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## TRACEABLE CALIBRATION OF ELECTRICAL COMPONENTS REGARDING AMPLITUDE AND PHASE DISPLACEMENT IN THE FREQUENCY RANGE OF 0.15 HZ TO 12.5 (50) KHZ

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**Abstract** – It is mandatory that for every calibration system intended for the determination of the complex transfer coefficients of vibration sensors, the phase-frequency response of its electronic subunits is determined and the total system is calibrated in its entirety. Since currently no national metrological institution is equipped with the necessary measurement facilities, a research project was carried out jointly by the SPEKTRA GmbH Dresden, Germany, and PTB, the German metrological institute, in the context of which such instrumentation was developed, constructed, and linked to the German national standard.

**Keywords:** phase measurement, voltage measurement, traceability

### 1. INTRODUCTION

The traceable calibration of the electrical components of a calibration system is of particular relevance to obtain low measurement uncertainty. This is true in addition to the periodic retracing of the employed reference standards to the relevant national standard, which continues to be a necessary measure but is not enough.

Manufacturers of calibration systems by the method of comparison according to [1] ignore this issue more often than not. They argue that due to the evaluation of quotients many error variables such as the amplitude and phase frequency responses simply divide out. However, this argument is true only for an almost identical setting of both channels. It does not hold any more if reference standard and calibration item have transfer coefficients that differ widely.

The situation worsens if the electronic unit of a calibration system is used without any limitation also in a primary calibration system according to [2], since in this case total uncertainty is no longer determined by the small uncertainty contribution resulting from the measurement of quotients. The absolute uncertainty of measuring very small alternating voltage or alternating charge is important, as in this case both the vibration quantity and the electrical quantity are measured at the same time using different physical methods.

Actually the required high absolute precision in measuring extremely small alternating voltage and alternating charge in the frequency range 0.15 Hz to 50 kHz at amplitudes down to a few  $\mu\text{V}$  or pC calls for enormous

efforts in the final testing of the electronic units, requiring, among other things, the determination of the complex frequency response at all possible settings of gain and at all cut-off frequencies of the high-pass and low-pass filters. To do so, instruments and standards are needed that have to be recalibrated periodically by the metrological institute of the relevant country (PTB in Germany). Apart from that, it is mandatory that the electronic unit tested must have excellent stability with time.

On the one hand, a wide variety of instruments for measuring alternating voltage and alternating charge is available in the commercial market, covering wide ranges of frequency and amplitude and safely traceable to the respective national standards. On the other hand, no measuring instrument has been available as yet for measuring the phase difference between two alternating voltages having different amplitudes. SPEKTRA, a manufacturer of calibration systems, was forced, therefore, to develop an instrument of this kind and retrace it to the national standards. This was achieved in close cooperation with the PTB in Braunschweig, Germany.

In two work stages a solution was developed for the frequency range between 0.15 Hz and 12.5 kHz (without traceability it works up to 50 kHz), where the results of the first stage have been demonstrated in [3].

### 2. CONCEPT AND IMPLEMENTATION

#### 2.1. Measurement system

The measurement system shown in Fig.1 is made up of two broad-band amplifiers of identical design with programmable gain, two identical analog-to-digital converters, and a computer-based evaluating unit.

In the first step, assume that the frequency response of their complex gain coefficients is known (see below for the method of determination). Furthermore, assume that both analog-to-digital converters have identical conversion and run times and that they are triggered by the same clock.

There are two different reasons for using sine approximation for evaluating the signals. The commonly used methods of FFT or DFT as demonstrated in [4] and [5], on the one hand, yield exact results only for signals obtained by absolutely coherent sampling according to (11). Furthermore ISO 16063-11 [2] as the international standard specifying binding requirements for primary calibration,

definitely demands sine approximation. For many years SPEKTRA has been able to gather experience especially with this algorithm because the instrumentation shown in Fig.1 has been similarly employed in primary calibration systems.

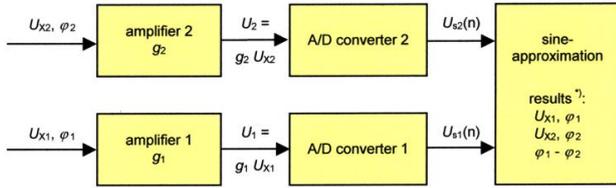


Fig. 1. Block diagram of the measuring setup \*) with reference to both inputs

### 2.2 Basic principle of sine approximation

Assuming clock-synchronized (i. e., phase-locked) sampling, the dual-channel measurement of a transient event will yield the following sequences of numbers at the outputs of the two analogue-to-digital converters:

$$\{u_1(0), u_1(T_a), u_1(2T_a), \dots, u_1((N-1)T_a)\} \quad (1)$$

$$\{u_2(0), u_2(T_a), u_2(2T_a), \dots, u_2((N-1)T_a)\} \quad (2)$$

where

$T_a$  : sampling period       $N$ : record length (block length)

The expression searched for is a set of two functions

$$s_1(n) = s_1(nT_a) = u_{OFS1} + \hat{u}_1 \sin(\omega nT_a + \varphi_1) \quad (3)$$

$$s_2(n) = s_2(nT_a) = u_{OFS2} + \hat{u}_2 \sin(\omega nT_a + \varphi_2) \quad (4)$$

that are selected such that the following relations apply:

$$SU_1 = \sum_0^{N-1} [s_1(n) - u_1(n)]^2 \Rightarrow 0 \quad (5)$$

$$SU_2 = \sum_0^{N-1} [s_2(n) - u_2(n)]^2 \Rightarrow 0 \quad (6)$$

These conditions are met if their partial derivatives disappear simultaneously:

$$\frac{\partial SU_i}{\partial u_{OFS i}} = 0, \quad \frac{\partial SU_i}{\partial \hat{u}_i} = 0, \quad \frac{\partial SU_i}{\partial \varphi_i} = 0, \quad \frac{\partial SU_i}{\partial \omega} = 0, \quad (7) - (10)$$

If the frequency is known, so is  $\omega = 2\pi f$ , and in the special case of coherence as given by

$$nT_a = mT_{meas} \quad \text{with } n, m \text{ integers, } n > 2m \quad (11)$$

and  $T_{meas} = 1/f$

the following solutions are obtained for  $i = 1, 2$

$$\tan \varphi_i = \frac{\sum_1^N U_i(n) \cos[x(n)]}{\sum_1^N U_i(n) \sin[x(n)]} \quad \text{where } x(n) = 2\pi f T_a n \quad (12)$$

$$u_{pi} = \frac{2}{N} \sum_1^N U_i(n) \sin[x(n) + \varphi_i] \quad (13)$$

$$u_{OFSi} = \frac{1}{N} \sum_1^N U_i(n) \quad (14)$$

The record length  $N$  is selected such that the memory contains an integer number of periods  $T_{meas}$ .

As this algorithm will also be used for measuring unknown vibration quantities – in which case the frequency is not normally known and the special case of coherence is met only at selected frequencies – a more general approach is needed. In this case the system of equations (5)–(10) is solved by the iteration method, with the approximate solutions according to (12)–(14) used as initial values.

These constraints do not apply to retracing to the national standard, as in this case the system can be calibrated at the selected frequencies according to (11). In this case, the DFT method may be preferred because it is easier to handle.

### 2.3 Structure of the analogue channels

Figure 2 shows the basic structure of one analogue channel.

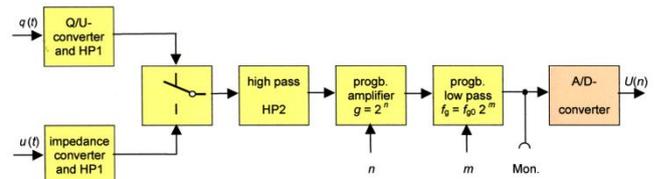


Fig. 2. Structure of an analogue channel

Depending on the physical quantity to be measured, the signal first passes through one of two converters with integrated high-pass filters. The charge-to-voltage converter is used if the input signal is alternating charge, whereas the impedance converter is needed for an alternating voltage signal. Having passed the converter, the signal is fed through a multiplexer and a second high-pass filter to a programmable broad-band amplifier, the gain of which is switchable in powers of 2 (approximately 6 dB). In this way the AD converter that follows the low-pass filter can be driven with a signal of optimum amplitude.

The two high-pass filters are switched such that their resulting total transfer function corresponds to a cut-off frequency of 0.2 Hz or 2 Hz, respectively. There is a choice of 14 possible values of gain in the range between 0 and 78 dB. The cut-off frequency setting of the programmable low-pass filter is normally linked rigidly to the sampling frequency of the analog-to-digital converter such that unwanted signals occurring outside the utilized frequency range are suppressed.

If the instrument is used as a reference standard, the amplitude-frequency and phase-frequency responses between the monitor output and the two possible signal inputs need to be determined for

- all settings of gain,
- both cut-off frequencies of the high-pass filters,
- all used cut-off frequencies of the low-pass filter

in the entire frequency range between 0.15 Hz and 50 kHz.

Since the characteristics of the complex transfer function can be described by relatively simple approximation functions, it will be enough to determine these characteristics only at selected frequency points.

An experienced measurement expert will have no trouble determining the amplitude-frequency response, particularly since the necessary instrumentation is available in the commercial market. The determination of the phase-frequency response, however, is actually an engineering challenge that will be shortly outlined in the following.

### 2.4 Determination of phase-frequency response

Initially, another instrument according to Fig. 1 was constructed with channels of identical design. If both channels of such a device are fed by a joint low-impedance source of alternating voltage and if both channels are set to identical values of gain and cut-off frequencies, the set value of the angle of phase difference is zero due to the identical design of the channels. In practice, however, there are always some unavoidable differences between the utilized components, so the obtained value will differ slightly from zero and should be stored as a systematic deviation that may later be used for corrections.

Figure 3 shows the total setup for the determination of the angle of phase difference between monitor output and signal input.

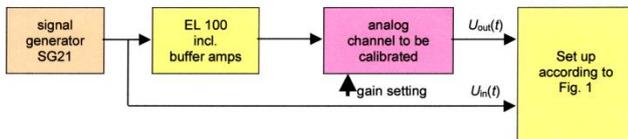


Fig. 3. Measuring setup of the angle of phase difference

Step by step the gain coefficient of the analogue channel under calibration is set to values of 0, 6, 12, ... 78 dB. At the same time the attenuation of the programmable attenuator type EL 100 is set to the same values. As a consequence, both voltages  $U_{out}$  and  $U_{in}$  are almost identical in magnitude and the instrumentation yields the angle of phase difference of the tandem connection of the analogue channel under calibration and the programmable attenuator.

To determine the yet unknown angle of phase difference of the attenuator, the analogue unit under calibration in Fig. 3 is temporarily replaced with a wired connection ( $\varphi = 0$ ) and the following procedure is executed:

**Step 1:** Apply voltage  $U$  in simultaneously to both analogue inputs. Assume that the measured angle of phase difference is  $\varphi_1 - \varphi_2 = \varphi_0$ . Thus the measuring deviation  $MA$  is  $MA_0 = \text{measured value} - \text{expected value} = \varphi_0 - 0 = \varphi_0$

**Step 2:** Connect analogue input 1 to input of attenuator, analogue input 2 to output of attenuator, set attenuator to 0 dB. The measured angle of phase difference is  $\varphi_1(0,0) - \varphi_2(0,0) = \varphi(0,0,0)$ . Thus the corrected value is **measured value** –  $MA_0 = \varphi(0,0,0) - \varphi_0 = \varphi_{DT}(0)$

**Step 3:** Same as Step 2, but with attenuator set to 6 dB. The measured angle of phase difference is  $\varphi_1(6,0) - \varphi_2(6,0) = \varphi(0,6,0)$ . The corrected value is **measured value** –  $MA_0 = \varphi(0,6,0) - \varphi_0 = \varphi_{DT}(6)$

**Step 4:** Same as Step 3, but with gain of channel 2 set to 6 dB. The measured angle of phase difference is  $\varphi_1(6,6) - \varphi_2(6,6) = \varphi(0,6,6)$ . If this value  $\varphi(0,6,6)$  differs from the result of Step 3  $\varphi_{DT}(6)$ , this difference can only be caused by the change of gain in channel 2. Consequently **MA**<sub>6</sub> of channel 2 for  $v = 6$  dB is **measured value** – **expected value** =  $\varphi(0,6,6) - \varphi_{DT}(6) = \varphi_{CHAN2}(6)$

**Step 5:** Same as Step 4, but with attenuator set to 12 dB. The measured angle of phase difference is  $\varphi_1(12,0) - \varphi_2(12,6) = \varphi(0,12,6)$ . The corrected value is **measured value** –  $MA_6 = \varphi(0,12,6) - \varphi_{CHAN2}(6) = \varphi_{DT}(12)$

In this way the basic principle becomes obvious: Any incremental increase of phase measured in the above step-by-step procedure is caused

- by the attenuator (DT) for those steps where gain has not been changed
- or by the analogue channel 2 for those steps where the attenuation  $D$  of the attenuator DT has not been changed.

Figure 7 (at the end of this paper) shows the phase-frequency response characteristics of the programmable attenuator determined in the manner described above for attenuation settings of 0, 6, 12, ... 66 dB.

A by-product of this procedure is the phase error of the analogue channel 2 as functions of frequency and gain setting, related here to the absolute value of the angle of phase difference for gain  $v = 0$  dB. By swapping the channels and performing this procedure one more time, the phase error of analogue channel 1 will be obtained. This additional information may be very helpful for any possible verification procedure that will be carried out later.

Figure 4 shows two typical phase-frequency response characteristics for the arbitrarily selected gain coefficients 12 dB and 60 dB.

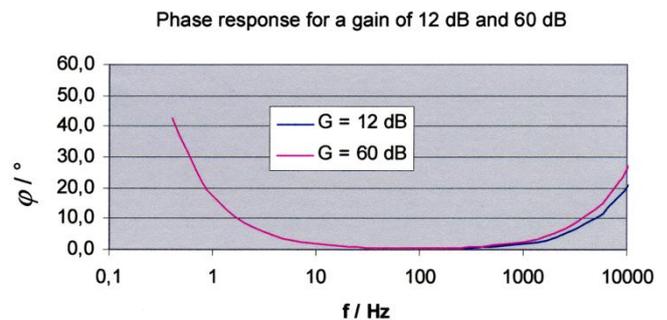


Fig. 4. Phase-frequency response of one analog channel

A well-suited approximation function is

$$\varphi(G_i, f) = \alpha_i f + \arctan\left(\frac{f_{g1}}{f}\right) + \arctan\left(\frac{f_{g2}}{f}\right) \quad (15)$$

where  $f_{g1}, f_{g2}$  = cut-off frequencies of high-pass filters HP1, HP2 in Fig. 2 provided that the cut-off frequency of the low-pass filter is approximately 65 kHz or higher.

The characteristics shown in Fig. 4 show quite clearly that in primary calibration no usable results can be obtained unless phase is adequately corrected, since the laser vibrometer used for measuring the vibration quantity has an absolutely linear phase response, caused solely by the

constant run time, whereas the analogue output signal of the device under test is subject to the phase response according to Fig. 4.

**2.5 Application**

After the angles of phase difference

$$\Delta\varphi_1 = \Delta\varphi_1(g_i, \alpha_i, f_{g11}, f_{g12}, f) \tag{16}$$

$$\Delta\varphi_2 = \Delta\varphi_2(g_j, \alpha_j, f_{g21}, f_{g22}, f) \tag{17}$$

have been determined for all possible values of i, j, using an equation similar to (15), and their values have been stored in a table, the instrumentation according to Fig.1 can be used for measuring the angle of phase difference between any two signals of the same frequency, regardless of whether their amplitudes are equal or different.

The two input signals are described by their amplitudes and zero phase angles at their common frequency  $f_n$  as follows:

$$U_{X1}, \varphi_{X1} \text{ and } U_{X2}, \varphi_{X2};$$

the uncorrected measurement result is named  $\Delta\varphi_{MEAS}$ .

In that case, the following function applies for selected frequencies  $f_n$ , on the basis of (16) and (17):

$$\varphi_{X1} - \varphi_{X2} = \Delta\varphi_{MEAS} - (\Delta\varphi_1 - \Delta\varphi_2) = \Delta\varphi_{MEAS} - \partial\varphi \tag{18}$$

Table 1. Technical Data of SPEKTRA CS18 PH calibration hardware

Item	Value
Physical principle:	Sine approximation
Measuring range phase	0 to 360 ° or -180° to +180°
Measuring range amplitude	1 mV to 10 V / 1 pC to 10 nC
Frequency range	0.15 Hz to 50 kHz
Maximum level difference	60 dB
Ranges with traceable phase at present	0.15 Hz to 12.5 kHz and 2.5 mV to 5 V
Planned extension of traceable voltage range	1 mV to 10 V
Expanded uncertainties without $\delta\varphi$ correction	< 0,2% ; < 0,01°



Fig. 5. SPEKTRA CS18 PH calibration hardware

**3. TRACEABILITY TO THE NATIONAL STANDARD**

In the context of the abovementioned research cooperation, a standard measuring instrument for phase meters with input voltages between 2.5 mV and 5 V in the frequency range from 0.15 Hz to 12.5 kHz was developed at the laboratory for electrical quantities of PTB Braunschweig, Germany, based on [5–7], (see Fig. 6). The results obtained, presented in the CIMM-PTB seminar in Braunschweig [3], were excellent. According to them, the measured deviations  $\delta\varphi$  and  $\delta U$  between SPEKTRA and PTB were less than  $\pm 0.01^\circ$  or 0.15%, respectively, in the entire frequency range and even for signals with extremely different amplitudes (level difference 66 dB).

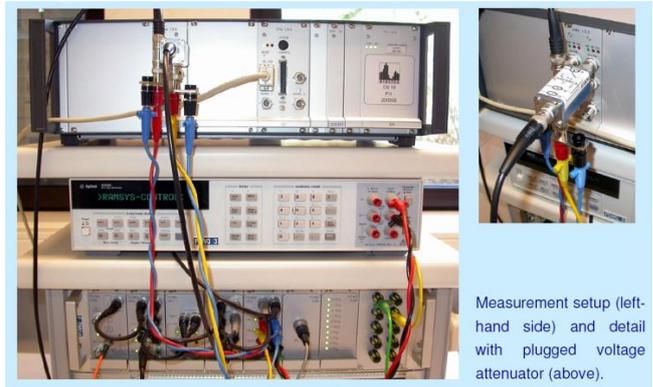


Fig. 6. Measuring setup of the PTB

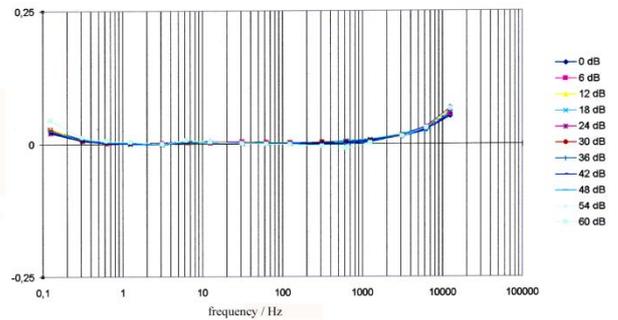


Fig. 7. Phase deviation between SPEKTRA and PTB

On the one hand, the traceability to the national standard allows the verification of results obtained by means of (18). On the other hand, the measurement error  $\delta\varphi$  stated by the PTB can be used for correcting the measured values, as the PTB value is regarded as more trustworthy. Now the attenuator EL 100 (see. Fig. 3) can be calibrated directly with respect to its angle of phase difference, and hence, much smaller measurement errors can be expected when performing the next traceability calibration.

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