

HIGH-PRECISION MEASUREMENT OF STRAIN GAUGE TRANSDUCERS AT THE PHYSICAL LIMIT WITHOUT ANY CALIBRATION INTERRUPTIONS

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Abstract: The maximum resolution when measuring transducers, which operate on the strain gage principle, is physically limited. Possible sources of errors and their compensation to achieve the high accuracy class of 0.0005 (5 ppm) are shown. In addition to the actual extraordinary high measuring accuracy and long term stability, for the new precision instrument DMP41 the auto-calibration cycles could also have been omitted. An innovative method allows to measure for the first time in this accuracy class without any interruptions caused by an auto-calibration.

Keywords: precision instrument, strain gauge, physical limit, high resolution, high stability, auto- and background-calibration

1. INTRODUCTION

The resistance of a strain gauge changes under mechanical load. If several strain gauges are combined to a bridge circuit the ratio of the bridge output to the bridge excitation voltage is nearly proportional to the mechanical applied force. For the electrical measurement of mechanical quantities using strain gauges, it is the ratio of the voltages expressed in mV/V, which has great importance. The measured mechanical quantities are captured using transducers and are mapped into the unit mV/V.

The highest accuracies in the area of force and force comparison measurements are required at national and international levels from government institutions (e.g. Physikalisch Technische Bundesanstalt PTB (Germany), National Institute of Standards and Technology NIST (USA)). The goal is a reproducible error from just a few millionths. In the industrial sector precision measuring instruments are used as special measuring equipment for the development of precision force transducers or as reference instrument in calibration laboratories. For these applications, the precision measuring instruments are required. Since the launch of the DMP series in 1980 [1], HBM reached in the class of precision sensor amplifier, that operate on the strain gauge principle, the absolute physical limit in terms of the resolution. Metrological institutes worldwide therefore rely on DMP precision amplifiers from HBM.

Figure 1 shows the new model of the DMP series from the front and back side, the DMP41. Although the predecessors of DMP41 have been measuring at the physical limit, the technical measurement characteristics could be further improved. The new

precision amplifier now supports the simultaneous measurement of all channels and is much less sensitive to electromagnetic influences (EMC).

For the users a completely new feature will bring a decisive advantage. It is called the "background-calibration" (or "background-adjustment"). In addition to the current extraordinary high measuring accuracy and long term stability, the auto-calibration cycles are also omitted now for the new precision instrument DMP41. For the first time, it is possible to measure in this accuracy class without interruptions by an auto-calibration. Compared to previous devices of the DMP series, any auto-calibration cycle caused a data stream interruption, even by auto-triggered calibration cycles. In that case, that the user cannot tolerate any interruption during the measurement, it was always necessary to trigger the calibration before executing a measurement.

From the user's point of view mainly three questions arise: With respect to the DMP39 and DMP40, is there any way to increase the precision? How is the operating principle of the carrier frequency technology and which benefits arise thereby? Is the background-calibration process visible in the measured data, what influence has the method on the measurement accuracy and how is it realized? This paper will give an answer to these questions.

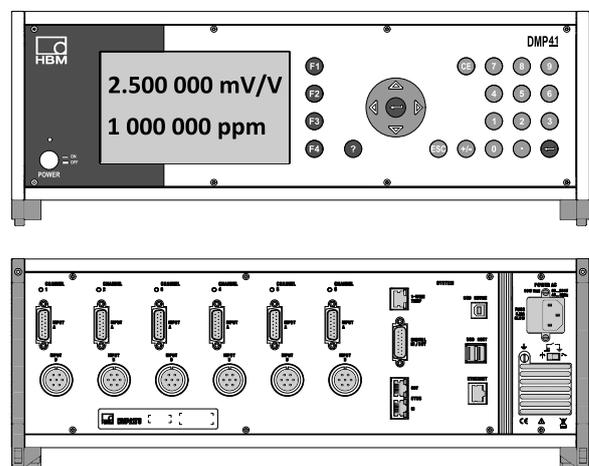


Figure 1: Front and back side of the DMP41 as table housing version

2. THERMAL NOISE VOLTAGE (RESOLUTION LIMIT)

The purpose of an amplifier is a desired change (e.g. gain) of the input signal. In the real world, there are also noise sources which cannot be eliminated. This noise sources will be amplified with the useful signal and limit the signal to noise ratio and therefore the maximum resolution. Noise is a random signal. It changes the actual value of the signal and cannot be predicted at any time.

The noise is internally generated in the amplifier, but also by the external passive components (e.g. resistors) and that's why noise is always a significant problem for engineers by the development of highly sensitive amplifiers, like the DMP41. With a good understanding of the basics of noise in amplifier circuits, the noise can be reduced significantly.

When measuring strain gauges, the noise voltage of the sensor and amplifier electronics limits the reachable resolution [2]. With high effort, the noise voltage of the amplifier can be kept very small. The noise of the amplifier depends mainly on the used operational amplifiers and the circuit technology. Even if the amplifier would be ideal and has no intrinsic noise, the maximum possible resolution is still limited by the thermal resistance noise of real resistors (strain gauges are resistive).

Disorderly movements of electrons in real resistors evoke thermal noise and generate a noise voltage density. This noise is present in all passive resistive components. The mobility of electrons increases with temperature and therefore also the thermal noise increases with temperature. Increasing of the resistance shows the same behavior.

The noise voltage density e_R of a resistive resistor is given by the temperature T , the resistance R and the Boltzmann constant k (1). A noise density is a noise voltage or current relative to the root of the frequency. Normally the spectral noise densities specify the noise parameters. For high-precision measurements with strain gauge transducers bridge resistance of 350Ω are typically used in full bridge circuits. The source resistance seen by the amplifier is therefore also 350Ω . For the absolute room temperature T of 295.16 K ($22 \text{ }^\circ\text{C}$) and an ideal amplifier, the transducer R of 350Ω itself produces a noise voltage density e_R of about $2.39 \text{ nV}/\sqrt{\text{Hz}}$.

$$e_R = \sqrt{4TRk} = \underline{2.39 \frac{\text{nV}}{\sqrt{\text{Hz}}}} \quad k = 1.38 \cdot 10^{-23} \frac{\text{J}}{\text{K}} \quad (1)$$

The noise corresponds to a statistically Gaussian distribution and it is called white noise. It is independent of frequency (uniform power distribution). To get the effective noise voltage $V_{N,rms}$, the voltage noise density e_R is integrated over the frequency f (2). Because the thermal noise is constant over frequency, the noise voltage density e_R is just multiplied by the root of the bandwidth B to get the effective noise voltage $V_{N,rms}$ (taking into account the equivalent noise bandwidth Eq_{NBW}). A limitation of the bandwidth by low-pass filtering leads to a much lower voltage noise. For a measurement with a low pass filter with a bandwidth B of 1 Hz , the transducer R of 350Ω itself produces an effective noise voltage $V_{N,rms}$ of about 2.39 nV_{rms} .

$$V_{N,rms} = \sqrt{\int_{f_{low}}^{f_{high}} e_R(f)^2 df} \stackrel{\text{white noise}}{\triangleq} e_R \sqrt{B} = \underline{2.39 \text{ nV}_{rms}} \quad (2)$$

The effective noise voltage is not direct representing the achievable resolution. Essential for the maximum resolution is the peak-peak value of the noise.

Figure 2 shows distribution of the noise. The peak-peak value can be estimated from the effective noise voltage. The noise corresponds statistically to a Gaussian distribution. This distribution is used to calculate the peak-peak noise. Therefore the standard deviation must be known.

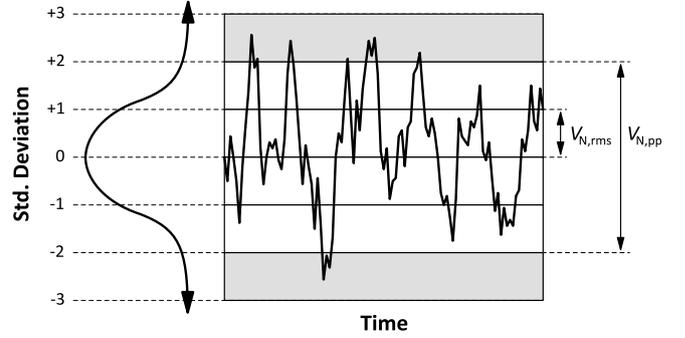


Figure 2: Noise distribution

The effective noise voltage $V_{N,rms}$ is equal to the standard deviation σ for that case, the constant component μ of the noise voltage is zero (3). This simplification applies only for that cases where the effective noise voltage contains no dc component. Thermal noise has no constant component (dc component), the standard deviation and the effective noise voltage are identical. The standard deviation is thus normalized to one ($\sigma = V_{N,rms}$).

$$\sigma = \sqrt{\frac{1}{n} \sum_{i=0}^{n-1} (x_i - \mu)^2} \triangleq \sqrt{\frac{1}{T} \int_0^T x(t)^2 dt} = \underline{V_{N,rms}} \quad (3)$$

In the world of technology it is common to calculate with a standard deviation of two (2σ). This means that the peak-peak noise voltage $V_{N,pp}$ is four times larger than the effective noise voltage $V_{N,rms}$ (4). The peak-peak noise voltage $V_{N,pp}$ of a full bridge is therefore 9.56 nV .

$$V_{N,pp} = \underbrace{2}_{\text{peak-peak}} \cdot 2\sigma = 4V_{N,rms} = \underline{9.56 \text{ nV}} \quad (4)$$

With a standard deviation of $n = 2$ the noise floor is 95.45% (probability Φ) of the measured time below the peak-peak noise (5). A probability specification is part of every noise value.

$$\Phi(n) = \int_{-n}^{+n} \frac{1}{\sqrt{2\pi}} e^{-\frac{1}{2}x^2} dx \triangleq \underline{95,45 \%} \quad (5)$$

When converting the actual values into peak-peak values, it is very important to note, that the outlets of the Gaussian distribution are infinite, so much higher noise voltages can occur. The probability of noise voltages with extremely high momentary values is very low. For a threshold of $n = 4$ (a doubling of the peak-peak noise), values exceed the threshold 0.006% (probability Φ) of the time (6).

3. CARRIER FREQUENCY TECHNOLOGY

$$\Phi(n) = 2 \int_n^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{1}{2}x^2} dx \triangleq \underline{0.006 \%} \quad (6)$$

The maximum resolution is reached by definition, when the resolution step is equal to the peak-peak noise voltage. In that case, the measured value display shows a constant value with a given probability. The measured value changes with a standard deviation of two only 4.55 % of the time to a different value and can be considered as stable.

The resolution limit can be determined for a common measuring range by the signal to noise ratio SNR of the measurement [3]. The signal to noise ratio is the measurement signal (excitation voltage V_e multiplied by the measuring range M) in relation to the noise signal $V_{N,pp}$ (7). So the maximum resolution only of a transducer with a resistance of 350Ω in a full bridge circuit using a measuring range of 2.5 mV/V at $5 V_{rms}$ is limited to 0.76 ppm at a low pass filter of 1 Hz .

$$SNR = \frac{V_e M}{V_{N,pp}} \triangleq \underline{0.76 \text{ ppm}} \quad (7)$$

A high resolution can be achieved only in a small bandwidth. In the calculating, the amplifier's noise was not taken into account. Figure 3 shows that only slow measurements can achieve high resolution. Reverse, fast measurements can be performed with a lower resolution. Precision measurement technology means slow measurements. Dynamic measurements cannot be performed with the highest accuracy and resolution.

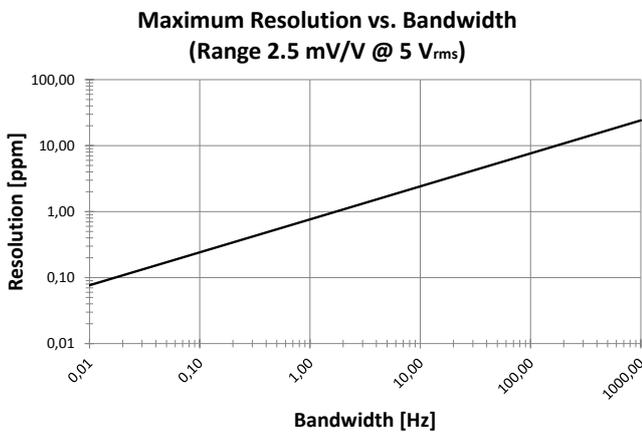


Figure 3: Maximum resolution vs. Bandwidth

The resolution of the strain gauge technology is limited by the resistive resistance and limits the resolution of the strain gauge technology. Through the amplifier, the signal to noise ratio is further reduced. This is due to additional resistive resistances and to the voltage and current noise of the operational amplifiers.

With excellent amplifiers without extra features or protection circuit it is possible to reach a noise floor of the complete measuring chain around twice the noise of the transducer itself. The devices of the DMP series have converged since the DMP39 up to a factor of 2 to the absolute physical resolution limit. An improvement is not possible due to the physical conditions.

Besides the noise, which physically limits the maximum achievable resolution, there are other error sources. These include quasi-static interferences like thermocouple voltages and offset drifts of the amplifier, but also interferences in certain frequency ranges due to inductive or capacitive couplings (e.g. power line and high frequency electromagnetic interferences).

To eliminate this type of error influences the carrier frequency method has been proven successful for decades [4]. The transducer is supplied with an alternating excitation voltage and works as a modulator. In this way, an interfering signal (AC or DC) can be separated from the measurement signal of the same frequency by the carrier frequency method.

Using the Fourier Transformation and the convolution [5] in the frequency domain, the modulated input signal $F_m(\omega)$ will be determined (consisting of the transducer signal $f_t(t)$ and the carrier frequency $f_c(t)$) (8).

$$f_t(t) = \cos(\omega_t t) \quad \text{O} \text{---} \bullet \quad F_t(\omega) = \pi[\delta(\omega + \omega_t) + \delta(\omega - \omega_t)]$$

$$f_c(t) = \cos(\omega_c t) \quad \text{O} \text{---} \bullet \quad F_c(\omega) = \pi[\delta(\omega + \omega_c) + \delta(\omega - \omega_c)]$$

$$f_m(t) = f_t(t)f_c(t) \quad \text{O} \text{---} \bullet \quad F_m(\omega) = \frac{1}{2\pi}[F_t(\omega) * F_c(\omega)]$$

$$F_m(\omega) = \frac{1}{2}[F_t(\omega + \omega_c) + F_t(\omega - \omega_c)]$$

$$= \frac{\pi}{2} \left[\underbrace{\delta(\omega + \omega_c + \omega_t) + \delta(\omega + \omega_c - \omega_t)}_{\text{negative side bands}} \quad (8) \right. \\ \left. + \underbrace{\delta(\omega - \omega_c + \omega_t) + \delta(\omega - \omega_c - \omega_t)}_{\text{positive side bands}} \right]$$

Figure 4 shows the signals at the amplifier input and at the demodulator output in the frequency domain. The amplitude of the carrier frequency represents a static measurement signal (1). A dynamic measurement signal (2) within the bandwidth of the amplifier is divided into two frequency components. These frequency components are located symmetrically around the carrier frequency with half of the amplitude. The distance to the carrier frequency corresponds to the frequency of the measurement signal.

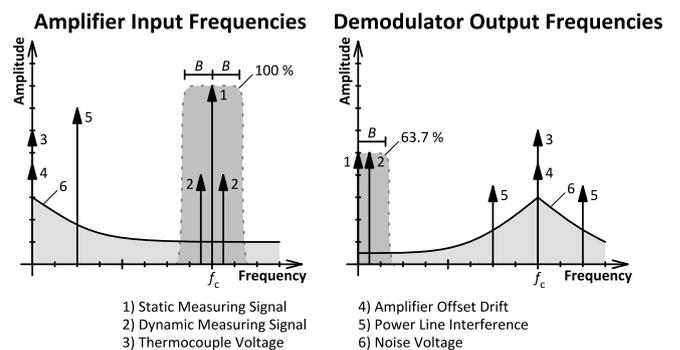


Figure 4: Transformation by the carrier frequency technology

Most of the interference signals are located close to zero. These include thermocouple voltages (3), the amplifier offset drifts (4), the crosstalk of the supply voltage of 50/60 Hz (5) and the 1/f noise of the operational amplifiers (6). By demodulation the signals are transformed. By using again the Fourier Transformation and the convolution in the frequency domain, the demodulated output signal $F_s(\omega)$ will be determined (consisting of the modulated input signal $f_m(t)$ and the demodulator signal $f_d(t)$) (9).

$$f_d(t) = \frac{4}{\pi} \sum_{k=0}^{\infty} \frac{1}{2k+1} \cos((2k+1)\omega_c t)$$

$$F_d(\omega) \underset{\substack{\text{band-pass} \\ k=0}}{\approx} \frac{4}{\pi} [\pi[\delta(\omega + \omega_c) + \delta(\omega - \omega_c) + \dots]]$$

$$f_s(t) = f_m(t)f_d(t) \quad \text{O} \text{---} \bullet \quad F_s(\omega) = \frac{1}{2\pi} [F_m(\omega) * F_d(\omega)] \quad (9)$$

$$F_s(\omega) = \frac{2}{\pi} [F_m(\omega + \omega_c) + F_m(\omega - \omega_c)]$$

$$\underset{\text{low-pass}}{\approx} 1[2[\delta(\omega + \omega_c) + \delta(\omega - \omega_c) + \dots]]$$

The low frequency interference signals are mirrored in the region of the carrier frequency and the measuring signal is reflected in the area around zero. All the interference signals are outside the bandwidth of the amplifier and therefore are all filtered completely out. The measurement signal corresponds again to the pure modulated transducer signal without any interference. All interference signals located outside the narrow bandwidth around the carrier frequency are completely suppressed by the carrier frequency method.

However, the carrier frequency method limits the bandwidth. Precision measurement technology means slow measurements. The carrier frequency technology with a low sinusoidal carrier frequency of 225 Hz (free of harmonics) is for high precision measurements the right choice.

4. OPERATION OF THE DMP41

Figure 5 shows a simplified block diagram of a single measuring channel of the DMP41 [6]. In order to reach the required noise suppression, zero and display stability, the sensors are supplied with a low carrier frequency of 225 Hz.

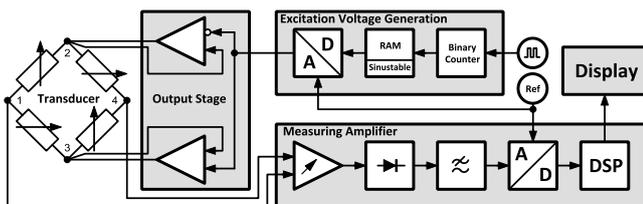


Figure 5: Simplified block diagram of a DMP41 channel

To obtain a very frequency- and amplitude stable bridge excitation voltage, the 225 Hz sine wave voltage is digitally generated. The discrete amplitude values for one period are stored in a table.

All values stored in this table are transmitted 225 times per second to a digital-to-analog converter (DAC). At the output of the converter occurs a composite from many levels stable sine wave. The waveform is smoothed and converted into a symmetrical bridge supply voltage referenced to ground by an additional circuit. By varying the time-discrete values of the table, a variation of the amplitude depending on the reference voltage can be realized. The nominal excitation voltage is fed to an output stage. The bridge excitation voltage can be finely adjusted in amplitude, offset and phase and has no significant harmonics.

As well as its predecessor the DMP41 operates with a six-wire circuit to eliminate the cable influences, two sense leads from the excitation voltage of the transducers are fed back to the output stage [7]. The output stage regulates voltage drops across the cable from the supply voltage to the transducer. A circuit part (not shown) monitors the sensor lines and the deviation of the output stage. Errors are immediately reported.

Internally, the measurement signal inputs (diagonal) and the sense lines of the supply voltage inputs are implemented identically. Voltage drops and phase shifts thereby effect on the sense lines in the same way as on the measuring lines (using balancing resistors $R_s/2$ for long cables to get the same source resistance for the sense leads like the measuring signal) [8]. The voltage drop across the cables can be compensated by the ratio metric measurement method.

The connected sensor divides the excitation voltage down depending on the modulation in a certain ratio (mV/V). The resulting measuring signal is first amplified in the amplifier, band pass filtered (not shown in the figure), demodulated, low pass filtered and then digitized. Depending on the measuring range, the gain of the amplifier stage can be adjusted. The amplifier stage is constructed using low-noise operational amplifiers in order to reach the required resolution. The input impedance of the amplifier is very high to avoid a loading of the transducer by the amplifier. The sensitivity remains constant regardless of the internal resistance of the sensor.

Because of the ratio metric measurement method the stability of the measurement depends only on the resistance ratio of some precision resistors, small changes are additionally corrected by the auto-calibration cycles (now background-calibration cycles). The measurement signal is adjusted in the DSP (Digital Signal Processor) in accordance with the auto-calibration values and the selected filter. The entire control of the amplifier section and the data acquisition is realized in real time via a DSP and a soft core processor on the FPGA (Field Programmable Gate Array).

5. NEW BACKGROUND-CALIBRATION

In a background-calibration (precise: adjustment) the drift of the amplifier is periodically compensated with an internal reference to keep the high class accuracy constant over temperature and lifetime. An internal calibration divider attenuates the sensed excitation voltage down by a certain factor and provides the required reference signal in mV/V depending on the selected measuring range. This ratio must be very accurate, temperature- and long-term stable.

The required temperature and long-term stability cannot be achieved only by using high precision resistors. Due to the alternating voltage by using the carrier frequency method an inductive divider can be used. Through the inductive division method, the temperature drift is extremely low and the long term stability is ensured.

Figure 6 shows a simplified circuit of the internal inductive calibration divider. The inductive divider is a special transformer. Because no inductive dividers are available on the market which meets the requirements of the DMP precision amplifiers, an own inductive divider has been developed by HBM. This specially designed divider provides the required dividing ratios by the number of windings and generates the calibration signal V_{cal} out of the sensed excitation voltage V_{sens} . This is very temperature- (< 2 ppm per 10 K) and long-term stable - *a winding cannot disappear!*

For almost ideal inductive dividers relatively large cores and twisted wires are used. The losses of the primary winding's copper resistance are compensated by a current-free sensor winding. In this way, the magnetic flux is controlled and the division ratio is exactly represented by the winding ratio alone. Different taps on the secondary side provide reference signals for all measuring ranges.

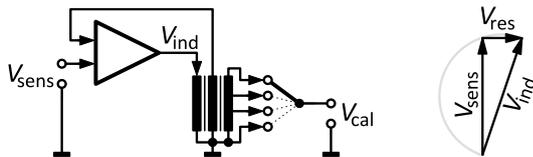


Figure 6: Circuit of the internal inductive calibration divider

The accuracy and stability of the precision measuring amplifier depends entirely alone on this inductive component. This method of generating the required reference signals has been proven for decades in the predecessors of DMP41. The long term stability is monitored at a pair of devices and remains within a range of 2.5 ppm since 1981. Within the device, this is the stable and reliable reference.

The block diagram shown in Figure 5 will be extended for the background-calibration by two other circuit components. The specially developed inductive divider is used as the stable reference. Compared to its predecessors the DMP41 has got an additional calibration amplifier (identical to the measuring amplifier) for background-calibration purposes yet. This second amplifier performs the measurement for a short moment during the background-calibration.

The extended circuit with the components for the background-calibration is highly simplified shown in Figure 7. The actual excitation voltage of the transducer is fed back via the six-wire circuit and applied to the inductive divider. The calibration signals at the output can be picked for the background-calibration.

The innovation is implemented in the DMP41 background-calibration process with the now available additional calibration amplifier. While executing a background-calibration, a normal and unchanged measurement within the device specifications can be taken on all channels. The measurement is not interrupted or affected in terms of accuracy.

At the beginning of each background-calibration process the calibration amplifier itself is calibrated and the data acquisition is object of the measuring amplifier. The sense line of the excitation voltage is applied to the inductive divider and the relevant calibration signal (0) for the calibration amplifier is picked (1). Zero and gain errors are determined based on two calibration points (2). After the calibration amplifier is successfully calibrated, it can be connected in parallel to the measuring amplifier (3). After the calibration, the additional amplifier could take over the measuring task with the same accuracy.

In the DMP41 are two additional steps added. The identified minimal intrinsic errors of the measuring amplifier at the operating point (e.g. offset and linearity errors) will be transferred to the calibration amplifier (4). To avoid the settling of the calibration amplifier, all filter coefficients and state memories of the digital filter are copied from the measuring amplifier (5). Waiting for the settling of the very slow filter is no longer necessary. Another advantage is the takeover, which is not recognizable in the noise any more.

At the same time the measurement signal of the calibration amplifier is used as the new value (5). The actual measuring amplifier is no longer involved in the measurement and can now be calibrated (6) without interrupting the measurement. A calibration of the measuring amplifier is identical to the calibration process of the calibration amplifier, zero and gain errors are determined (7).

After calibration is accomplished the measuring signal is switched in parallel to the calibration amplifier (8). The filter coefficients and the state memories are copied back to the measuring amplifier (9). At the same time the measured value of the measuring amplifier is again used as the signal (9). Then, the calibration amplifier is separated from the measurement path and parked in an idle state. After a certain time the same background-calibration is triggered again by the automatic system.

The background-calibration cycle takes only a few seconds, depending on the selected filter. During the warming up phase of the device the adjustment of the background-calibration is visible and desirable because the amplifier is drifting slightly and this is compensated. The calibration amplifier has the same structure as the actual measuring amplifier and it is calibrated before each

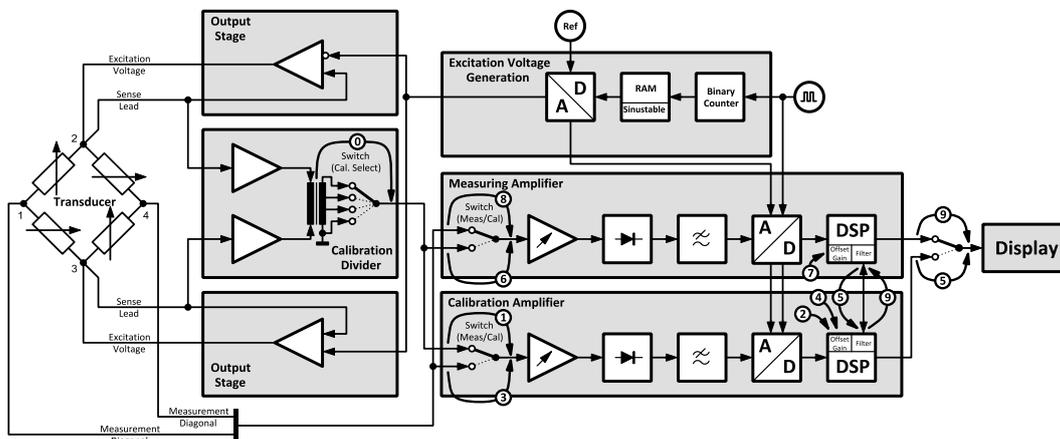


Figure 7: Extended block diagram of a DMP41 channel with an additional calibration amplifier

takeover. Therefore measurements can be taken during the background-calibration process with the same accuracy. The calibration amplifier is additionally minimally adjusted to the calibration of the measuring amplifier to take over its intrinsic errors (e.g. offset and linearity errors).

Measurements at different boundary conditions have shown that this method has no negative or disturbing influences. Figure 8 illustrates a measurement plot of a 2.5 mV/V signal over six hours for a 0.04 Hz filter, where a background-calibration was triggered every minute. During the measurement are many background-calibrations performed. The adjustments are in the sub ppm range. Only the constant noise of the sensor and amplifier are recognizable. There are no interferences visible in the noise floor. Quite the opposite, it is now possible to calibrate during measurements to minimize errors.

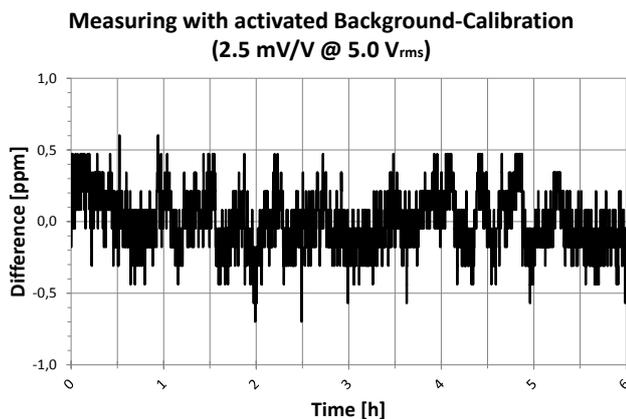


Figure 8: Measuring record with a running background-calibration at a measuring range of 2.5 mV/V and an excitation voltage of 10.0 V_{rms}.

Deactivation of the background-calibration does not improve the accuracy, but it will worsen the accuracy. Because the background-calibration for the user remains completely unnoticed and has also no effect on the accuracy, there is no more control of the auto-calibration provided. The user does no longer have to worry about the active control of the previous auto-calibration to get no interruption when measuring.

Besides the excellent temperature and long-term stability is the absolute measured value important. For the calibration of the DMP41 the transducer is replaced by a bridge calibration unit. In this way, defined voltage conditions can be generate in a purely electrical way with high precision and added to the amplifier [9].

The first factory calibration of the DMP41 is done during production with a recirculated bridge calibration unit (BN100A with an accuracy class of 0.0005 (5ppm)). With a special measuring device, the deviations of the bridge calibration unit can be identified and taken into account during a calibration at the Physikalisch-Technische Bundesanstalt (PTB) [10]. In relation to the values acquired by the Physikalisch-Technische Bundesanstalt the end values of all HBM devices with a carrier frequency of 225 Hz, for reasons of history and continuity, are given 10 ppm lower than the nominal values.

There is no field calibration, the stability is guaranteed through the internal inductive reference and the background-calibration feature. Calibration certificates could confirm and document the accuracy of the DMP41 with respect to national standards, which realize the units of measurement according to the International System of Units (SI).

6. CONCLUSIONS

The resolution of the strain gauge technology is physically limited and cannot be increased further by new amplifiers. This has been demonstrated and justified. There is a contradiction between a high resolution and a high dynamic (high bandwidth). The DMP41 from HBM measures along the physical limit.

The operation of the new internal background-calibration was described in detail in this paper. It was explained and shown that this new feature has no negative influence on the measurement. There is no longer a freezing during the live measurement. Measurements could be done on the physical limit without any interruption. For this reason the possibility to disable the background-calibration cycle in the DMP41 has been omitted.

It should be mentioned at this point that in addition to the new measuring features (e.g. background-calibration, absolutely simultaneously measurements of different channels), the DMP41 comes with up-to-date hardware interfaces and a totally new dual mode graphical user interface (GUI). The user has a great benefit from these new features, options and enhanced usability.

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