Whitepaper

How To Achieve The Best Possible Resistance Measurements

Learn about noise reduction and device choices

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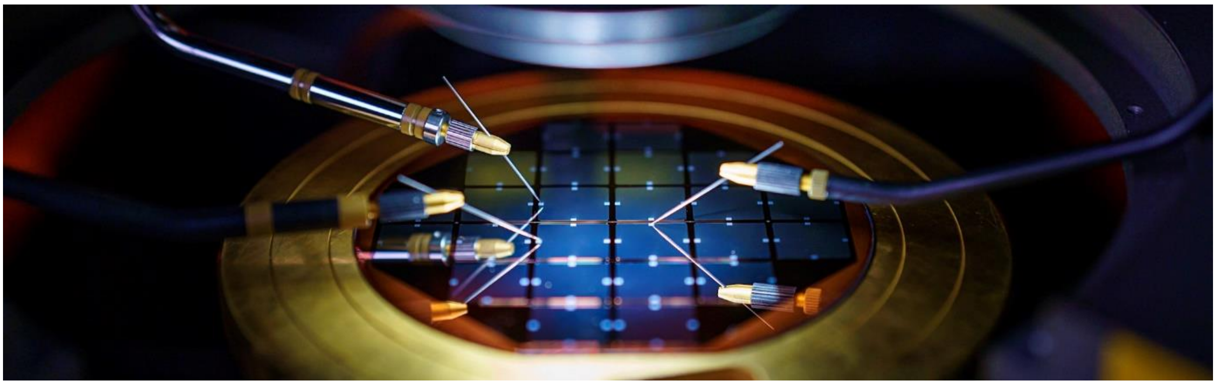

The **electrical resistance of materials** is one of the properties that is most pervasive in modern applications ranging from nanoscale to megascale. Electrical resistance characterization is a simple, yet powerful tool in diverse fields from the research of new materials, over the qualification of new applications all the way to actual field operation.

To stay ahead of your competition, you want to make the best possible resistance measurements. This means: Get the most precise result in the shortest amount of time. But neither can measurements be infinitely accurate, nor infinitely fast. It is important to understand, which bottlenecks in your signal chain can be resolved and which physical limitations exist.

Looking for the best performance usually also means that the basics of resistance measurements and instrumentation are known. Therefore, the main focus of the article is on **precision measurements** for research, sensorics and characterization purposes. In such measurements, integration times range from milliseconds to seconds and excitation frequencies range from DC to 10s of kHz. The measurement times per device span a large range from milliseconds – such as in production monitoring – all the way to days to study subtle effects in condensed matter physics.

Contents

1. Noise Myths

Obtaining the best measurements is equivalent to maximizing the signal-to-noise ratio (SNR) in the target bandwidth. As we take the signal for granted here, this translates to picking up as little noise as possible. Many myths surround this topic and some of them will be discussed in the following. Many of these myths contain situational truth, so it is helpful to know when to do what.

1.1. Longer integration times reduce noise

It is a well-known statistical rule that performing a greater number of experiments brings you closer to the true result as all random variations cancel out. For resistance characterization, this means that increasing the integration time by 2x is expected to reduce the RMS noise by $1/\sqrt{2}$.

However, this trick is effective only until the integration time reaches the inverse of the *1/f-noise corner frequency* f_C. Every active electronic component has both frequency independent noise (*Johnson noise*) as well as frequency dependent noise that rises at low frequencies. When plotted together in a log-log scale, they form a characteristic corner (**Figure 1**). When integrating longer than $1/f_C$, the noisy part of the spectrum below f_C enters the measurement rendering the noise suppression effect of longer integration increasingly useless.

The time $1/f_C$ is a measure of the stability of the underlying measurement system and typical values span from milliseconds (generic amplifiers), to seconds (voltage references and precision amplifiers) all the way to hours (zero-drift equipment). The longest useful integration time is tied to this stability time of the signal chain. More specifically, the longest useful integration time is limited by the shortest stability time of any of the blocks in the signal chain, which also includes the ambient control of the experiment.

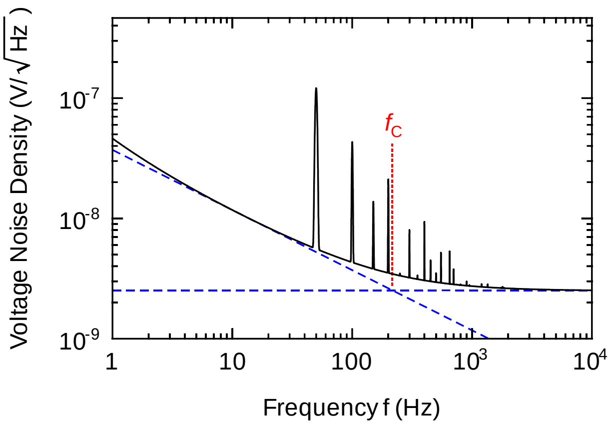

Figure 1: Typical Voltage Noise Spectral Density shown depending on frequency. The black line shows the compound noise spectrum including broadband sources and narrowband electromagnetic interference. The blue dashed lines represent the flat and 1/f broadband noise sources, respectively. Their intersection (red) is called the noise corner frequency f_C.

1.2. More sample current reduces noise

When thinking of electrical characterization as a statistical experiment, the electrons are the test probes. More electrons equal more experiments and thus a better result. This can be achieved in two ways, namely a longer integration time and more electrons per time. As the useful integration time is limited (Section 1.1), the other approach is to improve statistics through a larger probe current.

This works indeed, but again is not infinitely effective. Obviously, increasing the probe current can have disturbing or even destructive effects on the device under test due to resistive heating. But even when not taken to such extremes does increased current flow lead to an increase in noise generated by the sample itself. This additional noise generation is governed by the material *noise index (NI)* and is much more severe for badly conducting compounds, even when comparing samples with the same resistance. Materials such as organic and ionic conductors, conducting oxides and ultra-thin films have a high NI, whereas clean bulk metals, sheets, cables and most metal resistors have a low NI. For badly conducting compounds, using large probe currents rarely offers an improvement in measurement performance, even if the specimen is bulky and has a low resistance.

This also applies to clean pure metal samples when they are covered in e.g. an oxide that is a badly conducting materials. Contacts made to this oxide do not markedly change the resistance reading in a 4-wire measurement, but they do introduce noise when carrying more current.

1.3. AC measurements at irrational frequencies reduce noise

Low frequency electronic noise is fundamentally hard to avoid, as most electronic components in the signal chain generate some level of 1/f noise. To circumvent this problem, the whole signal chain can be shifted in frequency by amplitude modulation using e.g. a fixed frequency sinusoidal carrier. In practice, a modulated current is driven through the sample and the recorded voltage is demodulated using the known modulation pattern. This greatly helps to remove several (not all) sources of low frequency noise (**Figure 1**) and thus useful integration times can be much longer. In addition, phase sensitive measurements become possible, which grants the real and imaginary components of the complex sample impedance.

The drawback of this method is that AC current flows not only through resistive specimens but also through parasitic capacitances and inductances surrounding the experimental stage. As a result, it becomes increasingly challenging to apply AC measurements to specimens with a high resistance of about 1 MΩ and more. In general, measurements results can be harder to interpret and fewer measurement devices are available for precision AC measurements than for DC measurements.

Contrary to widespread belief, there is no advantage in picking an excitation frequency that is a rather odd or even irrational number. Instead, the excitation and demodulation frequencies (which usually have an integer relationship) should be merely far away from narrowband electromagnetic interference such as mains frequency harmonics.

1.4. Shielding reduces noise

Electromagnetic interference (EMI) is ubiquitous in almost every laboratory in the world. It is therefore a prudent idea and a good habit to shield the measurement environment from as much of the interference as possible. Usually, shielding refers to using a metallic container that is connected to ground.

While properly grounded shielding is almost never detrimental to the measurement performance, it also does not always help. First, generic metallic containers cannot screen all EMI. Low frequency magnetic fields can be screened only by special magnetic sheet material with low magnetic coercivity, which is expensive and rarely used. Moreover, low frequency EMI occurs predominantly at the harmonics of the mains frequency. As a result, fixed frequency lock-in amplifiers do not benefit from shielding when tuned to the quiet gaps between the mains harmonics.

1.5. Preamplifiers reduce noise

Preamplifiers are often used for signal conditioning when the signal source is passive or has a large output impedance.

On the other hand, a preamplifier – like any electronic component – *adds* noise to the signal chain. So introducing a preamplifier should always be carefully planned. If the final measurement device has a poor input amplifier, a preamplifier with gain can help. But most modern wall-powered measurement devices have very similar and good input amplifiers and signal conditioning built-in, so that preamplifiers are usually not required for resistance characterization.

2. Reducing Noise

Following general guidelines and myths about noise reduction can sometimes yield good results. For more deterministic performance even in challenging environments, it is necessary to understand the main noise sources. Three different classes of noise will be distinguished in the following three Sections. This distinction is motivated by the different respective countermeasures.

When discussing the topic of precision resistance measurements, there is often an immediate focus on the broadband noise of the input amplifiers. Low input broadband noise allows reaching the base sensitivity in a short amount of time, which is very useful when doing precision long-term measurements (see examples in Section 3). While important, it is – by far – not the one deciding factor for the measurement performance. In many usual measurement scenarios, the input broadband noise is actually the least important of the three sources. At low frequencies and long integration times in particular, 1/f noise arises due to many sources (**Figure 1**). It is this 1/f noise, which always determines the absolute base sensitivity of the measurement. Finally, the output noise of the measurement device can contribute a major part to the total detected noise.

2.1. Reducing broadband input noise

Broadband input noise is frequency independent, which makes it pervasive and unavoidable. Broadband noise has several contributions and it is important to reduce their combined power, instead of only a single contribution. **Figure 2** shows broadband voltage noise densities as a dependence of the sample resistance. The straight lines show the broadband noise caused by the sample resistance at three different sample temperatures, namely at room temperature (300 K), at liquid Helium temperature

(4.2 K) and at typical dilution fridge temperatures (50 mK). There is no known way to avoid this contribution from *thermal Johnson Noise* as it emanates from the sample itself, even when it is completely passive. However, one can attempt to add minimal further noise via the measurement process, which equates to using an amplification stage with low noise at the given sample resistance.

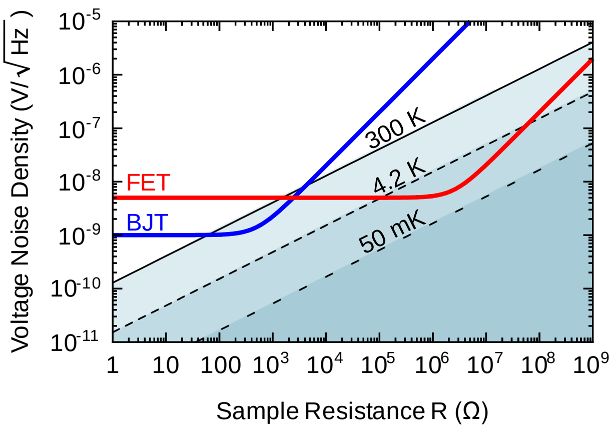

Figure 2: Broadband Voltage Noise Spectral Density shown for a wide range of sample resistances. The black lines indicate thermal resistor noise at 300 K, 4.2 K and 50 mK, respectively. Blue and red curves show the input noise of precision BJT and FET operational amplifiers, respectively.

Ideally, the noise curve of the used input amplifier thus stays below the noise curve of the sample resistance in **Figure 2**. At low sample resistances, the input amplifier noise is independent of the resistance and governed only by the voltage noise e_n. Most precision measurement devices use input amplifiers with a broadband voltage noise density of $e_n \lesssim 5\frac{\mathrm{nV}}{\sqrt{\mathrm{Hz}}}$. It is probably for the unit *Volts* why this number appears directly relevant for many electrical measurements and why many device makers place it on the forefront of their marketing sheets. Equally important however, amplifiers also have a small input current and an associated current noise i_n (expressed in $\frac{\mathrm{pA}}{\sqrt{\mathrm{Hz}}}$). This input current noise flows through the sample and drops a noise voltage of $i_n \cdot R$. When the sample resistance rises, this contribution rises linearly and becomes the dominant contribution for large resistances.

This general pattern of sample noise and amplifier noise has a number of consequences: Depending on the resistance of the specimen, the input amplifier should have either low voltage noise or low current noise. And even very good amplifiers will excel only for a limited range of sample resistances. Very high and very low sample resistances are always difficult to probe without adding electronic noise, especially when the sample is at cryogenic temperatures and produces very little Johnson noise.

Figure 3 expresses the same relationships in terms of the *Noise Power Spectral Density*. Resistors have a constant noise power density of $4 \cdot k_B \cdot T$. In contrast, the noise power contribution of electronics amplifiers varies depending on sample resistance. Typical precision amplifiers based on bipolar junction transistors (BJT) offer low voltage noise and good performance at resistances of around 100 – 1000 Ω. The minimum of the noise power spectral density is given by $2 \cdot e_n \cdot i_n$, while the optimum sample resistance is $\frac{e_n}{i_n}$. In comparison, FET amplifiers usually have higher voltage noise and thus perform worse at low resistances. On the other hand, FETs have low input current noise and the minimum noise power of precision FET amplifiers is orders of magnitude lower than that of BJT amplifiers. The noise power curve for FET amplifiers reaches deeply into the cryogenic territory of resistor temperatures.

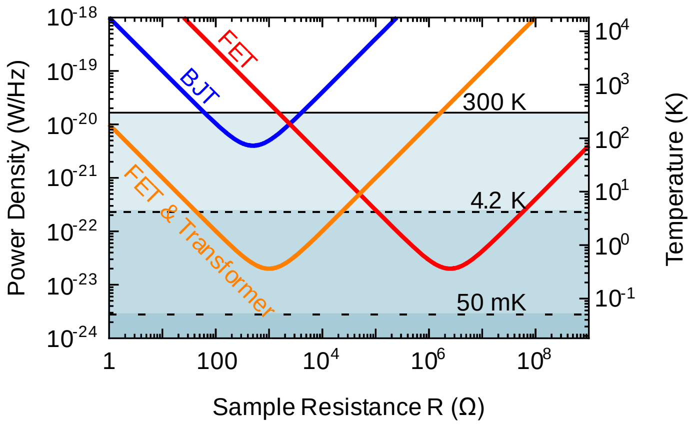

Figure 3: Noise Power Spectral Density shown for a wide range of sample resistances. The black lines indicate thermal resistor noise at 300 K, 4.2 K and 50 mK, respectively. Blue and red curves show the input noise power of precision BJT and FET operational amplifiers, respectively. The orange curve shows the effective input noise power for a transformer coupled FET amplifier.

So the big question is: *What is more important* – the low voltage noise of BJT amplifiers or the low noise power of FET amplifiers? In most precision measurement applications, the FET is the better choice, even for low sample resistances below 100 Ω. The reason is a simple trick. When using an input transformer with low resistance windings, the noise power curve can be shifted towards higher or lower resistances by the square of the winding ratio. The orange curve in **Figure 3** is obtained by adding a 1:50 transformer in front of the FET input.

2.2. Reducing 1/f noise and drift

Although 1/f noise and drift have different origins, they share several common aspects and will be treated together in the following. Loosely, 1/f noise can be thought of as a random drift around a mean state of the system. Both phenomena have in common that their noise power rises at least linearly towards lower frequencies and thus, both limit the base sensitivity of measurements (**Figure 1**). Conversely, both become irrelevant above the noise corner frequency f_C, where the noise power becomes frequency independent as broadband noise becomes dominant.

The two main sources of electronic 1/f noise are the measurement equipment (Section 2.2.2) and the sample (Section 2.2.3) including wiring and contacts (Section 2.2.4).

2.2.1. Experimental sources

The main source of drift is changing ambient conditions, in particular temperature. This refers not only to the temperature of the device under test, but also to that of the measurement device. Keeping the ambient conditions more stable alleviates problems associated with drift.

2.2.2. Electronic sources

The classical method to reject 1/f noise is by measuring and subtracting it from the input signal. To do so, *Chopper amplifiers* interleave each measurement of the device under test with an equal reference measurement of only the 1/f noise source. As a result, any 1/f noise below the chopping frequency can be subtracted from the compound signal.

In the simplest case, the reference measurement is a short between the terminals of the input amplifier, which includes the input offset voltage, its drift and input voltage 1/f noise. This functionality is offered in integrated operational amplifiers nowadays. The drawback is that half of the measurement time is lost to the determination of the 1/f noise. Therefore, such amplifiers have the penalty of $\sqrt{2}$ times higher broadband noise. One way around this penalty is switching the input terminals of the differential input stage, instead of measuring an input short. This allows measuring the input signal all the time, while still being able to reject the input offset, drift and 1/f noise.

Another drawback of chopper amplifiers is that they reject only 1/f noise caused at their own inputs. However, the concept generalizes other reference signals. The more noise sources can be made present in the reference signal, the more 1/f noise can be removed from the input signal. In particular, chopping the sample current driver allows the compensation of the driver 1/f noise and offset in addition to that of the input amplifier. Finally, alternating the sign of the driven current instead of merely chopping it, leads to the alternating current (AC) measurement concept. AC measurements condense most or all signal power at the alternating frequency. Therefore, a very narrow detection bandwidth can be used, which also greatly helps with immunity to parasitic signals such as EMI. Moreover, the AC frequency can be chosen to be substantially above the noise corner frequency (**Figure 1**). In this case, noise within in narrow detection bandwidth will be at a minimum.

2.2.3. Sample current noise

In an ideal measurement of the device under test, the dominant noise source is the sample itself, while the other noise sources are reduced as much as possible. When this is achieved, the final objective is thus, to make the sample generate a large signal and little noise.

One part of sample noise is a passive component due to the sample resistance. Broadband thermal Johnson noise will be always present with a constant voltage noise spectral density of $\sqrt{4 \cdot k_\mathrm{B} \cdot T \cdot R}$. The signal voltage, on the other hand, depends – often linearly – on the probe current flowing through the sample. As the thermal noise voltage is constant, the SNR rises when the probe current is increased. This relationship is shown in **Figure 4**: The resistance sensitivity reaches increasingly finer values when staying at the same demodulation time and increasing the probe current level.

If Johnson noise was the only component of sample noise, the sensitivity could be boundlessly increased by either using larger probe current or longer integration time. However, both of these approaches are impossible because the current flowing through the sample causes another noise contribution that is usually called *Current Noise* or *Excess Noise* and which has 1/f noise character. Moreover, excess noise voltage rises linearly with sample current. As a result, the SNR stops increasing when a certain current level is reached.

The main consequence of excess noise is that it places a fundamental lower bound to the precision of a resistance measurement. Trying to overcome this limit via longer integration times or higher sample current is futile. The characteristics of excess noise are summarized by the single dashed red line in **Figure 4**. The base sensitivity follows from any point on that line by $\frac{R_\mathrm{n}}{\sqrt{t_\mathrm{demod}}}$, where R_n is the resistance sensitivity. For the example of the 1.1 kΩ resistor, the sensitivity limit is approximately 250 µΩ.

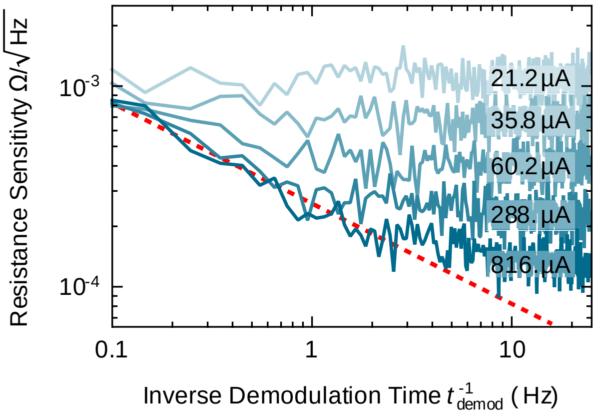

Figure 4: Attainable sensitivity of the resistance measurement of a standard 1.1 kΩ resistor at different AC probe current levels (values are AC amplitude). The dashed red line indicates the sample specific sensitivity limit, given by the noise index of the sample.

The *Noise Index* (NI) indicates the base sensitivity in relation with the base resistance value. The definition of the noise index is in terms of RMS noise voltage per spectral decade and per applied voltage. This leads to a dimensionless number – usually expressed in dB – that is not very intuitive for the sake of resistance measurements. A more helpful representation might be the following expression for the precision limit:

$$\frac{\Delta R}{R} > \frac{10^{\frac{\mathrm{NI}}{20\,\mathrm{dB}}-6}}{\sqrt{\log 10}} \approx 6.6 \cdot 10^{-7} \cdot 10^{\frac{\mathrm{NI}}{20\,\mathrm{dB}}}$$

The equation leads to a precision limit of **$6.6 \cdot 10^{-7}$** of the base value for a Noise Index (NI) of 0 dB. While this number does not appear to be very low in relation with the capabilities of state-of-the-art measurement devices, the material Noise Index can be much lower (e.g. -40 dB) for well conducting metallic specimens. Measuring such samples down to their base sensitivity requires great care at all stages of the experiment.

2.2.4. Contact noise

While the sample properties are usually given, contact issues are avoidable in most cases. The sample current flows through the contacts making them, too, a potential source of excess noise. Contacts should be ohmic, large, thermally stable and made with good conductors that have a low material noise index such as metals. If the contact surface is covered in a natural or protective coating (e.g. an oxide) an effort should be made to remove or penetrate this coating. Many metallic surfaces form tough insulating oxides, e.g. Aluminum and Tantalum. If a large area contact is placed over such an oxide, the contact resistance will be low, but the noise index of the connection will be high regardless. Therefore, such a contact will create a lot of 1/f noise when carrying current.

2.3. Output noise

The driver noise has to be considered, when attempting low noise resistance measurements. Any output voltage and current noise of the driver will affect the sample and will be picked up again partially by the measurement device. Drivers usually consist of a voltage reference, a DAC and buffer amplifiers and have typical output voltage noise densities of around $e_{\mathrm{n,out}} \approx 100 \frac{\mathrm{nV}}{\sqrt{\mathrm{Hz}}}$, which is many times larger than the input noise of most precision equipment.

In a typical 4-wire Kelvin type measurement of longitudinal sample resistance, the driver noise drops completely over the device under test, and appears again on the probing side of the measurement. However, several kinds of 4-wire resistance measurement geometries such as Wheatstone Bridges or Hall Crosses allow strongly suppressing the driver noise. These solutions convert a large fraction of the driver noise from differential mode to common mode with respect to the probe terminals. The large common mode rejection ratio of typical differential amplifiers then removes the driver noise.

3. Examples and Device Choices

Noise sources and as well as their remedies are diverse and situational as discussed in the previous sections. In the following, three specific application examples will be analyzed and the most suitable measurement procedure will be determined for each case. The examples become increasingly more specialized. So most considerations for one example also apply to the next.

3.1. Precision long-term measurement of longitudinal resistance

This scenario is often encountered in both science and sensorics. The device under test can be e.g. a resistive thin film device with a typical base resistance of 100 – 1000 Ω. The goal is detecting small resistance changes during the course of the experiment or field operation.

Parameters
Start by spotting the operation conditions (sample resistance and temperature) in **Figure 3**. A BJT input amplifier is generally superior to a FET amplifier to measure low resistance samples as in this example. However, if the operation conditions include lower temperatures such as in Helium cryostats or in space applications, the noise power of BJT inputs is still too high and a transformer coupled FET input should be used instead.

For long-term measurements, AC offers great benefits such as higher stability and lower 1/f noise than DC measurements. The excitation frequencies of the AC measurement should be at least around 1 kHz, to exclude most electronic 1/f noise sources from contributing significant noise. On the other hand, the AC frequency should not be needlessly higher, because a) higher frequencies increase parasitic conductivity through stray capacitances and b) the detrimental effect of sample clock jitter on SNR increases linearly at higher frequencies.

To pick the exact AC frequency, the excitation and demodulation frequencies should be as far away from EMI source frequencies as possible. In this example, the excitation and demodulation frequencies are the same, which simplifies the frequency selection. If there are no strong sources of narrowband EMI near the experiment, it is usually safe to assume that the mains harmonics are the dominant EMI frequencies (**Figure 1**). For example, an AC frequency of 1075 Hz (in between the mains harmonics at 1050 Hz and 1100 Hz) would thus be a good choice granting the maximum $f_{\mathrm{sep}} = 25$ Hz of frequency separation to the offending signals. When measuring several independent devices in parallel, different frequencies should be used for each measurement to minimize crosstalk.

The demodulation time per data point t_{demod} should be at least $\frac{4}{f_{\text{sep}}} = 160$ ms, because at shorter times the EMI signals at 25 Hz away cannot be fully distinguished from the target signal. Making integration times even longer can reduce the noise in each of the data points. However, this is usually a bad tradeoff. Measuring instead more and slightly noisier data points yields the same level of accumulated precision and allows the signal processing stage to perform more versatile filtering.

In this example, the expected range of resistance change is much smaller than the base resistance of the sample. Therefore, it is useful to place the sample in a Wheatstone bridge to reject most of the driver noise.

Finally, an optimum sample current level can be chosen according to **Figure 4**. For optimum SNR, the sample current would be so high that the base sensitivity is reached at the inverse of the chosen demodulation time. Setting current any higher will yield no benefit. Lower current can be necessary if the optimum current level turns out to be prohibitively high. The base sensitivity could still be reached at reduced current levels by averaging several data points until their compound integration time reaches the base sensitivity line.

Devices

The simplest solution is a lock-in amplifier in combination with a Wheatstone bridge. Many lock-in amplifiers offer at least a 5-digit (100 dB) continuous dynamic range. This is not to be confused with the higher "dynamic reserve" rating of some lock-in amplifiers, which is inconsequential for this measurement application. When the Wheatstone bridge is tuned to suppress the base resistance to a 100^{th}, one can resolve resistance steps down to $\sim 10^{-7}$, which is sufficient for most applications.

A more advanced solution also addresses gain drift: For long-term measurements, a low gain drift of the measurement device is important to detect slowly changing signals. Typical Lock-in amplifiers do not have excellent gain drift ratings and could thus show signal changes even when the specimen resistance does not actually change. This can be solved by carefully adding a switching unit that interleaves sample measurements with reference measurements of a precision impedance reference. This approach does not combine well with lock-in amplifiers due to their infinite impulse response low-pass filters used for demodulation.

An integrated solution that offers > 8-digit AC measurements and has integrated support for impedance reference measurements is available. The *Tensormeter RTM1* offers ultra-low gain drift of around 0.1 ppm/K in ratiometric mode with suitable impedance references. Due to the used demodulation process, reference and specimen measurement can be tightly interleaved with minimum dead time.

3.2. High-throughput production monitoring

In contrast to precision measurements on a single specimen, in high-throughput monitoring the time spent per device is usually only a few milliseconds. As such, it is central to employ a sophisticated stepping and contacting procedure to maximize the actually available measurement time. Despite the short measurement time per device of e.g. 5 ms, it is important to understand that production monitoring requires stability over the entire monitoring time, which can be hours and beyond.

Therefore, high-throughput monitoring is actually a special case of a precision long-term measurement (Section 3.1) with the additional constraint of a very short demodulation time.

Parameters

As with the previous example, a Wheatstone bridge can help to reject driver noise, especially when all samples are expected to have very similar properties.

The short available demodulation time prevents an AC measurement from reaching sufficient frequency definition to be distinguishable from mains EMI. Therefore, shielding of the test stage against both electric and magnetic fields is advised. Even then, an AC measurement still offers an important advantage. As shown in **Figure 1**, measuring at a high frequency reduces the noise impact from 1/f noise. Mains EMI is also reduced at kHz frequencies, while switch-mode power supply EMI only becomes a concern above approximately 20 kHz. Usual frequencies in high-throughput monitoring should thus be even higher than in long-term precision measurements – usually several kHz. This demands an impedance controlled test stage – with controlled stray capacitances and inductances and fixed cable layout. The better this impedance control, the higher the test resistance that can be reliable probed with the AC measurement. If reaching a high precision for large resistors (e.g. Megaohms) is mandatory, it will still be necessary to reduce the AC frequency and the test throughput.

Besides high AC frequency, the second important ingredient is to use the highest possible probe current. By measuring the sample noise spectra at different current levels (**Figure 4**), it is easy to identify the necessary current level to reach base sensitivity at a frequency of $\frac{1}{t_{\mathrm{demod}}}$. If the target precision is high and the available demodulation time is small, the necessary current level can be prohibitively high causing excessive Joule heating. In this case, the demodulation time will have to be extended to reach the target precision, which in turn reduces test throughput.

Special aspects

Finally, an important consideration is the AC demodulation process. This must use only data from the time period, during which the sample was actually contacted. Therefore, infinite impulse response filters must not be used in the signal chain. Such filters often occur in Lock-in amplifiers with a “time constant” setting. Moreover, a windowing function should be applied to the valid contact period. This sharpens the AC demodulation and suppresses spectral leakage and crosstalk with spurious AC signals.

Devices

The simplest approach for this application is a stable low noise source unit, a 7-digit voltmeter and a smart use of the dead time imposed by sample changes. When shortcutting the voltmeter during the dead time, the input offset drift of the voltmeter can be nulled. The source unit drift becomes the limiting stability issue, so the source unit should be a very stable type or also modulated. Driver noise can be tackled with a Wheatstone bridge made with low-drift resistors, if it becomes an issue.

In order to compensate for gain drift of the measurement devices as well, sample measurements must be interleaved with measurements of an impedance reference, leaving only a third of the time for actual device tests.

An integrated solution that also offers the additional benefits of AC measurements (lower noise, complex impedance information) is available. The *Tensormeter RTM1* needs only a single reference measurement in AC ratiometric mode, enhancing by 50% the effective measurement time over the combo of source unit and voltmeter.

3.3. Tensor measurements

Measuring several components of the Resistance Tensor is an important task in physics and sensorics, e.g. to distinguish real and parasitic signal changes. The simplest case of the Tensor measurement is the *Zero-Offset Hall* thin film measurement configuration, which yields independent values for the longitudinal and transverse (Hall) resistances of arbitrarily shaped thin films. To achieve this, one has to use at least two noncollinear current flow directions through the device under test. This alternation between these current directions is practically realized by switching the sample contacts, so different contacts carry the current and probe voltages at different times. The contact switch matrix also makes other related non-stationary measurement protocols possible, such as van-der-Pauw resistivity measurements.

Parameters

Tensor measurements share many aspects with the high-throughput monitoring application. It is desirable to alternate the current directions in quick succession, to minimize drift of the sample properties. The contact switching frequency follows from the used demodulation time as $f_{\mathrm{Sw}} = \frac{1}{t_{\mathrm{demod}}}$. This reduces the permissible demodulation time – with the same consequences as discussed for high-throughput monitoring. If the sample properties do not change fast, a longer demodulation time of at least $t_{\mathrm{demod}} = \frac{4}{f_{\mathrm{sep}}} = 160$ ms should be used to provide much better immunity to mains EMI as explained in Section 3.1.

While the duration of a single contacting configuration is short, the current still flows through the same sample every time. The thermal dissipation budget is thus much lower than in high-throughput monitoring and the permissible current is similar to long-term precision measurements. Such limited current levels imply longer averaging times to reach base sensitivity (**Figure 4**). On the hand, lower current also removes the need for very short demodulation times.

The necessary switch matrix should switch fast, reliable and very often, which excludes mechanical switches. However, electronic switches introduce parasitic capacitances, which limit the useful signal frequency range to 10s of kHz. As a result, usual AC frequencies in Tensor measurements are in the low kHz range – ideally in the gaps between mains harmonics.

Special aspects

In contrast to both previous examples, both the drive resistances and the 4-wire resistance can change substantially between successive contacting schemes in a Tensor measurement. The detection ranges of the instrumentation should be such that they accommodate the largest signals of all of the contacting schemes. The varying resistances can also lead to varying probe current levels. However, the changing current levels pose no problem for the analysis when calculating the resistances separately for each contacting scheme. Ideally, the power input from resistive heating should be roughly the same in all contacting schemes. This helps keep the sample in thermal equilibrium while switching current directions.

Another implication of the varying resistances between contacting schemes is that Wheatstone bridges are more difficult to implement than in measurements with stationary contacting. However, even when not using a Wheatstone bridge, this does not automatically mean that driver noise will contribute a lot to the detected signal. In a Zero-Offset Hall measurement, most of the driver noise below f_{Sw} is rejected from the transverse channel. This brings driver noise to a very low level where it is usually overshadowed by other noise contributions. However, a larger fraction of the driver noise enters the longitudinal resistance reading – typically between a 10^{th} and a 1000^{th} of the full driver noise. As a result, driver noise, can be the limiting factor for SNR in the longitudinal resistance channel and low noise drivers with a SNR $\gtrsim$ 140 dB should be used.

Devices

The advantages of AC measurements are too large in this application to recommend a combo of DC source unit and voltmeter. On the hand, lock-in amplifiers cannot be used as they would require significant dead time after switching due to their demodulation process.

A sensible solution involves an AC source, a reliable matrix switch unit and a sophisticated signal processing that maximizes the signal usage of the chopped signal. The *Tensormeter RTM1* is an available solution with out-of-the-box capabilities for exactly such measurements.

Author Profile

Dr. Tobias Kosub works as a senior scientist at Helmholtz-Zentrum Dresden-Rossendorf (Germany), and is CTO of Tensor Instruments (a division of HZDR Innovation GmbH). His research interests lie in the field of transport phenomena in thin films. Dr. Kosub holds a Ph.D. in physics (on thin-film spintronics) and has 10 years of experience in the development of resistance measuring devices.

About Tensor Instruments

Tensor Instruments is devoted to ending the stagnation in resistance measurement methods. Although measurement devices have become much finer and faster in the recent decades, the general methods and thought of resistance metrology has remained largely unchanged, namely 4-wire resistance measurements using stationary contacting. We are convinced that dynamic contacting schemes have much more to offer than only van-der-Pauw measurements. The Tensormeter product line offers state-of-the-art DC and AC resistance measurements with integrated switch matrices in an easy-to-use package that sets you ready to go into uncharted measurement territory. Moreover, our lab automation hardware and software allow creating sophisticated automated experiments in no time.