers are used to convert the different 6100A output voltage and current levels to nominally 800 mV. The DMM is always used on its 1.2 volt DC range for all measurements to reduce the relative phase uncertainty contribution from the DMM.

6.2.1 Coaxial Current Shunts

Current shunts are based on a design from the Mendelejev Institute, St Petersburg. The shunt design is described by Svensson [1]. The shunts were built and measured measured by SP, the Swedish National Testing and Research Institute to have a mutual inductance of 0.5 nH ± 0.5 nH. The resistors used have low temperature coefficients, less than 1 ppm/degree.

6.2.2 Voltage Dividers

Resistive voltage dividers were constructed using the same type of resistors as are used in the shunts. The voltage dividers are made up from a large number of resistors to minimise the power dissipated in each resistor. To minimise the effects of stray capacitance and compensate for the input capacitance of the DMM (approximately 260 pF) a parallel capacitive divider is used. The voltage dividers were also built and measured by SP.

7 Error Analysis

The discussion of errors and uncertainties will concentrate mainly on the use of a single DMM for phase measurement. However, brief outline of transducer performance is required to place the influence of other errors and uncertainties in context. The coaxial current shunts contribute the largest proportion of transducer phase uncertainty. They typically have phase displacement error uncertainty of around 0.0003° at 60Hz and 0.013° at 1500Hz. Voltage divider uncertainty is typically 0.0002° at 60Hz, 0.002° at 1500Hz.

7.1 DMM Phase Angle Errors

7.1.1 Bandwidth

Because the analog input of the DMM acts like a single pole low pass filter it contributes a phase shift. For a single pole low pass filter the phase error (change) can be estimated from equation (2) and a correction applied.

$$E_{BW} = \arctan\left(\frac{f}{f_{BW}}\right) \tag{2}$$

Where f is the frequency to be measured and f_{BW} is the upper bandwidth of the DMM.

The effect of the DMM bandwidth is to make the analog waveform appear to be lagging in phase with respect to the sample reference signal.

7.1.2 Sample Time Aperture and Trigger Delay

The DMM Sample Time Aperture and Trigger Delay also affect the measurement. The falling edge of the Sample Reference Output triggers the DMM to take a sample. There is then an internal delay (Trigger Delay) within the DMM before the signal is integrated over a period of 1.4 µs (the Sample Time Aperture). The DMM stores the result and waits for the next trigger. Figure 6 shows the relationship between the Sample Trigger signal and the DMM samples. For simplicity the analog signal is shown with zero phase error, crossing zero at $\mathbf{t_0}$. Note that the value stored by the DMM for a sample is the mean of the analog signal during the sample period. This mean value to all intent and purpose occurs in the center of the sample, at $\mathbf{t_S}$. It can be seen that the DMM will store the value of the analog signal at time $\mathbf{t_D} + \mathbf{t_A}$ after the trigger event. This has the effect of making the waveform appear to be leading in phase, i.e., in the opposite direction to the effect of bandwidth.

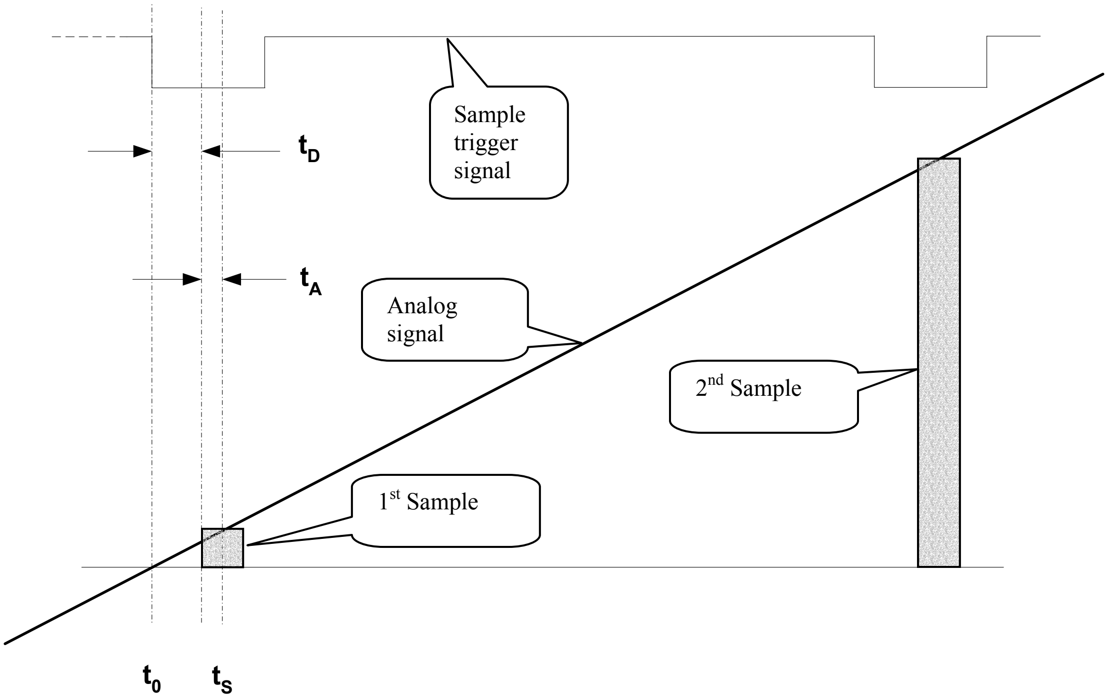

Figure 6. Effects of trigger delay and sample aperture

The actual phase errors due to trigger delay and sample aperture may be calculated using equations (3) and (4).

$$E_{Trigger} = t_D \times f \times 360 \text{ degrees} \tag{3}$$

$$E_{Aperture} = t_A \times f \times 360 \text{ degrees} \tag{4}$$

Where *f* is the frequency to be measured, and trigger delay and sample aperture are in seconds. Note that t_A is half the DMM aperture setting.

7.1.3 Combination of DMM phase errors

The errors described at (2), (3) and (4) above are combined to derive a frequency dependant value to be used to correct for phase measurement errors:

$$E_{Total} = E_{BW} - E_{Trigger} - E_{Aperture}$$

- Measurements gave DMM bandwidth between 130 kHz and 140 kHz. The typical value 135kHz ± 5kHz is used below.

- The DMM specified maximum trigger delay is 175 ns with a maximum variation between instruments of 125 ns. The worse case is a trigger delay of 113ns ± 63ns.

- The sample time aperture used is 1.4µs and the DMM timebase specification is ±0.01%, This gives: $t_A = 700 \pm 0.07 \;\; ns$

Using these values, typical errors at 60 Hz and 6 kHz are shown in Table 1, and the uncertainty of those errors in Table 2.

Frequency	E_{BW}	$E_{Trigger}$	$E_{Aperture}$	E_{Total}
60 Hz	0.0255	0.0024	0.0151	0.0079
6 kHz	2.545	0.2441	1.5120	0.7887

Table 1. Phase errors in degrees due to the DMM

Frequency	Uncertainty in E_{BW}	Uncertainty in $E_{Trigger}$	Uncertainty in $E_{Aperture}$	Combined uncertainty	Expanded Uncertainty (k = 2)
60 Hz	0.0004	0.0008	0.0000	0.0009	0.0018
6 kHz	0.0441	0.0786	0.0001	0.0901	0.1802

Table 2. DMM phase error uncertainty (degrees).

As previously discussed, the target uncertainty for the phase calibration system was <0.000825° at power line frequencies. The DMM contribution to phase uncertainty is alone more than twice that. However, using the same DMM to measure both voltage and current against a common reference signal means that all DMM related uncertainties other than short-term stability and measurement noise cancel. These remaining contributions are estimated to be 0.00023°, which combined with transducer contributions gives a total expanded system uncertainty of 0.00049° (k = 2) for current to voltage phase. The target phase measurement uncertainty was thus easily met.

The phase relationship between the voltages of two different instruments can also be calibrated with this system but as the instruments may be calibrated at different service centers, all DMM phase uncertainties must be included. The resulting expanded system uncertainty for the voltage to voltage phase error between a 6100A Master and any 6101A Slave instrument is 0.0019°.

8 Conclusion

There were two principle objectives for the development of the calibration system for the 6100A instrument. The first of these was to develop a calibration system with uncertainties for phase several times better than is available from commercially available equipment. The second objective was that the calibration method was cost effective for third party calibration facilities and those who maintain their own traceability. Close collaboration between metrology and instrument design functions from the very beginning of the development cycle enabled both objectives to be met.

As is often the case, the solution turned out to be simple but extremely effective. Instrument design was optimized to provide built in facilities which provide simple, easily maintainable traceability for all key parameters; particularly phase accuracy. Any competent laboratory with traceability for amplitude and phase displacement on suitable current shunts and voltage dividers can traceably calibrate the Fluke 6100A Electrical Power Standard for sinusoidal and nonsinusoidal power.

9 Reference

1. S. Svensson, *Power Measurement Techniques for Nonsinusoidal Conditions*, Chalmers 1999, pp.86-89.