High Resolution
Resistance Measurements

Bachelor’s Thesis

Tobias Meier

Advisor: Dr. Sebastian T. B. Goennenwein

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Chapter 1

Introduction

In 1964, Arno Penzias and Robert Wilson started experimenting with a highly sensitive horn antenna [1] for radio astronomy. In their experiments, they found a steady, persisting noise, which was independent of the daytime and the direction of the antenna [2]. When Penzias and Wilson checked their equipment, they found pigeons nesting in the antenna. But even after expelling them from the antenna and removing their droppings, the noise persisted in their measurements.

A possible explanation for this phenomenon was given at the same time by Robert Dicke, Jim Peebles and David Wilkinson [3], who were preparing to measure radiation in the microwave range. They stated that the big bang sent out not only matter but also radiation, which should be observable in the microwave-range due to a massive redshift. The noise measured by Penzias and Wilson fully met the characteristics (isotropic, unpolarized radiation) postulated by Dicke, Peebles and Wilkinson. This way it became clear that the accidentally measured noise had exceptional significance for astrophysics. For their discovery of the cosmic microwave background, Penzias and Wilson received the nobel prize for physics in 1978.

This historical example illustrates what an important role noise can play in physics. Especially when measuring a low level effect, one should always consider the noise level in the experiment and how it affects the measurement. If the occurring noise level is higher than the magnitude of the effect, there is no point in trying to measure the effect. Even nowadays, many experiments and also commercial measurement electronics are limited by noise.

Very recently, a novel magnetoresistance called the spin Hall magnetoresistance (SMR) was covered in thin platinum layers in close vicinity to a ferromagnetic insulator [4][5]. This effect, resulting in a $10^{-3}$ change in resistance in nm-thick Pt layers, was ascribed to a spin sensitive transport across the interface. However, a debate on the origin of this effect is still seething [6][7] and control experiments on other material combinations are necessary to clarify this issue. A very promising metal to validate the SMR theory is gold, but the prediction of the SMR in Au samples leads to very small ($10^{-7}$) changes in resistance. Therefore, the objective of this Bachelor thesis is to compare different measurement methods to detect low changes in resistance (respectively voltage).
This thesis splits into 4 chapters. In Chapter 2 a short introduction to the theoretical background of the spin Hall magnetoresistance and its competing effect, the proximity magnetoresistance, are given. Chapter 3 contains the experimental techniques and setups as well as a detailed discussion of the theoretical limits in comparison to the results of the measurements. A conclusion of the results as well as suggestions for further proceeding in chapter 4 rounds off this thesis.
Chapter 2

Theoretical background

Magnetoresistance is the property of a solid body to change the value of its electrical resistance when applying an external magnetic field [8]. An example is the anisotropic magnetoresistance (AMR) observed by W. Thomson in 1856 [9], where the electrical resistance of a ferromagnet depends on the angle between the direction of the electric current and the magnetization direction. When applying an external magnetic field $\mu_0 H > M_{\text{sat}}$, the magnetization direction in the ferromagnet adjusts parallel to the external field and the electrical resistance changes. Today, the AMR-effect is commonly used in a lot of industrial or automotive applications [10].

In many recent experiments, a new kind magnetoresistance has been observed in a nonmagnetic metal (NM) layer attached to a ferromagnetic insulator (FMI) [4][5][7]. To explain the origin of this magnetoresistance, two different models have been developed, the spin Hall magnetoresistance (SMR) based on spin current flow across the FMI|NM interface [4] and a magnetic proximity effect based model, which assumes a static spin polarization in the NM at the interface [7]. One qualitative test which of these two models is more appropriate thus would be to study FMI|NM heterostructures in which no static magnetic proximity would occur. For a material far off the Stoner-criterion (cf. Sec. 2.4), only the SMR-theory predicts an effect and therefore measuring this magnetoresistance would support the SMR-theory. This was the starting point of this thesis.

This chapter gives a brief overview of the mechanisms behind the SMR, starting with the concept of charge and spin currents in Sec. 2.1. The generation of spin currents due to the so called spin Hall effect as well as some basic spin transport properties are introduced in Sec. 2.2 followed by a discussion of the spin Hall magnetoresistance itself in Sec. 2.3. A short description of the proximity effect based model in Sec. 2.4 rounds off this chapter.

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1The term NM describes paramagnetic metals, so these are conductors without a permanent magnetization. Examples are Pt, Au or Pd.

2FMIs are insulating materials that possess a permanent magnetization like a (e.g. ferromagnet). A commonly used example is YIG (Yttrium Iron Garnet, $Y_3Fe_5O_{12}$).
2.1 Spin currents

A charge current consists of uniformly moving electrons. However, electrons possess an additional degree of freedom, their spin. The projection of the electron spin angular momentum parallel to a certain quantization axis is called spin-up, the projection anti-parallel spin-down. Both spin-up (↑) and spin-down (↓) electrons contribute equally to a charge current. Therefore the charge current density $J_c$ can be written as the sum of the charge current densities generated by spin-up and spin-down electrons, as already suggested by N. Mott:

$$J_c = J_↑ + J_↓$$

(2.1)

In contrast to the charge current density, the spin current density can be defined as a net spin momentum $J_↑ - J_↓ \neq 0$ carried along the transportation direction [12]:

$$J_s = -\frac{\hbar}{2e} (J_↑ - J_↓),$$

(2.2)

where $\hbar$ is the reduced Planck constant and $e$ the absolute value of the electron charge. With this model, one can divide electric transport in three main regimes as shown in Fig. 2.1. If the total current density consists of the same number of spin-up and spin-down electrons flowing in the same direction, the expression for the spin current density 2.2 equals zero. This means a pure charge current is generated. An unsymmetrical distribution of spin-states leads to non-zero values for both charge and spin current density. In this case, one calls the total current density spin-polarized. The most interesting case for investigating spin transport is the following: If spin-up electrons flow along one direction and the same number of spin-down electrons along the opposite direction, the effective charge current will become zero, while the spin current density $J_s$ and therefore a pure spin current is generated.

![Figure 2.1](image)

Figure 2.1: Different forms of current: (a) Pure charge current (b) Spin-polarized charge current (c) Pure spin-current; Illustration from [12].

Spin-polarized currents and pure spin currents can be generated in paramagnetic metals by different effects, which stem from spin-orbit interaction. Two extrinsic mechanisms called skew scattering [13] and side-jump [14] are caused by spin-dependent scattering of electrons at impurities with a charge. In addition, there is

$^3$The spin quantum number of an electron is $s = 1/2$
an intrinsic effect called *Berry-phase mechanism* [15]. All of these convert a charge current into a perpendicular spin current. A more detailed description, which is beyond the scope of this thesis, can be found in e.g. [16].

### 2.2 Spin transport

The skew scattering and side-jump mechanisms mentioned in Sec. 2.1 can cause macroscopic effects, predicted by Dyakonov and Perel [17] as well as Hirsch [18]. If one assumes a charge current flowing through a nonmagnetic metal (NM), a perpendicular spin current $J_s$ is generated because of spin-orbit interaction. This process shown in Fig. 2.1(a) is called the *spin Hall effect (SHE)* in analogy to the ordinary Hall effect (OHE), where an external magnetic field induces a perpendicular charge current$^4$. As the spin current can not exit the NM, the two different spin states accumulate on each side of the sample and thus a gradient in the spin-chemical potential $\mu_{\text{chem}}$ emerges. As a consequence, a spin-diffusion current $J_{\text{diff}}$ flowing anti-parallel to the initial spin current builds up. The whole system reaches an equilibrium when both spin currents compensate each other. Based on the same mechanisms, it is also possible to generate a perpendicular charge current out of a spin current by the *inverse spin Hall effect (ISHE)* as depicted in Fig. 2.1(b). The two effects can be described by the following equations [12]:

$$
J_{\text{SH}}^s = \alpha_{\text{SH}} \left( -\frac{\hbar}{2e} \right) J_c \times s, \quad J_{\text{ISH}}^c = \alpha_{\text{SH}} \left( -\frac{2e}{\hbar} \right) J_s \times s
$$

Here, $s$ is the spin vector and $\alpha_{\text{SH}}$ is the so called spin Hall angle, which describes how effectively charge and spin currents are converted. It is defined as the ratio of spin conductivity $\sigma_s$ and electrical conductivity $\sigma_{\text{el}}$ [19]:

$$
\alpha_{\text{SH}} = \frac{\sigma_s}{\sigma_{\text{el}}}
$$

![Figure 2.2](image)

**Figure 2.2:** (i) The spin Hall effect converts a charge current into a spin current. (ii) The inverse spin Hall effect converts a spin current into a charge current. Illustration taken from [12].

The value of the spin Hall angle depends on the material used. As the origin of the SHE is spin-orbit interaction, elements with strong spin-orbit coupling have a

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$^4$notwithstanding that the SHE occurs in the absence of a magnetic field
larger spin Hall angle than ones with weak spin-orbit coupling. Example values are $\alpha_{\text{SH}} = 0.11$ for Pt \cite{4} or $\alpha_{\text{SH}} = 0.0016$ for Au \cite{20}.

Apart from the spin Hall angle, there are two additional important properties that describe spin transport phenomena. The *spin diffusion length* $\lambda$ is the mean distance a non-equilibrium spin state can propagate before it is re-oriented. Typical values are $\lambda = 1.5\text{nm}$ for Pt \cite{4} or $\lambda = 35\text{nm}$ for Au \cite{21}. In a bilayer consisting of a ferromagnetic insulator (FMI) and a NM, the *spin mixing interface conductance* $G_{\uparrow\downarrow}$ characterizes how effectively a spin current can be transferred across the interface between those two different materials. Although $G_{\uparrow\downarrow}$ is a complex quantity, interface-effects like spin-pumping \cite{22} \cite{23} and the longitudinal SMR are characterized by the real part $G_r$ of $G_{\uparrow\downarrow}$. For a YIG|Au-bilayer, a value of $G_r = 2 \times 10^{14}\Omega^{-1}\text{m}^{-2}$ \cite{24} has been obtained from spin pumping measurements, for a YIG|Pt-bilayer a value of $G_r = 4 \times 10^{14}\Omega^{-1}\text{m}^{-2}$ has been determined in \cite{4} with spin Hall magnetoresistance measurements.

### 2.3 Spin Hall magnetoresistance

Now we will take a closer look at the bilayer-system of a nonmagnetic metal and a ferromagnetic insulator introduced above (cf. Fig. 2.3). Applying a charge current in the NM generates a spin current perpendicular to the charge current via the SHE. Depending on the orientation of the magnetization $M$ of the FMI, there are two different situations. If $M$ is parallel to the spin orientation $s$ of the spin current, it is not possible to transfer angular momentum across the interface, so the spin current is blocked. The situation is then effectively the same as in the considerations with only the NM in Sec. 2.2. In the second case, $M \perp s$ (or more generally, $M$ not exactly parallel to $s$), the spin current can transfer spin angular momentum to the magnetization of the FMI via spin transfer torque \cite{25} \cite{26}. As this process is accompanied by transferring energy, the resistance in the NM rises. The resulting additional resistance is called the *spin Hall magnetoresistance* (SMR).

In \cite{25}, Chen *et al.* developed a theoretical model of the SMR-effect by evaluating the spin diffusion equation for FMI|NM bilayers. It predicts the following relations for the transverse resistivity $\rho_{\text{trans}}$ (the resistivity in the direction of spin polarization $t$ as indicated in 2.3) and of the longitudinal resistivity $\rho_{\text{long}}$ (the resistivity in the current direction $j$ as indicated in 2.3):

$$\rho_{\text{long}} = \rho_0 + \rho_1 m^2_t$$ (2.5)

$$\rho_{\text{trans}} = \rho_2 m_n + \rho_3 m_t m_t$$ (2.6)

In this equations, $\rho_0$ is the magnetization-independent part of the specific resistance, $\rho_1 = -\rho_3$ stem from the SMR-effect and $\rho_2$ is a Hall-effect related resistivity. The vectors $m_j$, $m_t$ and $m_n$ represent the magnetization orientation $m = \frac{M}{M_{\text{sat}}}$ projected in the direction of current ($j$), in the transversal direction ($t$) and in the normal direction ($n$)\footnote{$M_{\text{sat}}$ is the saturation magnetization of the FMI}.
Proximity based magnetoresistance

Figure 2.3: Sketch of the spin Hall magnetoresistance in a NM|FMI-bilayer for two configurations of the magnetization $\mathbf{M}$ and the spin direction $\mathbf{s}$: (a) $\mathbf{M} \parallel \mathbf{s}$: The spin current can not cross the NM|FMI interface, which leads to spin accumulation on both sides of the NM and a spin diffusion current $J_{\text{diff}}$. In the resulting equilibrium state, there is no additional resistance contribution in the NM. (b) $\mathbf{M} \perp \mathbf{s}$: The spin current can transfer spin angular momentum to the FMI (spin transfer torque), raising the resistance in the NM. Illustration taken from [27].

Furthermore, the relative change in the longitudinal resistance is given by [4]:

$$\frac{-\rho_1}{\rho_0} = \frac{\alpha_{\text{SH}}^2 (2\lambda_{\text{NM}}^2 \rho_{\text{NM}}) (t_{\text{NM}})^{-1} G_r \tanh \left( \frac{t_{\text{NM}}}{2\lambda_{\text{NM}}} \right)}{1 + 2\lambda_{\text{NM}} \rho_{\text{NM}} G_r \coth \left( \frac{t_{\text{NM}}}{\lambda_{\text{NM}}} \right)}$$

(2.7)

Here $\alpha_{\text{SH}}$ is the spin Hall angle, $\lambda_{\text{NM}}$ the spin-diffusion length in the nonmagnetic metal, $\rho_{\text{NM}}$ its specific resistance, $t_{\text{NM}}$ the layer thickness of the nonmagnetic metal and $G_r$ the real part of the spin mixing interface conductance. This effect was observed in YIG|Pt [4] with a maximum amplitude $\frac{-\rho_1}{\rho_0} = 1.6 \times 10^{-3}$. The SMR-magnitude strongly depends on the NM-layer thickness $t_{\text{NM}}$, because the spin current can not reach the FMI|NM interface if $t_{\text{NM}} \gg \lambda_{\text{NM}}$. The maximum of the SMR-effect is at $t_{\text{NM}} \approx 2\lambda_{\text{NM}}$. An example for the thickness dependency of the SMR is shown in Fig. 2.5 in Sec. 2.4 for a YIG|Au-bilayer.

2.4 Proximity based magnetoresistance

In [29], Huang et al. presented a different approach to explain the change in resistance in a NM/FMI-hybrid depending on an external magnetic field. It states that the magnetization in the FMI induces magnetic moments in the NM in close proximity to the FMI|NM-interface (cf. [24]).

This results in an anisotropic character of the magnetoresistance in an initially (e.g. pure) non-ferromagnetic normal metal. Different elements or alloys show a varying affinity to orient their magnetic moments, depending on their electron configuration.
Figure 2.4: Sketch of a hypothetic proximity at the FMI[NM interface. The green spheres represent electrons in the FMI, the gray spheres are electrons in the NM and the yellow spheres are electrons in the NM in close proximity to the FMI[NM interface. In (a), the NM and the FMI are not in contact, so the magnetic moments in the FMI are oriented, whereas the magnetic moments in the NM show no specific orientation. When bringing the two materials together as shown in (b), the FMI induces oriented magnetic moments in the NM. This effect is limited to the close proximity of the FMI[NM-interface, so only the electrons in yellow color are affected by this mechanism.

The so called Stoner-criterion describes whether an electron gas can minimize its energy by orientating spins parallel \[ \frac{1}{2} UD(E_F) > 1 \] (2.8)

Here, \( U \) is the characteristic energy density and \( D(E_F) \) the density of states at the Fermi-level. Pt is close to the Stoner-criterion, as evidenced experimentally by Wilhelm et al., who showed the presence of induced magnetic moments in Ni[Pt-layers with XMCD\(^6\) measurements in [32]. XMCD-measurements have also been done for a YIG|Pt-bilayer with contrary results: In [6], Geprägs et al. found no evidence for induced moments in Pt, while Lu et al. state the presence of magnetic moments in [7]. This was the starting point of this thesis. If we choose a different NM like Au, which is far off the Stoner-criterion, one would not expect any induced magnetic moments in the proximity of YIG|Au-bilayers. In contrast to that, the SMR-model from Sec. 2.3 does predict a magnetoresistive effect for these samples, since Au has a small but finite spin-orbit coupling. Investigating the magnetoresistance of YIG|Au-bilayers can thus give a clue which of the above two models is more appropriate to describe the physics behind the observations in FMI[NM-hybrids.

To estimate the longitudinal SMR-magnitude in a YIG|Au-bilayer, we use Eq. 2.5 from Sec. 2.3. Figure 2.5 shows the thickness dependence of the longitudinal SMR using the literature values \( \alpha_{SH} = 0.0016 \) [20], \( \lambda = 35\text{nm} \) [21], \( G_r = 2 \times 10^{14}\Omega^{-1}\text{m}^{-2} \) [24] and \( \rho = 3\mu\text{Ωcm} \) [20]. Even if choosing the optimal thickness for the sample, the SMR magnitude \( \rho_1/\rho_0 \) is in the range of \( 10^{-7} \), which is about three to four orders of magnitude smaller than the value for a YIG|Pt-bilayer [4]. Thus it is much

\(^6\)XMCD is the abbreviation for X-ray magnetic circular dichroism, for more information see [31]
more difficult to measure the SMR in YIG|Au-hybrids. A detailed discussion about measurement setups and associated problems is therefore given in Ch. 3 to determine if it is even possible to measure such a small effect.

Figure 2.5: The longitudinal SMR-amplitude \(-\rho_1/\rho_0\) is plotted against the NM thickness \(t\) using eq. 2.7 for a YIG|Au-bilayer. The maximal amplitude is \(2.2 \times 10^{-7}\) at a layer thickness of \(t = 78\text{nm}\).
Chapter 3

Low level measurements

Every electrical measurement has a certain level of noise background. If one wants to measure a very small effect like the expected SMR in YIG|Au-hybrids, it is important to determine the resolution limit given by intrinsic noise sources and keep the noise background as low as possible. Therefore, the different noise sources and their influence are discussed at the beginning of this chapter (cf. Sec. 3.1). In addition to these effects, it has to be considered that every device used for measuring acts as an additional noise source. The main part of the chapter is devoted to a description of the different measurement techniques and measurement electronics used will be followed by a discussion of internal noise sources in each device. The section will end with a summary of the resolution limits for each method followed by a discussion of the pros and contras of each setup.

3.1 Noise sources

Mathematically, noise can be interpreted as a time dependent stochastic function restricted to a certain observation interval [27]. This can be physically observed in electronic measurements as undesired, random fluctuations of an electronic signal. The noise sources discussed in this work are assumed to be uncorrelated, which means the total noise from different contributions can be determined by incoherent addition, \( \Delta V_{\text{tot}} = \sqrt{\Delta V_1^2 + \Delta V_2^2 + \ldots} \). In addition to that, every type of noise is treated as Gaussian noise. This means that the probability density function of the noise is equal to a Gaussian distribution. In this statistical regime, the relation \( \Delta V(\text{pp}) = 6.6 \cdot \Delta V(\text{RMS}) \) is commonly used to convert RMS-noise into peak-to-peak noise and vice versa [33]. Fig. 3.1 schematically shows this for a voltage measurement. We define the peak to peak noise level shown in red as the difference between the highest and lowest values during the measurement. The RMS noise level drawn in blue is then by a factor 6.6 lower.
3.1.1 Johnson-noise

Johnson-noise (also called Nyquist- or thermal noise) \([34, 35]\) has its origin in thermal agitations of charge carriers and can be observed in every resistor. Because its spectral power density is constant over the whole frequency spectrum, it is often also called white noise. Johnson-noise (RMS) can be calculated by

\[
\Delta V_{\text{Johnson}} = \sqrt{4k_BTR\Delta f},
\]

where \(k_B\) is the Boltzmann-constant, \(R\) the resistance value, \(T\) the resistor’s temperature and \(\Delta f\) the bandwidth.

3.1.2 Shot-noise

Shot-noise \([36]\) originates from the fact that charge is transported in single quanta. If one considers a constant charge current as a statistical process of many electrons, there is a mean number of particles per time propagating through a sample. But there are always deviations from this value which sum up to changes in the macroscopic current. Shot-noise (RMS) can be calculated by

\[
\Delta I_{\text{Shot}} = \sqrt{2eI\Delta f},
\]

where \(e\) is the absolute value of the electron’s charge, \(I\) the applied current and \(\Delta f\) again the bandwidth.
3.1.3 Flicker-noise

Flicker-noise \(^{37}\) is also called 1/f-noise or pink noise, as its spectral power density is not constant but gets lower for higher frequencies. It is caused by a big variety of effects, e.g. impurities in a conductive channel or generation or recombination noise in a transistor \(^{38}\). 1/f-noise determines the input noise in devices for low frequencies and therefore has to be taken into account when measuring in this regime. It can be neglected for higher frequencies, as there is a certain corner frequency at which white noise becomes dominant over 1/f-noise.

3.2 Experimental setups and results

In this section, different approaches to detect very small resistance changes (as those expected in SMR in YIG\(|\)Au\) in resistance are presented.

3.2.1 Boundary conditions

This subsection gives a short overview of the used samples as well as the conditions and conventions for all measurements. Three YIG\(|\)Au-samples have been prepared at the WMI, which were patterned into Hall-bars. The three samples have different Au-layer thicknesses \(t_1 \approx 6\text{nm}, t_2 \approx 15\text{nm}, t_3 \approx 27\text{nm}\)\(^{1}\). Thus, the Hall-bar resistance values of these samples vary between \(10\Omega \leq R \leq 500\Omega\). However, for determining the resolution limit of measurement setups, we do not need these particular samples. The most important properties for noise measurements are the resistance of a sample and its temperature coefficient. Therefore, we use carbon- or metal film resistors whose resistances are of the same magnitude as the resistances of the YIG\(|\)Au samples to simulate their behavior. These resistors have a specified temperature coefficient of 50ppm/K (metal film) and 600ppm/K (carbon film) and will be used as "dummy"-resistors instead of the YIG\(|\)Au-bilayer samples in all the presented measurements. For each measurement, the "dummy"-resistor was soldered onto a chip-carrier in a four-wire connection scheme to minimize the influence of lead resistances. As we want to measure at room temperature, the resistors on their respective chip carrier were mounted onto a magnet cryostat dipstick\(^{2}\) placed in a measurement-rack without any cooling or heating. To detect the temperature fluctuations in the lab, a cernox temperature sensor was connected to a LakeShore LS340 temperature controller placed alongside the sample.

For our experiments the following conventions are used: As measurement time \(\Delta t_{\text{meas}}\), we take the observation time that is quoted by the manufacturers of the measurement electronics in its specifications. If no time is quoted, the measurement

\(^{1}\)The sample with 15nm has been prepared by Felix Schade, whereas the other two samples were grown by Sibylle Meyer

\(^{2}\)An example of a dipstick is shown in \[39\]
Low level measurements

time is set to $\Delta t_{\text{meas}} \simeq 30\text{min}$. The resolution limit of each method is determined by $\Delta V/V_{\text{sample}}$, where $\Delta V$ is the peak-to-peak noise in the measurement and $V_{\text{sample}}$ the voltage drop over the sample.

Temperature and temperature drift are essential parameters for noise measurements, which affect both the measurement electronics and the sample itself. We will therefore take a closer look at the temperature stability of our non-isolated, room temperature setup. Fig. 3.2 shows the evolution of room temperature during a 2-minute and a 30-minute observation time. Exemplary temperature changes are $\Delta T_1 = 15\text{mK}$ for 2 minutes and $\Delta T_2 = 140\text{mK}$ for 30 minutes. As the sample resistors were not specifically isolated, we can assume that their temperature is directly linked to the room temperature. Therefore, the values $\Delta T_1$ and $\Delta T_2$ are used as typical values when discussing the influence of temperature fluctuations on our measurements.

Figure 3.2: Measured fluctuations of the laboratory temperature for different time periods: $\Delta T_1 = 15\text{mK}$ is the measured temperature fluctuation for 2 minutes (a) and $\Delta T_1 = 15\text{mK}$ for 30 minutes (b). As the sample resistors were not isolated specifically, the sample resistor’s temperature drift should be of the same order of magnitude as the change in laboratory temperature. The values obtained from these diagrams are used to estimate the influence of temperature drift on the devices and samples used in this work.

An overview of the specifications of all the used measurement electronics in the following setups can be found in the appendix of this work (cf. Chap. 5).

3.2.2 ”Conventional” 4 point resistance measurement

A first approach to measure small voltage changes is using a highly sensitive voltmeter, a so called ”nanovoltmeter” (Keithley Instruments K2182A) in combination with a sourcemeter as current source (Keithley Instruments K2400). We will now take a closer look on the operation principle and noise performance of these two devices, starting with the Keithley Instruments K2182A. Fig. 3.3 shows a block diagram of a chopper type dc nanovoltmeter. The dc input signal is chopped with a certain frequency, converting it into an ac signal. The chopper frequency is set
to a level where the influence of pink noise can be neglected (a typical value is \( f = 60\text{Hz} \)). Afterwards the signal is amplified and reconverted to a dc signal by a second chopper. The last stage is a low-pass filter. The main advantage of this method is that the flicker-noise of the amplifier’s input can be minimized by choosing a high enough chopper frequency. The noise level generated by the nanovoltmeter is then determined by the bandwidth of the low-pass filter. It is also remarkable that the device actually measures in an ac regime, although the input signal is dc.

![Figure 3.3: Schematic diagram of a dc nanovoltmeter (chopper type): The low level dc potential \( V_S \) is converted into an ac signal by a chopper (CH1) and amplified by an RC amplifier (A1). Then the signal is demodulated by a second chopper (CH2) and sent through a low-pass filter (IOA). Illustration taken from [38].](image)

The Keithley 2182A manual [40] contains information about the dc peak-to-peak noise for the chosen range and filter parameters for an observation time of two minutes or less. In particular, the noise for 1 power line cycle (PLC) and the filter count set to 10 is specified as 35nV peak-to-peak in the 10mV-range. So, if the full range is used and the total signal amplitude is \( V_{\text{sample}} = 10\text{mV} \), this would result in a resolution limit of \( \Delta V/V_{\text{sample}} = 3.5 \times 10^{-6} \). In general, the relative fluctuations of a measurement variable can be reduced by raising the amount of data points. Therefore the voltage resolution improves according to the relation \( \Delta V/V \propto \frac{1}{\sqrt{N}} \) and the peak-to-peak noise can be reduced by massive averaging. For a filter value 75 of and 5 PLCs, the manual states a peak-to-peak noise of 6nV (over a 2-minute time period). The theoretical resolution for this settings would be \( \Delta V/V_{\text{sample}} \geq 6 \times 10^{-7} \). However, such high averaging leads to a very long response time \( t_{\text{resp}} = 25\text{s} \), which describes the time "required for the reading to be settled in noise levels" [10]. Due to this fact, any SMR-type measurements will take a fairly long time. This fact makes it more difficult to keep the noise level low, because long-term stability of the device as well as the ambient temperature has to be ensured.

For the measurements with the nanovoltmeter, we use the Keithley Instruments K2400 sourcemeter as a dc current source. Its manual [11] lists up the accuracy for
a given programming range. The applied current 29µA require a programming range of 100µA, which has an accuracy of 0.031% of reading + 20nA. The total accuracy is therefore 29nA and leads to a relative current accuracy $\Delta I/I = 10^{-3}$, which is converted to a voltage accuracy $\Delta V/V_{\text{sample}} = 10^{-3}$ at the resistor. The accuracy according to the manual is specified for a 1-year timespan and a temperature tolerance of ±5°C. As this is a very wide range, we can assume that this specification will be fulfilled in our noise measurements and also in SMR-type measurements, which have a typical duration of about one hour. However, for our two-minute measurements, the assumed error might be way higher than the actual occurring error.

We now describe our experiment to test these theoretical sensitivity limitations. Figure 3.4 (a) shows the connection scheme. The used resistor had a resistance $R_{\text{sample}} = 332\Omega$ and the applied current was $I = 29\mu\text{A}$. With this current level, no additional heating of the sample should occur and the sample voltage $V_{\text{sample}}$ is still below 10mV, which means it can be observed within the 10mV-range of the K2182A. The nanovoltmeter’s filter count was set to 10 with 1 PLC, which results in an effective averaging over 10 measurement values per data point. Fig. 3.4 (b) contains the measured voltage over a two-minute time period. The peak-to-peak noise is $\Delta V = 90\text{nV}$ and the total signal $V_{\text{sample}} = 9.61948\text{mV}$, thus the resolution is $\frac{\Delta V}{V_{\text{sample}}} = 9.3 \times 10^{-6}$. This is at least one order of magnitude higher than the required resolution for investigating the SMR in YIG|Au.

![Figure 3.4: (a) Experimental setup for a conventional voltage measurement: The K2400 provides a current $I = 29\mu\text{A}$ which is sent through the metal film resistor $R_{\text{sample}} = 332\Omega$. The resistor is connected in a four-wire setup to minimize the error caused by lead resistances. The K2182A is set to 1PLC and a filter count of 10, so each data point is obtained by averaging over 10 measurement values. (b) Measured voltage over a 2-minute time period. The total signal $V_{\text{sample}} = 9.6195\text{mV}$ shows a peak-to-peak noise $\Delta V = 100\text{nV}$, leading to a resolution limit of $\Delta V/V = 9.3 \times 10^{-6}$.](image-url)

We now want to compare the measured data with the estimations concerning the noise contributions of the used devices. The specified peak-to-peak noise of the K2182A derived above, $\Delta V/V_{\text{sample}} = 3.5 \times 10^{-6}$ is about 3 times lower than the
observed noise level. So, the K2182A makes its contribution to the observed noise, but there have to be other influences in the measurement which add to the total noise. The metal film "dummy"-resistor in this measurement has a temperature coefficient of 50 ppm/K. With the temperature fluctuation $\Delta T_1 = 15$ mK for 2 minutes of observation time, the relative resistance fluctuation caused by thermal drift is $\Delta R_{th} \simeq 7.5 \times 10^{-7}$. This is one order of magnitude lower than the observed noise and therefore is a negligible influence. The observed peak-to-peak noise is far below the estimated error of the K2400, which suggests that the accuracy for the short time period of the measurement is much better, but it could still contribute to the total noise level in a significant way. However, the following discussion in Subsec. 3.2.3 reveals that the measurement electronics actually have a better noise performance than observed in this conventional 4 point measurement and the additional noise stems from non-linear effects.

### 3.2.3 Current reversal method

The setup with a nanovoltmeter and a sourcemeter can be improved by using the so called current reversal (or delta) method, which works the following way: First, a current with given – say positive – polarity is applied to the sample and the voltage drop $V_1$ over the sample is measured in 4 point configuration. After this, the current direction is reversed and the corresponding voltage drop $V_2$ is measured. The resistance is then calculated from the effective voltage $V_{\text{sample}} = \frac{V_1 - V_2}{2}$. The main advantage of this method is that effects which are independent or of even power of the current direction (e.g. thermoelectric voltages) cancel out \[38\]. However, the same electronics as in the conventional method are used and therefore the resolution is still limited by the noise performance of the devices. To implement the delta method in the setup from Fig. 3.4(a), the K2182A and K2400 must be connected with a trigger link cable. Apart from that, the experimental setup shown in Fig. 3.5(a) is effectively the same as the one with the conventional 4 point measurement. The current level for the measurement was again set to 29 $\mu$A and the filter settings were also the same (filter count 10, 1PLC). The experimental data for the delta method is plotted in Fig. 3.5(b). A peak-to-peak noise of 45 nV results in a resolution of $\Delta V/V_{\text{sample}} = 4.7 \times 10^{-6}$, which is about a factor of 2 lower than the $\Delta V/V_{\text{sample}}$ obtained with the conventional voltage measurement.

As the same resistor is used as in the conventional method, the relative noise caused by temperature drift has the same value $\Delta R_{th} \simeq 7.5 \times 10^{-7}$ and therefore does not have a significant influence on the noise level. The accuracy of the delta method is specified in the K2182 manual as the sum of the accuracies of the used current source and of the used channel of the nanovoltmeter. But as we have seen before, the accuracy specified for the K2400 is much worse than the actual observed noise, because this accuracy is specified for a wide time and temperature range. However, for the filter settings used in these two experiments, a dc peak-to-peak noise performance of 35 nV is stated in the K2182A manual. Even if the delta method improves this

\[3\]claimed from the specifications
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Figure 3.5: (a) Experimental setup for the delta-method: The same devices as in Fig. 3.4 are used, connected identically and also the current magnitude $I = 29 \mu A$ and filter settings (1PLC, 10counts) are the same. (b) Measured voltage over a 2-minute time period. The total signal $V_{\text{sample}} = 9.62243mV$ shows a peak-to-peak noise $\Delta V = 45nV$, leading to a resolution limit of $\Delta V/V = 4.7 \times 10^{-6}$. The resolution has improved by a factor of 2 compared to the conventional voltage measurement.

value a bit, it is likely that the observed noise is close to the performance limit of the measurement electronics. A higher averaging can improve this performance limit, but will also require longer observation times to get the same amount of data points. As our setup is not isolated or temperature controlled, the noise caused by temperature drift during a longer time period will cancel out the improved performance with higher averaging. However, when using a cryostat, the long-term temperature stability is much better (cf. Subsec. 3.2.6), so the noise performance will not get significantly worse for a longer observation time. Choosing a high averaging value can therefore be an effective way to improve the noise performance in this measurement environment.

3.2.4 Lock-in amplifier

We have seen that a conventional approach to measure small voltage changes with a nanovoltmeter and a dc current source (and also the improvement by applying a delta-method) can not provide a high enough resolution for measuring the predicted SMR in YIG|Au-bilayers. Therefore, we now want to investigate a second method to measure voltages with high sensitivity using a so called lock-in amplifier [42][38], whose operating principle is shown in fig. 3.6. The lock-in multiplies the input signal $V_{\text{in}}$ with an internal reference signal $V_{\text{ref}} = V_r \sin(\omega_r t + \varphi_r)$ with frequency $\omega_r$ and

![Diagram of experimental setup](image)

![Graph with measurements](image)
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Phase $\varphi_s$. Assuming that the input signal is also a sine function $V_{\text{in}} = V_s \sin(\omega_s t + \varphi_s)$ with frequency $\omega_s$ and phase $\varphi_s$, the multiplied function can be written as

$$V_{\text{mult}} = V_r V_s \sin(\omega_r t + \varphi_r) \sin(\omega_s t + \varphi_s) = \frac{1}{2} V_r V_s \left[ \cos((\omega_r - \omega_s)t + \varphi_r - \varphi_s) - \cos((\omega_r - \omega_s)t + \varphi_r + \varphi_s) \right]$$  (3.3)

by using the identity $\sin \theta \sin \vartheta = \frac{1}{2} (\cos(\theta - \vartheta) - \cos(\theta + \vartheta))$. This reformulation suggests that $V_{\text{mult}}$ consists of two ac signals with frequencies $\omega_+ = \omega_r + \omega_s$ and $\omega_- = \omega_r - \omega_s$. If both frequencies are the same, the difference component becomes a dc signal and it can be separated by applying a low-pass filter. The latter has an integrative effect on the signal, thus the output function equals

$$V_{\text{out}} = \frac{1}{2} V_r V_s \cos(\varphi_s - \varphi_r).$$  (3.4)

The whole process is called phase-sensitive detection (PSD). To determine the output signal, it is necessary to know both the magnitude and phase shift. Therefore a second PSD-mechanism is included in the lock-in amplifier with a reference function shifted by $90^\circ$, $V_{\text{ref}}' = V_r \cos(\omega_r t + \varphi_r)$. The output function of this channel can be derived in the same fashion as before, but due to the phase shift it is now proportional to $\sin(\varphi_s - \varphi_r)$. For a given reference signal (with $\varphi_r = 0$) the two outputs are then given by $X = V_r \cos(\varphi_r)$ and $Y = V_s \sin(\varphi_s)$. The initial input signal’s magnitude and phase can be determined by the relations

$$V_s = \sqrt{X^2 + Y^2}, \quad \varphi_s = \tan^{-1} \left( \frac{X}{Y} \right).$$  (3.5)

To provide a sinusoidal excitation signal, lock-in amplifiers have an internal oscillator, which can be set to a certain frequency and amplitude. It is also possible to use an external source instead, as the internal reference signal can be locked to both.

---

**Figure 3.6:** Basic diagram of a lock-in amplifier. After passing the input operation amplifier, the input signal is fed through two separate phase-sensitive detectors (shifted by $90^\circ$), consisting of a multiplier and a low-pass filter. Illustration from [27].

An important property of operation amplifiers (and thus of any device containing one) when investigating noise is the so called common mode rejection ratio (CMRR)

---

4Not to be confused with "power spectral density". In this work, the abbreviation "PSD" always means phase sensitive detection/detector.
An ideal differential amplifier with input voltages $V_+$ and $V_-$ produces an output voltage $V_{out} = A_d \cdot (V_+ - V_-)$, where $A_d$ is its differential gain. But this is never the case in real measurements. The output function can be better described by $V_{out} = A_d \cdot (V_+ - V_-) + \frac{1}{2} A_{cm} \cdot (V_+ + V_-)$, where $A_{cm}$ is the common-mode gain. The common mode rejection ratio (CMMR) is then defined as

$$\text{CMMR} = \left( \frac{A_d}{A_{cm}} \right) = 10 \log_{10} \left( \frac{A_d}{A_{cm}} \right)^2 \text{ dB} = 20 \log_{10} \left( \frac{A_d}{A_{cm}} \right) \text{ dB.} \quad (3.6)$$

The higher the CMMR of an electronic, the less unwanted common mode signal will appear in a measurement. Usually, the common mode signal is negligible, but for our low level measurements we should also consider this effect. The used EG&G5302 lock-in amplifier has a built in EG&G5317 input amplifier with a CMRR of 100dB. So, for example, an applied signal of $V_+ = V_- = 10\text{mV}$ and a gain of 1 will produce a common mode signal of 100nV.

The EG&G5302 manual [43] only contains information about the input noise of the direct input channel, but no input noise for the EG&G5317 preamplifier is stated. However, it is likely that the preamplifier’s input noise is in the same range as the direct channel’s or even better, so we assume a value of $20\text{nV}/\sqrt{\text{Hz}}$. The manual also does not include information about the ideal operating point of the EG&G5302 or the EG&G5317 preamplifier (noise figure, cf. the discussion of the SR560s later in this Subsec.).

![Figure 3.7: Noise figure of the SR560 as function of frequency and resistance](image)

Figure 3.7: Noise figure of the SR560 as function of frequency and resistance: A high noise figure value means the amplifier’s output noise is dominated by the device’s own noise, while a low value indicates that the output noise is governed by the source’s thermal noise. The noise figure of the SR560 has its lowest value for frequencies between $10^1 - 10^4\text{Hz}$ and source resistances between $10^5 - 10^6\Omega$. Illustration from [44].

The used EG&G5302 lock-in has a 4-digit resolution, so the observable change in voltage is limited to $10^{-4}$ of the chosen range, which can be set from 1V to 100nV.
Hence, the setup must be tuned to obtain more effective digits. A possible solution is implementing a voltage compensation. The basic idea is to recreate the sample’s resistance very accurately with a compensation box containing ohmic resistors and potentiometers. After applying an excitation signal, the voltage drop over the sample and the compensation box will be nearly the same. Taking the difference of both then will leave a compensated signal with a much smaller magnitude, which effectively raises the resolution by the factor of compensation. One can also see this method as a background subtraction. The SMR-effect, which is a modulation of the voltage drop over $R_{\text{sample}}$ is not affected by this method and its modulation $\Delta V$ will still depend on $V_{\text{sample}}$. As introduced in Subsec. 3.2.1, the longitudinal resistances of the already existing YIG|AU samples with $6\text{nm} \leq t_{\text{Au}} \leq 27\text{nm}$ vary in a range $10\Omega \leq R_{\text{Au}} \leq 500\Omega$. To recreate resistance values of this range, we re-designed an already existing so called compensation box consisting of $R_{\text{current}} = 10k\Omega$ and two potentiometers with ranges between 0 and 500 Ohms and 0 and 10 Ohms, respectively mounted on the outer surface of a aluminum box. The two potentiometers are linked in series, such that one can create $0 \leq R_{\text{Box}} = R_1 + R_2 \leq 510\Omega$ by hand. The additional resistor $R_{\text{current}}$ can be implemented in series to the potentiometer by a manual switch. The application of this additional resistor can be used to determine the current through the box. As also shown in Fig. 3.8 two BNC connectors on the left and right side allow to wire the box to the $R_{\text{sample}}$ and electronic devices, whereas the two BNC connectors on top can be used for a connection to a voltmeter to measure the voltage drop over $R_1 + R_2$.

![Figure 3.8](image)

**Figure 3.8:** Schematic diagram of the compensation box: The resistance $R_{\text{current}} = 10k\Omega$ is used to define a certain current level when provided with a certain voltage. $R_1$ is a potentiometer with a resistance of 0-500$\Omega$ and $R_2$ a potentiometer with a resistance of 0-10$\Omega$. The two potentiometers can be adjusted manually to reproduce a certain resistance in the range of 0-510$\Omega$. All resistors are connected in series.

To check how effectively a signal can be compensated, the setup shown in Fig. 3.9(a) was used. The box was connected in series to the test resistance $R_{\text{sample}} = 47\Omega$ and a sourcemeter provided a dc signal $I = 1mA$. The voltage drop over the sample $V_{\text{sample}}$ as well as the voltage drop over the compensation box $V_{\text{box}}$ were fed in the two input channels of a nanovoltmeter. The nanovoltmeter was set to show the difference signal between both input channels. The potentiometers $R_1$ and $R_2$ of the compensation box were then adjusted until the difference signal of channel 1 and channel 2 reached a minimal value. Fig. 3.9(b) shows the measured voltage in the difference channel. In our test experiment, the original voltage drop along the sample, $V_{\text{sample}} = 48.6\text{mV}$, could be reduced to $V_{\text{comp}} = 1.3\text{\mu V}$. This leads to a reduction factor $Q = V_{\text{comp}}/V_{\text{sample}} = 2.7 \times 10^{-5}$. This means we managed to compensate more
than 4 significant digits. Combined with the lock-in, relative changes of $2.7 \times 10^{-9}$ in voltage could be detected by the combination of the compensation box and the lock-in (of course, this is only the technical limit that does not include the additional limitation by noise).

**Figure 3.9:** (a) Experimental setup for the voltage compensation: The compensation box is connected in series with a "dummy"-resistor $R_{\text{sample}} = 47\Omega$ and a dc current $I = 1\text{mA}$ provided by the *Agilent B2911A* sourcemeter is fed through the circuit (the resistance $R_{\text{current}}$ is set to zero for this experiment. The voltage drop over the box is measured in channel 1, the voltage drop over the resistor in channel 2 of the *Agilent 34420A* nanovoltmeter. The potentiometers $R_1$ and $R_2$ are adjusted such, that the difference signal (channel 1 - channel 2) of the nanovoltmeter reaches a minimal value.

(b) Measured voltage of the difference channel over time ($t_{\text{meas}} = 2\text{min}$): The total signal of the sample $V_{\text{sample}} = 48.6\text{mV}$ can be compensated down to $V_{\text{comp}} = 1.3\mu\text{V}$.

To convert the voltage drop over the sample and the box into a differential signal, we use two *SR560* differential preamplifiers. The SR560 manual contains information about the so called noise figure (NF), which is defined as

$$\text{NF} = 20 \log \left( \frac{\text{OutputNoise}}{\text{Gain} \cdot \text{SourceThermalNoise}} \right)$$

(3.7)

The value of the NF depends on the frequency and source impedance and is plotted with contour lines. The frequency should be chosen such, that the influence of 1/f-noise can be neglected, which is the case for $f \geq 100\text{Hz}$. The used frequency $f = 1.105\text{kHz}$ in the measurement is therefore suitable. This frequency in combination with the sample resistance $R_{\text{sample}} = 323\Omega$ correspond to a noise figure $\text{NF} \approx 2.5\text{dB}$, which leads to an output noise $\Delta V_{\text{Output}} = 3.1\text{nV}/\sqrt{\text{Hz}}$. At the ideal operating point, the NF value is 0.05dB and therefore the output noise equals $\Delta V_{\text{Output}} = 2.3\text{nV}/\sqrt{\text{Hz}}$. To reach this value, the source impedance has to be raised by two orders of magnitude. Simply adding resistance will not improve the noise performance, as then the signal’s thermal noise will also rise proportional to $\sqrt{R}$ and the output noise will then be dominated by this term instead of the noise figure.

Now we combine the lock-in amplifier and the compensation to the experimental setup shown in Fig. 3.10(a) and measure the noise performance. The voltage drop
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over the sample as well as $R_{\text{box}}$ is fed into two SR560 preamplifiers (gain = 1, filter = bandpass). The output signals of both preamps are fed into channel A and B of the lock-in differential amplifier, respectively. The lock-in amplifier provides an excitation voltage $V_{\text{Osc}} = 1\text{V}$ with frequency $f_{\text{Osc}} = 1.105\text{kHz}$. The lock-in frequency used is high enough to avoid the influence of 1/f-noise or distortion caused by the 50Hz line frequency and its multiples. Also, the SR560-preamplifiers’ noise figure value is minimal in this frequency range. The potentiometers of the compensation box are set to a combined resistance $R_1 + R_2 = R_{\text{Box}} = 323\Omega$ equal to the sample resistance $R_{\text{sample}} = 323\Omega$.

\[\text{Figure 3.10: (a) Used setup for the measurement with voltage compensation and the lock-in amplifier: The internal oscillator of the lock-in provides an ac voltage } V_{\text{Osc}} = 1\text{V. } R_{\text{current}} = 10\text{k}\Omega \text{ is much larger than } R_{\text{sample}} + R_{\text{Box}} = 646\Omega \text{ and therefore defines the current level. Two differential preamplifiers measure the voltage drop over the sample respectively the compensation box. The two signals are then fed into the A and B channel of the lock-in amplifier, resulting in a small difference signal. The lock-in was set to a range of } 100\mu\text{V and an integration time } \tau = 1\text{s. (b) Measured voltage over a 30-minute time period: The total sample voltage } V_{\text{sample}} = 29.4\text{mV is compensated down to } V_{\text{comp}} = 82.9\mu\text{V. The peak-to-peak noise for this measurement is } \Delta V = 140\text{nV}, \text{so the resolution limit is given by } \Delta V/V_{\text{sample}} = 4.8 \times 10^{-6}.\]

The obtained experimental values are plotted in Fig. 3.10(b). The first thing to notice is, that the compensated signal $V_{\text{comp}} = 82.9\mu\text{V}$ is almost two orders of magnitude larger than the signal obtained in the test-setup of the compensation box (cf. Fig 3.9(b)), although the sample voltages were in the same range for both cases. Further investigation has revealed that this offset-voltage scales with frequency and that the offset-voltage is caused by a phase-shift between the two input signals at the A and B channel of the lock-in. This might be due to different length in the used wiring and could be improved by adding a phase shifting circuit in future investigations [45]. Due to the fact that the compensated signal $V_{\text{comp}} = 82.9\mu\text{V}$ is still comparably large, the observation range is limited to 100\mu V and so the last digit of the lock-in’s display is equivalent to 10nV. As the total noise in the measurement is $\Delta V = 140\text{nV}$ peak-to-peak, which is only by a factor of 14 larger than the sensitivity of the lock-in, the measured voltage values show a quantized character. The voltage
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Drop over the sample in this measurement is $V_{\text{sample}} = 29.4\text{mV}$, so the resolution limit is given by $\Delta V/V_{\text{sample}} = 4.8 \times 10^{-6}$.

Now we want to discuss where this noise level stems from. At the chosen frequency $f_{\text{Osc}} = 1.105\text{kHz}$, 1/f noise has no significant influence on the resistor itself and also the used devices have no more 1/f-portion in their input noise. So the intrinsic noise contributions of the resistor are thermal noise and Shot-noise. Using the values $I = 94\,\mu\text{A}$, $V_{\text{sample}} = 29.4\text{mV}$ and Eqs. 3.1–3.2 a total noise of $\sqrt{(1.8\text{nV})^2 + (2.3\text{nV})^2}/\sqrt{\text{Hz}} = 2.9\text{nV}/\sqrt{\text{Hz}}$ is produced by the "dummy"-resistor and, of course, the same noise is produced by the resistors in the compensation box. This adds to the input noise of the SR560-preamplifiers, which is specified as $4\text{nV}/\sqrt{\text{Hz}}$, resulting in a total input noise of $\sqrt{(2.9\text{nV})^2 + (4\text{nV})^2}/\sqrt{\text{Hz}} = 4.9\text{nV}/\sqrt{\text{Hz}}$. With $\text{NF} \approx 5\text{dB}$, the total output noise at each of the two SR560s is $8.7\text{nV}/\sqrt{\text{Hz}}$. Now this noise adds to the lock-in’s input noise at both channels A and B, which we assumed to be $20\text{nV}/\sqrt{\text{Hz}}$, resulting in an output noise of $\sqrt{(8.7\text{nV})^2 + (20\text{nV})^2}/\sqrt{\text{Hz}} = 22\text{nV}/\sqrt{\text{Hz}}$. The difference channel then has a total output noise of $\sqrt{(22\text{nV})^2 + (22\text{nV})^2}/\sqrt{\text{Hz}} = 31\text{nV}/\sqrt{\text{Hz}}$. To calculate the output noise in units of Volt, this value has to be multiplied by the square root of the equivalent noise bandwidth (ENBW). For the applied 12dB/oct-filter of the lock-in and the integration time $\tau = 1\text{s}$ it is determined by $\sqrt{\text{ENBW}} = \sqrt{1/8\tau} = 0.35\sqrt{\text{Hz}}$ [42].

So, the total noise produced by the devices and the resistors is $\Delta V(\text{RMS}) = 11\text{nV}$ or $\Delta V(\text{pp}) = 72\text{nV}$. This noise is by a factor of 2 lower than the actual observed noise level, so there must be other noise sources which add to the actual measured value.

The SR560s has a CMRR of 90dB and the EG&G5317 preamplifier has a CMRR of 100dB. This means, for $V_{\text{sample}} = 29.4\text{mV}$ the SR560 introduces $929\text{nV}$ common mode signal and the EG&G5317 $294\text{nV}$. However, these signals only lead to small offsets compared to the compensated voltage $V_{\text{comp}}$.

Another possible source of the additional noise appearing in the measurement are fluctuations caused by temperature drift. We therefore now discuss the temperature stabilities of the SR560-preamplifiers and the EG&G5302 lock-in. The SR560’s dc temperature drift is specified as $5\mu\text{V}/\text{K}$. Multiplying this value by the temperature fluctuation $\Delta T_2 = 140\text{mK}$ from Fig. 3.2 results in an voltage drift of $700\text{nV}$. Such a high value was not observed in the measurement, but even when assuming a 10 times better temperature stability, this thermal drift would still contribute in a significant way.

The amplitude stability of the EG&G5302 lock-in is specified as $0.02\%/\text{K} = 200\text{ppm}/\text{K}$. This leads to a relative fluctuation $\Delta V_{\text{Osc}}/V_{\text{Osc}} = 2.8 \times 10^{-5}$. However, as the lock-in amplifier has only a 4-digit resolution, this should have no significant influence on the noise level.

In this measurement, the used resistor was a carbon film resistor. It has a typical temperature coefficient of $600\text{ppm}/\text{K}$. The relative fluctuation of the sample is therefore $8.4 \times 10^{-5}$, so the temperature drift of the resistor itself is a massive source of noise, with even more impact than the two electronics. This effect can be reduced by using a metal film resistor instead of the carbon film resistor. The temperature
coefficient of the metal film resistor is by a factor of 0.12 lower than the temperature coefficient of the carbon film resistor. This means that the temperature drift would still lead to a relative fluctuation of $3.4 \times 10^{-6}$.

To sum it up, we figured out that not the electronic devices used in our method give the dominant part of the measured noise and thus limit the resolution, but the voltage noise from the thermal coefficient of the resistance used itself. The noise performance of this method could be significantly improved by using a cryostat-system. Of course, the temperature stability of such an isolated system is much better than the stability in our room temperature measurements with no isolation. All the temperature depending fluctuations of the devices and the resistor itself can be minimized this way. Another fact to consider is that Johnson noise gets lower $\propto \sqrt{T}$, thus measuring in a cooled system can reduce the intrinsic noise significantly compared to a room temperature setup.

### 3.2.5 Comparison

Table 3.1 sums up the obtained resolution limits for each measurement method. The delta method yielded the best value, but the measurements with the lock-in amplifier are not significantly worse. However, all three methods have a resolution limit in the range of $10^{-6}$, which is not sufficient for our goal to measure effects in the range of $10^{-7}$.

<table>
<thead>
<tr>
<th>Method</th>
<th>$\Delta V / V_{\text{sample}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conventional method</td>
<td>$9.3 \times 10^{-6}$</td>
</tr>
<tr>
<td>Delta method</td>
<td>$4.7 \times 10^{-6}$</td>
</tr>
<tr>
<td>Lock-in amplifier</td>
<td>$4.8 \times 10^{-6}$</td>
</tr>
</tbody>
</table>

**Table 3.1:** Resolution limits for the different measurement methods.

We have seen that the delta method can improve the resolution limit by a factor of 2 in comparison to the "conventional" method. So, when measuring with a dc setup using a combination of a nanovoltmeter and a sourcemeter, the delta method should always be preferred. The fact that non-linear effects in $I$ are canceled out is an important improvement. However, the measured noise performance for the current reversal method is close to the limit of the used electronics (for the used filter settings). Further improvements can therefore only be made by applying very high filter settings. The maximal theoretical resolution would then be approximately $6 \times 10^{-7}$ for 5 PLCs and 75 counts.

The resolution of the lock-in method was determined by the thermal fluctuations of the used resistor. However, even if this effect could be eliminated, the input noise of the SR560s and the EG&G5302 would still make a significant contribution to the total noise level and therefore limit the resolution. The noise performance can be improved by using a better lock-in amplifier, e.g. the *Stanford Research Systems SR830* has an input noise of 6nV/$\sqrt{\text{Hz}}$ (at 1kHz) [12], which is significantly lower than the
assumed input noise of $20\text{nV}/\sqrt{\text{Hz}}$ of the EG&G5302. Also, the SR830’s filter can be set to $24\text{dB/oct}$, so the equivalent noise bandwidth determined by $\text{ENBW} = 5/64\pi$ for this filter setting is also lower than the ENBW of the EG&G5302.

### 3.2.6 Temperature stability control

The previous discussion has shown that temperature stability is a crucial requirement to enable very sensitive resistance measurements. The measurements at room temperature were done without an effective isolation or temperature control of the sample. For SMR-type measurements, the sample is usually installed in a cryostat, which ensures a much better thermal stability. A typical fluctuation at $T = 300\text{K}$ is $\Delta T_{300\text{K}} = 10\text{mK}$. For lower temperature ranges of e.g. $T = 50\text{K}$, a better stability $\Delta T_{50\text{K}} = 3-5\text{mK}$ can be reached. So, for an estimation, we can take $\Delta T = 3\text{mK}$ as the maximum thermal fluctuation in a cryostat. Now the temperature coefficient of resistance of a YIG|Au-thin-film has to be investigated to see whether a temperature change of $3\text{mK}$ is low enough to enable a voltage measurement resolution of $10^{-7}$ or better. To do so, a YIG|Au-sample with a $15\text{nm}$-thick Au-layer was installed in a cryostat and cooled down to $10\text{K}$. Figure 3.11 shows the longitudinal resistance $R_{\text{long}}$ for a temperature sweep from 10 to 65K. Two linear fits in the ranges $10-12\text{K}$ and $30-65\text{K}$ yield temperature coefficients $C_1 = 6.6 \times 10^{-2}\text{Ω/K}$ and $C_2 = 0.64\text{Ω/K}$ respectively. The resistances in the respective ranges can then be calculated by

$$R_{\text{long,1}} = C_1 \cdot T + 220.7\text{Ω}, \quad R_{\text{long,2}} = C_2 \cdot T + 220.7\text{Ω}. \quad (3.8)$$

With the total value of the longitudinal resistance in these ranges, $R_{\text{long}} \approx 250\text{Ω}$, the relative temperature coefficients are $\Delta R_1 = 264\text{ppm/K}$ and $\Delta R_2 = 2560\text{ppm/K}$. The literature value of the temperature coefficient $\alpha_T \simeq 3.4 \times 10^{-3}/\text{K}$ for gold at $300\text{K}$ has the same order of magnitude as our experimentally determined value $\Delta R_2$ for $30-65\text{K}$. However, the measured coefficient for the lower temperature range $10-12\text{K}$ $\Delta R_1$ is distinctly lower than the literature value for room temperature.

If we now assume temperature fluctuations in the cryostat $\Delta T = 3\text{mK}$ during a measurement, the relative resistance fluctuations arising from the mK temperature fluctuations are $\Delta R_1/R = 7.92 \times 10^{-7}$ for a base temperature $10\text{K} \leq T_{\text{meas}} \leq 12\text{K}$ and $\Delta R_2/R = 7.7 \times 10^{-6}$ for $30\text{K} \leq T_{\text{meas}} \leq 65\text{K}$. This means that the temperature dependent changes in resistance even in the regime of the lowest temperature coefficient are greater than the expected SMR in YIG|Au from Fig. 2.5. In other words, thermal drift already limits the resolution to a level which is not sufficient to measure the predicted SMR-effect for YIG|Au-bilayers. To achieve a resolution of $10^{-7}$ with the measured relative temperature coefficient $\Delta R_1 = 264\text{ppm/K}$, the temperature drift must not be higher than $\Delta T = 380\text{µK}$. This would mean the temperature stability has to be improved by a factor 8.
Figure 3.11: Temperature evolution of the resistance of an Au thin film \( (t = 15\text{nm}) \) on top of YIG in the range \( 10\text{K} \leq T \leq 65\text{K} \). The red and blue dashed lines indicate linear fits to the experimental data in the range \( 30 \leq T \leq 65 \text{K} \) (red) and \( 10 \leq T \leq 12 \text{K} \) (blue).
The initial motivation of this thesis was to measure the spin Hall magnetoresistance (SMR) effect for YIG|Au-bilayers, which has a predicted magnitude $\Delta R/R \simeq 2.2 \times 10^{-7}$ (cf. Fig. 2.5). Since this is a rather small resistance change, very sensitive measurement methods are required. For this purpose, different measurement methods have been tested in Sec. 3.2.

The first approach (cf. Subsec. 3.2.2), a "conventional" 4 point resistance measurement using a highly sensitive voltmeter, yielded a resolution limit of $\Delta R/R = 9.3 \times 10^{-6}$ (for the chosen filter settings). To improve this noise performance, the current reversal method can be implemented in the first setup (cf. Subsec. 3.2.3). With this technique, each measurement point $V$ is determined by the difference between the voltage $V_1$ for an applied current $I$ and the voltage $V_2$ for the reversed current $-I$. With this procedure all noise caused by effects independent or of even power of $I$ cancel out. The obtained resolution limit for this method was $\Delta V/V_{\text{sample}} = 4.7 \times 10^{-6}$, so the noise in the "conventional" approach must have contained contributions from non-linear effects. The measured noise performance of the current reversal method is close to the performance limit of the used measurement electronics for the chosen filter settings. This means the noise performance can not be improved anymore, unless by going to massive filtering values. However, raising the filter level also increases the measurement time, which leads to higher temperature fluctuations during the measurement. The changes in resistance caused by thermal drift will then counteract the reduced noise level obtained from higher filtering. A different technique using a lock-in amplifier in combination with a compensation box has been discussed in Subsec. 3.2.4. The lock-in amplifier acts as a narrow bandpass and therefore reduces the detected noise in the measurement. The compensation box is needed because the lock-in has a resolution of only 4 digits, so, for detecting signals of the magnitude $10^{-7}$, the initial signal $V_{\text{sample}}$ has to be compensated by a factor of at least $10^3$ before it is sent into the lock-in. With this method, we obtained a resolution limit $\Delta V/V_{\text{sample}} = 4.8 \times 10^{-6}$. The noise level in the setup stems from two main contributions, the intrinsic noise of the used resistor and the input noise of the measurement electronics as well as resistance fluctuations.
caused by temperature drift. Taking everything into account, all three methods are limited to a resolution with a magnitude of $\Delta R/R \approx 10^{-6}$.

During the discussion of the noise performance of each setup, it became apparent that a high temperature stability is a crucial condition to maintain a low noise level. Therefore, we took a closer look at the resistance changes caused by temperature fluctuations in an isolated cryostat-system in Subsec. 3.2.6 of chapter 3. The result was that the typical thermal stability in a cryostat ($\Delta T = 3\,\text{mK}$) leads to a relative resistance fluctuation in an Au-film (any bulk-like gold sample) $\Delta R/R_{\text{sample}} \geq 7.9 \times 10^{-7}$. This means that the resolution is already limited to a level insufficient for the required $\Delta V/V_{\text{sample}} \simeq 2.2 \times 10^{-7}$ by thermal drift alone. This is the key result of this thesis: The experimental determination of resistance changes smaller than $\approx 10^{-6}$ is very difficult and possible only under particular circumstances.

A first option to reach a sufficient resolution for measuring the SMR in YIG$|$Au would be trying to reach a better thermal stability during the measurement. But, as already mentioned at the end of Subsec. 3.2.6, this would require a thermal fluctuation $\Delta T \leq 380\,\mu\text{K}$. In other words, the thermal stability has to be increased by approximately one order of magnitude in comparison to the stability typically observed in a cryostat-system. So, fundamental improvements of cryogenic technology are required and it is unlikely that this issue can be resolved in a short time period. The change in resistance caused by temperature drift also depends on the temperature coefficient of resistance of the sample. However, the coefficient of gold and other typical NMs used for SMR-type experiments are very similar ($\alpha_T \simeq 4 \times 10^{-3}/\text{K}$)\textsuperscript{[46]}, which means that no significant improvement can be made by changing the type of metal used.

Nevertheless, taking a closer look at the used sample itself, particularly the chosen NM, is a more promising approach. The reason why Au was chosen is that it is far away from the Stoner-criterion and therefore no proximity-based magnetoresistance is expected. So, measuring an additional resistance would have supported the SMR-theory based on spin currents. The fact that gold has only a small spin Hall angle $\alpha_{\text{SH}} = 0.0016$\textsuperscript{[20]} is the main reason that the predicted SMR-effect has such a small magnitude. Finding a different NM that is also nowhere near the Stoner-criterion, but has a significantly larger spin Hall angle would be a sensible. Ideally, the predicted SMR-magnitude should be in the range of $10^{-6}$ or larger so it can be detected by the methods discussed in this thesis. In\textsuperscript{[47]}, Wang et al. report a spin Hall angle $\alpha_{\text{SH}} = -0.14$ for a YIG$|$W-bilayer, obtained from spin pumping experiments. This value is close to the spin Hall angle of Pt ($\alpha_{\text{SH}} = 0.1$\textsuperscript{[3]} and therefore one would also expect a SMR-effect of the magnitude $10^{-3}$ (assuming similar values for the spin mixing interface conductance $G_{\uparrow\downarrow}$ and the spin diffusion length $\lambda$). Measuring the SMR with this comparably big magnitude can easily be done with any of the discussed methods. Of course, it has to be investigated first whether tungsten is also far away from the Stoner-criterion. Nevertheless, YIG$|$W-bilayers might be an interesting topic for future research.
Chapter 5

Appendix

This appendix contains the most important specifications of the used measurement electronics. More detailed information can be found in the respective manuals linked in the references.
Figure 5.1: Specifications of the direct input channel of the EG&G5302 lock-in amplifier as well as the specifications for the different available preamplifiers. The lock-in used in the experiments of this thesis had a built-in EG&G5317 preamplifier. Excerpt from the EG&G5302 manual [43].

<table>
<thead>
<tr>
<th>NOT APPLICABLE</th>
<th>100 DB</th>
<th>140 DB</th>
<th>100 DB</th>
<th>40 DB</th>
</tr>
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<tbody>
<tr>
<td>SINGLE</td>
<td>SINGLE</td>
<td>SINGLE</td>
<td>SINGLE</td>
<td>SINGLE</td>
</tr>
</tbody>
</table>

Low Frequency Performance
- 10 MHZ to 15 kHz
- 10 kHz to 1 kHz
- 1 kHz to 100 Hz

Input Impedance
- 50 Ohms
- 100 Ohms
- 500 Ohms

Preemphasis
- 0 dB
- 10 dB
- 20 dB

Conversion Gain
- 20 dB
- 40 dB
- 60 dB

Input Mode
- DIRECT
- NORMAL
- SHUNT

Input Amplifier Specifications
DC Noise Performance \^7 (DC noise expressed in volts peak-to-peak)

Response time = time required for reading to be settled within noise levels from a stepped input, 60Hz operation.

### Channel 1

<table>
<thead>
<tr>
<th>Response Time</th>
<th>NPLC, Filter</th>
<th>10 mV</th>
<th>100 mV</th>
<th>1 V</th>
<th>10 V</th>
<th>100 V</th>
<th>NMRR ^8</th>
<th>CMRR ^8</th>
</tr>
</thead>
<tbody>
<tr>
<td>25.0 s</td>
<td>5, 75</td>
<td>6 nV</td>
<td>20 nV</td>
<td>75 nV</td>
<td>750 nV</td>
<td>75 \mu V</td>
<td>110 dB</td>
<td>140 dB</td>
</tr>
<tr>
<td>4.0 s</td>
<td>5, 10</td>
<td>15 nV</td>
<td>60 nV</td>
<td>160 nV</td>
<td>1.5 \mu V</td>
<td>75 \mu V</td>
<td>100 dB</td>
<td>140 dB</td>
</tr>
<tr>
<td>1.0 s</td>
<td>1, 18</td>
<td>25 nV</td>
<td>175 nV</td>
<td>600 nV</td>
<td>2.5 \mu V</td>
<td>100 \mu V</td>
<td>95 dB</td>
<td>140 dB</td>
</tr>
<tr>
<td>60 ms</td>
<td>1, 10 or 5, 2</td>
<td>35 nV</td>
<td>250 nV</td>
<td>650 nV</td>
<td>3.3 \mu V</td>
<td>150 \mu V</td>
<td>90 dB</td>
<td>140 dB</td>
</tr>
<tr>
<td>60 ms</td>
<td>1, Off</td>
<td>70 nV</td>
<td>700 nV</td>
<td>6.6 \mu V</td>
<td>300 \mu V</td>
<td>60 dB</td>
<td>140 dB</td>
<td></td>
</tr>
</tbody>
</table>

### Channel 2\^ 9, 10

<table>
<thead>
<tr>
<th>Response Time</th>
<th>NPLC, Filter</th>
<th>10 mV</th>
<th>100 mV</th>
<th>1 V</th>
<th>10 V</th>
<th>100 V</th>
<th>NMRR ^8</th>
<th>CMRR ^8</th>
</tr>
</thead>
<tbody>
<tr>
<td>25.0 s</td>
<td>5, 75</td>
<td>150 nV</td>
<td>200 nV</td>
<td>750 nV</td>
<td>—</td>
<td>—</td>
<td>110 dB</td>
<td>140 dB</td>
</tr>
<tr>
<td>4.0 s</td>
<td>5, 10</td>
<td>150 nV</td>
<td>200 nV</td>
<td>1.5 \mu V</td>
<td>—</td>
<td>—</td>
<td>100 dB</td>
<td>140 dB</td>
</tr>
<tr>
<td>1.0 s</td>
<td>1, 10 or 5, 2</td>
<td>175 nV</td>
<td>400 nV</td>
<td>2.5 \mu V</td>
<td>—</td>
<td>—</td>
<td>90 dB</td>
<td>140 dB</td>
</tr>
<tr>
<td>85 ms</td>
<td>1, Off</td>
<td>425 nV</td>
<td>1 \mu V</td>
<td>9.5 \mu V</td>
<td>—</td>
<td>—</td>
<td>60 dB</td>
<td>140 dB</td>
</tr>
</tbody>
</table>

### Current Accuracy (Local or Remote Sense)

<table>
<thead>
<tr>
<th>Model</th>
<th>Range</th>
<th>Programming Resolution</th>
<th>Source Accuracy (1 Year) 25(^\circ)C ±5(^\circ)C</th>
<th>Default Measurement Resolution</th>
<th>Measurement Accuracy (1 Year) 25(^\circ)C ±5(^\circ)C</th>
<th>Source/Sink Limit</th>
</tr>
</thead>
<tbody>
<tr>
<td>2400, 2400-C, 2401</td>
<td>100 \mu A</td>
<td>50 pA</td>
<td>0.055% ± 600 pA</td>
<td>10 pA</td>
<td>0.025% ± 2000 pA</td>
<td>±1.05A @ ±21 V</td>
</tr>
<tr>
<td></td>
<td>10 \mu A</td>
<td>500 pA</td>
<td>0.033% ± 20 pA</td>
<td>100 pA</td>
<td>0.027% ± 700 pA</td>
<td>±10 mA @ ±210 V</td>
</tr>
<tr>
<td></td>
<td>100 \mu A</td>
<td>5 nA</td>
<td>0.031% ± 200 pA</td>
<td>1 nA</td>
<td>0.025% ± 6 nA</td>
<td>±10 mA @ ±210 V</td>
</tr>
<tr>
<td></td>
<td>1 mA</td>
<td>500 pA</td>
<td>0.034% ± 200 pA</td>
<td>10 pA</td>
<td>0.027% ± 60 nA</td>
<td>±10 mA @ ±210 V</td>
</tr>
<tr>
<td></td>
<td>10 mA</td>
<td>500 pA</td>
<td>0.034% ± 200 pA</td>
<td>100 pA</td>
<td>0.035% ± 600 nA</td>
<td>±10 mA @ ±210 V</td>
</tr>
<tr>
<td></td>
<td>100 mA</td>
<td>5 \mu A</td>
<td>0.066% ± 20 \mu A</td>
<td>1 \mu A</td>
<td>0.095% ± 6 \mu A</td>
<td>±10 mA @ ±210 V</td>
</tr>
<tr>
<td></td>
<td>1 \mu A</td>
<td>90 \mu A</td>
<td>0.27% ± 900 \mu A</td>
<td>10 \mu A</td>
<td>0.22% ± 570 \mu A</td>
<td>±10 mA @ ±210 V</td>
</tr>
</tbody>
</table>

Figure 5.2: Top: DC peak-to-peak noise performance of the K2182A nanovoltmeter for different ranges and filter settings. Excerpt from the data sheet [48]. Bottom: Current accuracy of the K2400 sourcemeter for different programming ranges. Excerpt from the data sheet [49].
**Appendix**

**Figure 5.3:** Data sheet of the SR560 preamplifier. Excerpt from the data sheet [50]
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Bibliography


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Erklärung

Mit der Abgabe der Bachelorarbeit versichere ich, dass ich die Arbeit selbständig verfasst und keine anderen als die angegebenen Quellen und Hilfsmittel benutzt habe. Die Prüfungsleistung wurde bisher bzw. gleichzeitig keiner anderen Prüfungsbehörde vorgelegt.


Ort, Datum

______________________________

Vorname Nachname