

The Analysis and Development of Sensors for Active Magnetic Bearings

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SUMMARY

This dissertation presents the analysis and development of contactless sensors for active magnetic bearings (AMBs). The rotors of AMB systems are suspended in air without any direct contact between the stator and the rotor. An AMB system requires rotor position feedback for control. To extract the position from the system some sort of contactless sensor must be constructed.

The aim of this project is to develop different types of contactless sensors for AMBs. Through this project basic knowledge of sensors is gained and expertise is established in the Engineering Faculty. These models can be used for further research investigations.

This project is divided into three different sub-projects; 1) an optical sensor, 2) an inductive sensor and 3) a self-sensing sensor. Each of these sensors are analysed, developed and tested on AMB models.

The optical sensor constitutes an infrared transmitter and an infrared receiver. This transmitting and receiving technique relies on the amount of light transmitted and is known as the barrier optical sensor. This sensor was evaluated on a magnetic bearing demonstrator. The optical sensor has a very high bandwidth and is ideal when low noise is a requirement.

The inductive sensor principle of operation is based on the changing effect of the inductance due to the air gap. If the coil is placed near a ferromagnetic material the inductance of the coil will change according to the size of the air gap. The sensor's coils are activated with a high frequency voltage. This constant frequency voltage will be applied to a constant reactance if the air gap is constant. As the distance between the rotor and the stator of the AMB changes, the air gap changes, thus the output of the sensor coils is proportional to the distance.

The inductive sensor was implemented on a homopolar AMB with a maximum rotating speed of 3000 rpm. It was determined that the inductive sensor has good linearity

properties over a specific measurement range. The bandwidth of the inductance sensor was designed to be above 500 Hz, which is ten times higher than the bandwidth of the homopolar AMB. In the case of very high rotating speeds, the bandwidth of the inductive sensor may be too low for geometric position control.

The self-sensing technique makes use of the current response of the power amplifier and the voltage across the power amplifier. The slope and amplitude of the high frequency current component changes according to the air gap. This changing current is demodulated to determine the position of the rotor. By using a self-sensing method the voltage and current measurements could be taken externally from the AMB. These voltages and currents are then converted to an estimation of the rotor position. This external electrical measurement has the advantage that there are no sensors present inside the AMB. The self-sensing method was evaluated on a heteropolar active magnetic bearing.

It was determined that the self-sensing sensor bandwidth was high enough to overcome disturbances up to 100 Hz. The sensor was successfully implemented on a heteropolar AMB with switching power amplifiers. The sensor output contains a 100 Hz noise component due to a ripple component on the power amplifier capacitor. The linearity of the sensor is impaired by eddy current losses in the AMB.

Future work will focus on the improvement of the self-sensing sensor. The eddy current losses could be minimized by using thinner laminations in the AMB magnetic circuit. This will reduce the losses and increase the linearity of the self-sensing sensor. By implementing a differential sensor where the air gaps on both sides of the rotor are estimated and subtracted a more sensitive and noise immune system could be obtained. Implementation by means of a digital signal processing (DSP) device is also an important future activity.

This project produces a firm basis on sensor technology in the research group.

OPSOMMING

In die verhandeling word die analise en ontwikkeling van kontaklose sensors soos van toepassing op aktiewe magnetiese laers (AMLs) voorgelê. Die rotors van AML stelsels word in die lug gesuspendeer sonder dat die stator direk kontak maak met die rotor. 'n AML stelsel vereis rotorposisie terugvoer vir beheer. Die posisie word bepaal deur van kontaklose sensors gebruik te maak.

Die doel van die projek is om verskillende kontaklose sensors te ontwikkel. Die projek sal bydra tot verbeterde deskundigheid asook basiese kennis in die veld van sensortegnologie in die Ingenieursfakulteit. Die modelle wat ontwikkel word kan ook vir toekomstige navorsing op die gebied van AMLs gebruik word.

Die projek word in drie sub-projekte verdeel; 1) 'n optiese sensor, 2) 'n induktiewe sensor en 3) 'n self-waarnemende sensor. Elk van die drie sensors word ge-analiseer, ontwikkel en op AML modelle ge-evalueer.

Die optiese sensor bestaan uit 'n infrarooi versender en 'n infrarooi ontvanger. Die sensortegniek maak gebruik van die hoeveelheid infrarooilig wat deurgelaat word. Die sensor staan bekend as die optiese versperring sensor. Die sensor is ge-evalueer op 'n AML demonstrasiemodel. Die bandwydte van die optiese sensor is baie hoër as dié van die beheerseine. Die ruisvlak van die optiese sensor is baie laag, dus ideaal vir lae ruisvlak posisiemetings.

Die induktiewe sensor maak gebruik van die veranderende induktansie as gevolg van die verandering in die luggaping. As 'n spoel naby 'n ferromagnetiese materiaal geplaas word, verander die induktansie van die spoel soos die luggaping verander. Die sensorspoele word geaktiveer deur 'n hoë frekwensie spanning. Hierdie konstante frekwensie sein word toegepas op 'n konstante reaktansie solank die luggaping konstant bly. Die luggaping verander egter soos die rotor ten opsigte van die middelpunt verskuif.

Hierdie verandering in die luggaping lei tot 'n verandering in die reaktansie. Die sensor se posisie-uitvoer is dus direk eweredig aan die veranderende reaktansie.

Die induktiewe sensor is op 'n homopolêre AML wat teen 3000 rpm roteer geïmplementeer. Die induktiewe sensor beskik oor goeie lineêre eienskappe oor 'n sekere meetgebied. Die bandwydte van die induktiewe sensor was ontwerp om tien maal hoër as die AML se bandwydte te wees. Die bandwydte van die sensor moet dus hoër as 500 Hz wees. In die geval van hoër rotasiesnelhede mag die bandwydte van die sensor te laag wees om die rotor om die geometriese middelpunt te suspendeer.

Die self-waarnemende tegniek maak gebruik van die stroom in die spoel asook die spanning oor die kragversterker. Die helling en die amplitude van die hoë frekwensie stroomkomponent verander soos die luggaping verander. Die veranderende stroom word gedemoduleer om die posisie van die rotor te bepaal. Deur van die self-waarnemende tegniek gebruik te maak, kan al die verlangde seine buite die AML gemeet word. Die voordeel van eksterne metings is dat daar geen sensorkomponente binne die AML teenwoordig is nie.

Die self-waarnemende sensor is suksesvol geïmplementeer op 'n heteropolêre AML wat van skakelmodus kragversterkers gebruik maak. Die bandwydte van die self-waarnemende sensor is hoog genoeg om steurings tot en met 100 Hz te oorkom. Die sensor se uitvoersein bevat 'n 100 Hz ruiskomponent wat afkomstig is van die kragversterker se kapasitor. Die lineariteit van die sensor is beperk deur werwelstroomverliese in die AML.

Toekomstige werk sal slegs op die self-waarnemende sensor fokus. Die werwelstroomverliese kan geminimeer word deur dunner gelamineerde plaatjies te gebruik in die magnetiese baan. Die verlaging van die werwelstroomverliese sal die lineêriteit van die self-waarnemende sensor verbeter. Deur 'n differensiële sensor te implementeer, waar die luggapings aanbeidekante van die rotor geskat word, en die verskil bepaal word kan 'n meer sensitiewe stelsel met 'n verlaagde ruisvlak verkry word. Implementering met digitale seinverwerking sal 'n belangrike rol speel in toekomstige werk.

Die projek plaas 'n stewige basis in sensortegnologie in die navorsings groep.

Die Here is my herder, ek kom niks kort nie. Hy laat my in groen weivelde rus. Hy bring my by waters waar daar vrede is. Hy gee my nuwe krag. Hy lei my op die regte paaie tot die eer van sy Naam – ***Psalm 23:1-3***

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LIST OF ABBREVIATIONS

ac	Alternating current
ADC	Analog to digital converter
AM	Amplitude modulation
AMB	Active Magnetic Bearing
BW	Bandwidth
CAD	Computer Aided Design
CCD	Charge-coupled device
dc	Direct current
DAC	Digital to analog conveter
EMF	Electromotive force
FPF	Forward path filter
FWR	Full wave rectifier

HFVIM	High frequency voltage injection method
IC	Integrated circuit
LPF	Low-pass filter
LED	Light emitting diode
MOSFET	Metal-oxide semiconductor field-effect transistor
PBMR	Pebble Bed Modular Reactor
PWM	Pulse Width Modulation
rpm	Revolutions per minute

LIST OF SYMBOLS

A_g	Air gap area
B	Magnetic flux density
C	Capacitance
C_0	Nominal capacitance
d	Nominal air gap
g, g_0	Air gap length
H	Magnetic field intensity
i	Instantaneous current
i_F	Fundamental current
K_I	Integral gain of the PI controller
K_P	Proportional gain of the PI controller
ℓ	Magnetic path length
L	Inductance
L_d	Nominal inductance
M	Mutual inductance
N	Number of coil turns
R	Electrical resistance
x	Change in air gap

Introduction

The aim of this project is to develop sensors for Active Magnetic Bearing (AMB) systems. The project was initiated by the Pebble Bed Modular Reactor (PBMR) project to establish a knowledgebase in the field of position sensor technology for AMBs. This chapter discusses the background on AMB sensors, the purpose of the research, issues to be addressed and the research methodology. An overview of the dissertation is also given.

1.1 AMBs and the PBMR

The PBMR is a helium-cooled reactor that uses the Brayton thermodynamic gas cycle to convert nuclear energy into electrical energy. The nuclear energy is generated in the reactor core by nuclear fusion. Helium gas will be used to transfer the energy from the reactor to a turbo-generator unit that will generate electrical power [1]. Figure 1.1 gives a schematic diagram of the PBMR system layout. The fundamental concept of the design is aimed at achieving a plant that has no physical process that could cause a radiation hazard beyond its site boundary. This means that all possible sources of nuclear contamination to the environment must be eliminated.

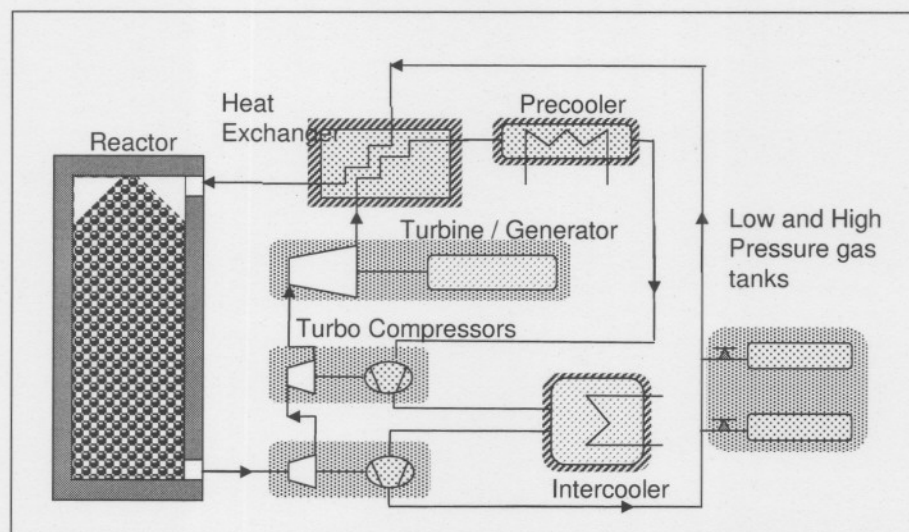


Figure 1.1: PBMR model

To prevent nuclear contamination to the environment helium gas is used as coolant since it is chemically and radiologically inert. If for some reason the helium would escape from the system into the atmosphere it will not hold a contamination risk for the environment. The only other possible source of nuclear contamination in the gas cycle would be oil from the oil film bearings of the compressors, turbines and generator.

Active magnetic bearings separate all contact between the rotating and stationary parts. This is done by suspending the rotor in mid air [2]. This contactless suspension eliminates the use of lubricants and there are thus no contamination risks.

The name active magnetic bearing means that the magnetic bearing is actively controlled. All actively controlled systems need some feedback [2]. In the case of AMB systems the position of the rotor is the feedback variable. Figure 1.2 illustrates the operation of an AMB. The vertical sensor measures the vertical position of the rotor. This position signal is subtracted from the reference position. This error is the input to the controller, which determines the amount of current delivered to the coils by the power amplifier.

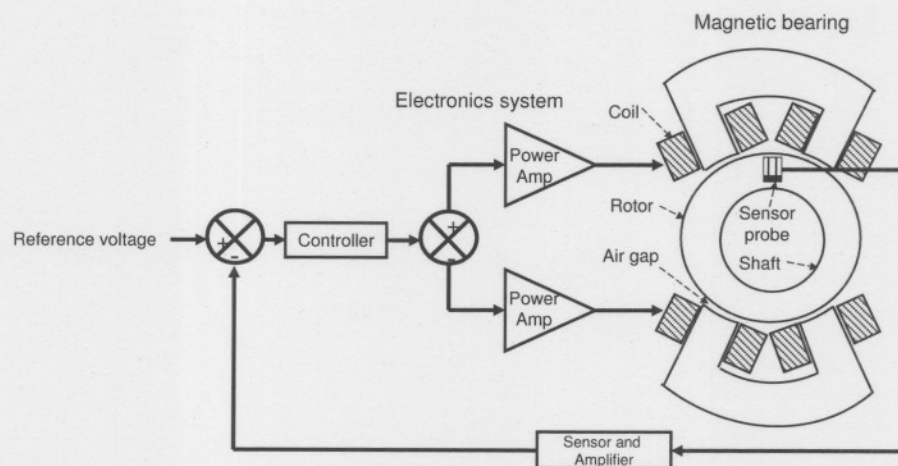


Figure 1.2: Active magnetic bearing system

The sensors are the main components affecting the controllability and stability of the AMB. If the sensor does not perform according to specification, the reliability and stability of the AMB will be greatly affected. The measurement range, linearity, sensitivity, resolution, frequency range, noise and temperature variations determine the performance of sensors [2]. These sensor properties will now be defined.

Measurement range

The output of a position sensor changes according to the distance [2]. The maximum useable distance, which gives a meaningful output and is in the linearity range of the sensor, is called the measurement range. The linear range is usually much smaller than the physical one. The measurement range of a sensor differs according to the type of topology used for the specific sensor.

Linearity

Sensors are rarely completely linear, but can normally be approximated as a linear across a percentage of the measurement range [2]. The deviation between the sensor output and the real distance determines the linearity of the sensor. The smaller the deviation, the more linear the sensor is.

Sensitivity

The sensitivity indicates the ratio of the output signal to the quantity to be measured [2]. The sensitivity of displacement sensors are measured in mV/ μm . The relatively small output of a sensor could be enhanced by an electronic amplifier circuit for further use.

Resolution

The resolution is defined as the minimum useful information that can be used where noise is present [2]. In the case of the displacement sensor, it is the peak value which can be distinguished from noise. The resolution is mostly indicated in absolute values, for instance μm . The resolution of the sensor could be increased by using a low-pass filter, but with the disadvantage of decreasing the frequency range. All external disturbances could drastically reduce the resolution.

Frequency range

Position sensors must operate linearly across a wide frequency range. This means that the output of the sensor must not be affected by a fast changing rotor [2].

This is an essential characteristic of the sensors, because in active magnetic bearings the frequency will change according to the speed of rotation of the rotor. The frequency at a 3 dB reduction in the sensor response is considered as the cut-off frequency, because a significant phase lag could be assumed.

Noise

Noise is the most undesirable component, but the most common component in sensors. Noise has two sources: 1) internal and 2) external. Great care must be taken to prevent noise from dominating the design. These noise problems can be overcome by good design and the use of filter techniques.

Internal noise

Internal noise is the non-ideal response of electronic components of the sensor itself. Sources of internal noise are the increase of temperature, the non-ideal PN junction of semiconductors and the ageing effect. Internal noise could be reduced by a good design.

External noise

External noise can be generated by almost anything [2], in our case it could even be the sun or dust. Figure 1.3 illustrated the effect on an optical sensor if dust is placed on the measured surface or if the surface is damaged.

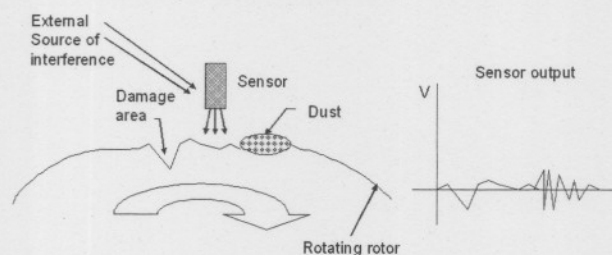


Figure 1.3: Rotor surface noise

Infrared optical sensors will be sensitive to sunrays, thus if the sensor is not isolated all the sensors will have great offset problems. In PBMR the greatest concern will be the extreme operating conditions and the exposure to large magnetic fields generated by the AMB.

Temperature variations

Changes in temperature lead to component drift [3]. When the components drift from their normal operating point due to the change of temperature, the reference point of the rotor will also change. These reference problems could lead to disaster if not compensated for.

The problem can be overcome by using the sensors in a difference layout as shown in Figure 1.4.

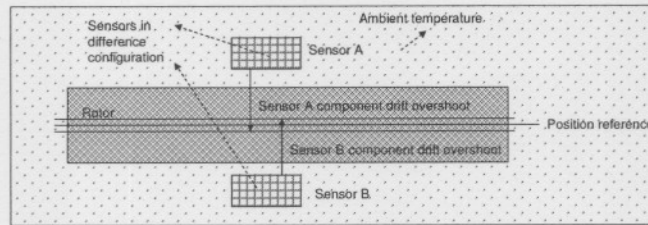


Figure 1.4: Temperature drift compensation

Assume that the drift of the sensors are only dependant on the ambient temperature and the sensors are identical. By subtracting the sensors outputs, the original reference point is found again. Sensors in the difference pair configuration overcome the temperature drift.

1.2 Problem statement

The block diagram in Figure 1.5 describes the problem statement at hand. A broad background study on position sensors for AMB applications will be done. The barrier optical sensor, inductive sensor and PWM-based self-sensing sensor will be developed. These sensors must be implemented and evaluated on AMB models. These working sensor models will increase the knowledge on AMBs sensors and establish a field of sensor technology in the Faculty of Engineering.

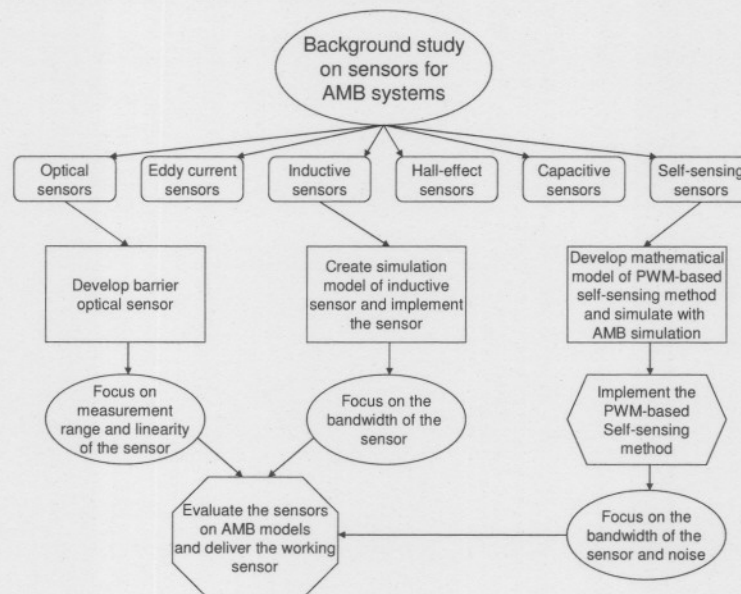


Figure 1.5: Problem statement block diagram

1.3 Issues to address and methodology

This project will be conducted in six phases: 1) Background study of position sensors, 2) barrier optical sensor design and implementation, 3) inductive sensor design and implementation, 4) self-sensing modelling, 5) self-sensing hardware implementation and 6) self-sensing hardware evaluation.

Background study

A comprehensive study on position sensors will be done. The basic types of position sensors will be investigated. The sensor properties described will be taken into consideration when the sensors are developed. The evaluation will also be based on these sensor properties.

Optical sensor design and implementation

The type of optical sensor topology must be chosen, and then this topology has to be simulated and developed. The sensor must meet the following specifications: 1) a linear measurement range of 8 mm and 2) the phototransistor and LED must be at least 5 cm apart. The sensor's performance must then be evaluated using the active magnetic bearing demonstrator.

Inductive sensor

An inductive sensor must be developed for a homopolar radial active magnetic bearing. The requirement for the sensor is a linear range of 1 mm. The bandwidth of the sensor must be larger than 500 Hz. After development the sensor must be evaluated using the homopolar radial AMB system.

Self-sensing modelling

The self-sensing method must be chosen. A mathematical model of the sensor must be developed. This model must be simulated in conjunction with the AMB model in Matlab[®]. This model will determine whether the sensor is suitable for use in AMB applications.

Self-sensing hardware implementation

The hardware implementation of the self-sensing sensor will be derived from the mathematical model. Small compromises have to be made to incorporate hardware imperfections. All the components are implemented with analog electronics.

Self-sensing hardware evaluation

This sensor will be evaluated using the heteropolar radial AMB. This evaluation will determine the performance of the sensor. This will also determine future study on self-sensing AMBs. The bandwidth and the signal to noise ratio of the sensor must be investigated.

1.4 Overview of the dissertation

The background study on different sensors is discussed in chapter 2. The basic model of each topology will shortly be explained. Some of these models will be implemented and evaluated.

In chapter 3 the optical sensor is designed and evaluated. The barrier topology will be used. This chapter consists of the design of the sensor, as well as the implementation and evaluation thereof. The sensor makes use of an infrared photo transistor and LEDs. The geometrical layout of the LEDs and photo transistors is also investigated.

In chapter 4 the inductive sensor is designed and evaluated. This sensor constitutes: 1) inductive coils which form the probes of the sensor, 2) an oscillator and 3) the signal demodulator. All these components are discussed and implemented. This sensor is evaluated using the radial AMB model.

Chapter 5 discusses the mathematical model and simulation of the self-sensing model in Matlab[®] and chapter 6 describes the hardware implementation and integration of the sensor model.

Chapter 7 discusses the evaluation of the self-sensing method. A step response, as well as a sinusoidal response, is given to the system. These tests evaluate the performance of the self-sensing sensor. From these results it was found that the self-sensing sensor output is extremely noisy.

The noise was found to be a 100 Hz component. After intense probing the power amplifier was identified as the source of noise. The sensor was implemented on a heteropolar AMB. The self-sensing sensor was used to close the control loop, but the self-sensing power amplifier was driven by a bias current reference only and not by the control reference. With this setup the system was moderately stable.

Chapter 8 discusses future work. The future work will be focused on the self-sensing sensor. The main action is to reduce the noise produced by the power amplifier. Other methods of self-sensing will also be investigated and developed.

Literature Study

Different types of sensor technology are available. The following sensors will be discussed: 1) optical sensors, 2) inductive sensors, 3) eddy current sensors, 4) Hall Effect sensors, 5) capacitive sensors and 6) self-sensing sensors. The knowledge gained in this chapter, will determine the best sensor for a specific AMB application.

2.1 Optical sensors

An optical sensor relies on light intensity. An LED is the light source and the emitted light is sensed by a photo sensitive sensor. The intensity of the light is directly proportional to the position of the rotor.

Different types of optical sensors are available, each with its own advantages and disadvantages [2]. The following sensors will be discussed:

- 1) light barrier principle,
- 2) reflection principle, and the
- 3) charge-coupled device (CCD) sensor.

2.1.1 Barrier topology

The barrier topology [2] relies on the obstruction of the light source. The rotor of the AMB is the medium that blocks a percentage of the light. Figure 2.1 shows the layout of the barrier topology.

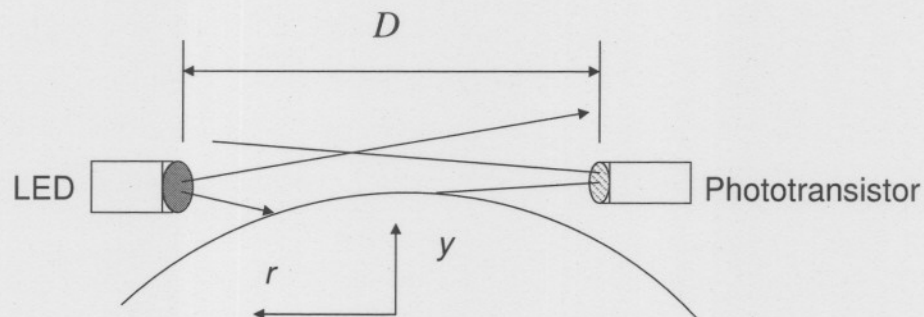


Figure 2.1: Barrier principle optical sensor

The infrared LED illuminates infrared light and the rotor blocks a percentage of the infrared light. If the rotor moves upwards, the light will decrease and the sensor output will decrease.

If the rotor moves downwards, the infrared light will increase and the sensor output voltage will increase accordingly. By choosing a compatible optical transmitter and receiver an excellent linear relation can be obtained between the rotor's position and the measured voltage. The relation between the distance D and the type of transmitter lens plays a critical part in the sensitivity, linearity and measurement range of the sensor.

2.1.2 Reflection topology

The reflection sensor relies on the reflection properties of the rotor. The reflection optical sensor also depends on the light emission [2]. When the distance x changes, the sensor output will also change accordingly. In this case the reflection incident of the rotor material must be a known factor. Anything that will alter the incident of reflection will be defined as noise, thus even a thin layer of dust will have an enormous effect on the linearity of the sensor. This sensor topology is limited to clean operation conditions. Figure 2.2 shows the reflection topology.

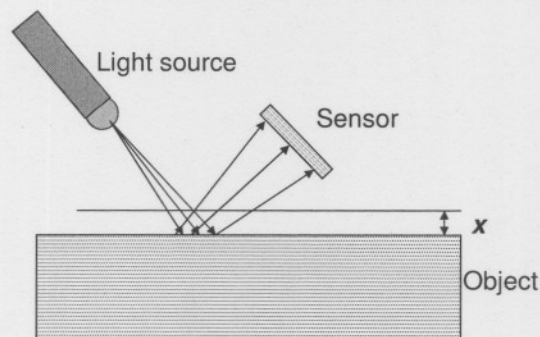


Figure 2.2: Reflection optical sensor

2.1.3 Linear camera topology

The position of the rotor could be determined by using a charge-coupled device (CCD) camera [2], [4]. Figure 2.3 shows a typical setup of a CCD sensor. The CCD camera consists of a linear array. Each of the array elements is light sensitive, thus an increase or decrease of light will result in a linear digital output. This sensor is also very susceptible to dust, and is also limited to clean operation conditions.

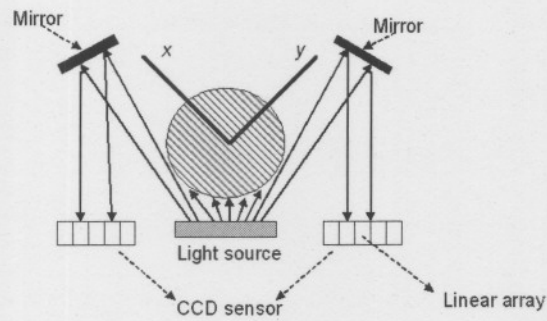


Figure 2.3: Linear CCD sensor

2.1.4 Advantages of optical sensors

Optical sensors have the advantage that they do not suffer from magnetic disturbances. Thus the sensor could be placed near the air gap of the AMB's coils and rotor. This placement of the sensor has the benefit of increased accuracy and minimizes noncollocation effects. Optical sensors also support excellent linearity. The bandwidth of the sensor is extremely high, thus it will exceed the bandwidth of the amplifier and other components. The simplicity of the sensor circuitry for general uses makes it a relatively low cost sensor, but when high quality measurements are required the sensors become expensive.

2.2 Inductive sensors

The inductive sensor is based on the variation of inductance as a function of the air gap [5],[6]. If two series coils are placed close to the rotor as shown in Figure 2.4, the inductance of the coils will change according to the distance from the rotor.

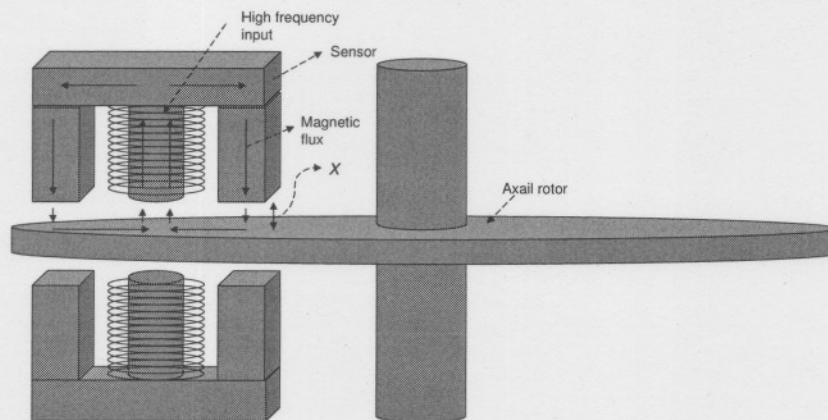


Figure 2.4: Inductive sensor application

The two coils are activated by a sine wave. The amplitude and the frequency of the sine wave are kept constant. The distance of the rotor changes continuously over time when the rotor is in operation, thus the reactance also changes continuously.

The inductance of the coil changes as the distance changes (2.1).

$$L = L_d \frac{d}{d + x} \quad (2.1)$$

where d is the nominal air gap and x is the change in air gap from the centre position.

L_d is the nominal inductance where x is zero.

Thus the distance is proportional to the inductance. The change in the inductance between the two coils, illustrated in Figure 2.5, leads to a changing voltage amplitude proportional to the air gap. This voltage is measured at the centre tap of the two series coils

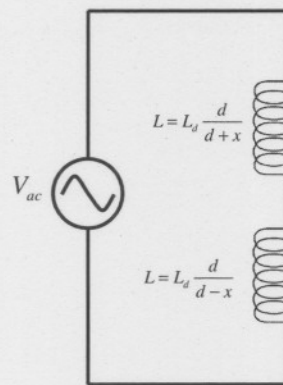


Figure 2.5: Changing inductance of series coils

Inductive sensors are very sensitive to external magnetic noise and in the case of AMBs, the sensor must be placed near the rotor where the exposure is the highest. The position of the sensor must be carefully considered and proper shielding will be inevitable. The rotor must consist of ferromagnetic material in the region where the sensor is to be placed.

The sensor has good linearity, but the bandwidth of the sensor is limited by the eddy current effect. Due to the eddy current effect, the conductivity of the rotor at high frequencies, delays the penetration of the flux generated by the sensor's coil, with a consequent alteration of the output signal.

2.3 Capacitive sensor

The capacitive sensor is based on the change in the gap between the two plates.

One plate (probe) is fixed and the other plate (target) is connected to the object to be positioned.

Since the plate size and the dielectric stays the same, the output of the sensor is proportional to the distance between the plates. By changing the position of the rotor the air gap changes and the sensor output will change accordingly [2]. Figure 2.6 shows a cylindrical capacitive sensor [7], [8].

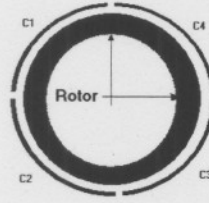


Figure 2.6: Cylindrical capacitive sensor

For simplicity the parallel plate capacitor in Figure 2.7 will be analysed [9].

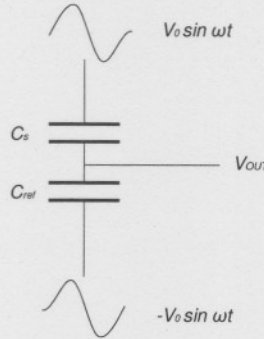


Figure 2.7: Parallel plate capacitive sensor

$$\frac{\partial C}{\partial x} = -\frac{C}{x_0 + x} \quad (2.2)$$

$$C_s = \frac{\epsilon_0 A}{x_0 + x} \approx C_0 - \frac{\partial C}{\partial x} x \quad (2.3)$$

where C_0 is the capacitance where $x = 0$ and $x \ll x_0$

$$V_{Out} \approx \frac{V_0}{2} (FM \times x \frac{C_0 - C_{ref}}{C_0}) \quad (2.4)$$

$$FM = -\frac{1}{x_0 + x} \quad (2.5)$$

where FM is the negative inverse proportionality of the distance.

The sensor is sensitive to switching noise. Another disadvantage is the small standoff distance. The bandwidth of the capacitive sensor is very high.

2.4 Hall Effect sensors

The Hall Effect sensor works on the changing flux of a magnet [10], [11]. The flux is inversely proportional to the distance. When current flows along a thin, band-shaped conductor and it lies perpendicular to a magnetic field, forces will act perpendicular to the band on the electrons which move at a drifting speed of v along the conductor. This has the result of positive and negative charges on both longitudinal sides of the band. The basic physical principle underlying the Hall Effect is the Lorentz force. When an electron moves along a direction perpendicular to an applied magnetic field, it experiences a force acting normal to both directions and moves in response to this force and the force affected by the internal electric field. For an n -type, bar-shaped semiconductor shown in Figure 2.8, the carriers are predominately electrons.

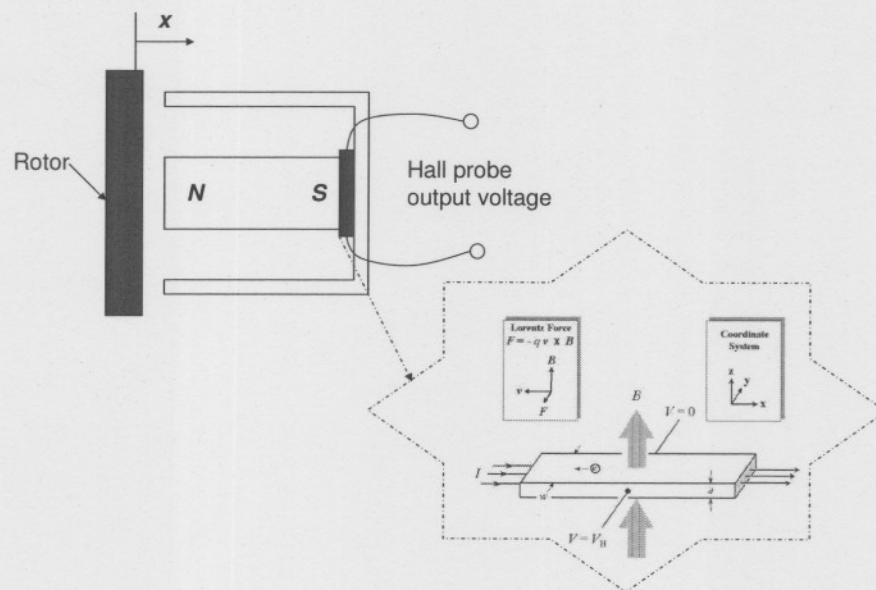


Figure 2.8: Hall Effect sensor

We assume that a constant current I flows along the x -axis from left to right in the presence of a z -directed magnetic field. Electrons subjected to the Lorentz force initially drift away from the current line towards the negative y -axis, resulting in an excess

surface electrical charge on the side of the sample. This charge results in the Hall voltage, a potential drop across the two sides of the sample. This transverse voltage is the Hall voltage V_H and its magnitude is equal to IB/qnd , where I is the current, B is the magnetic field, d is the sample thickness, and q (1.602×10^{-19} C) is the elementary charge.

These charges that are generated could be used for a sensing voltage. The sensor's output is dependent on the changing flux, which is proportional to the distance. The Hall Effect sensor has moderate linearity. The cost of the sensor is relatively low [12].

2.5 Eddy current sensor

As shown in Figure 2.9 when an alternating current of a constant frequency f is passed through a coil, then an alternating electromagnetic field will be produced around the coil. This electromagnetic field is called the primary magnetic field H_1 . When a metallic target is moved into the field, an eddy current is produced in the surface of the target. A secondary magnetic field H_2 is generated by the eddy current, which is opposite to the coil's field. The impedance of the coil changes when the target is placed in this field. The closer the target is to the coil the larger the change in the impedance and the mutual inductance M between the target and the coil [13].

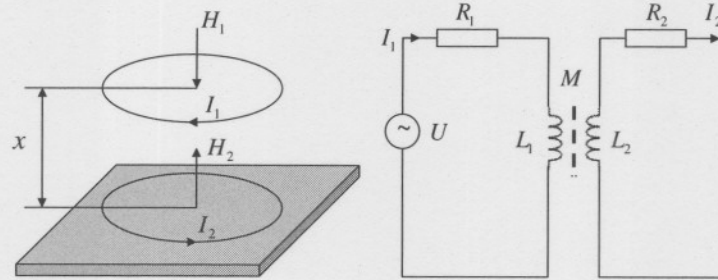


Figure 2.9: Principle of eddy currents

If x is the distance between the target and the coil, then x is proportional to the mutual inductance M . From Figure 2.9:

$$R_1 I_1 + j2\pi f L_1 I_1 - j2\pi f M I_2 = U \quad (2.6)$$

$$R_2 I_2 + j2\pi f L_2 I_2 - j2\pi f M I_1 = 0 \quad (2.7)$$

From (2.6) and (2.7) I_1 is calculated as follows:

$$I_1 = \frac{U}{[R_1 + 4\pi^2 f^2 R_2 / (R_2^2 + (2\pi f L_2)^2)] + j2\pi f [L_1 - 4\pi^2 f^2 L_2 / (R_2^2 + (2\pi f L_2)^2)]} \quad (2.8)$$

The impedance of the coil is as follows:

$$z = [R_1 + 4\pi^2 f^2 R_2 / (R_2^2 + (2\pi f L_2)^2)] + j2\pi f [L_1 - 4\pi^2 f^2 L_2 / (R_2^2 + (2\pi f L_2)^2)] \quad (2.9)$$

The inductance of the sensor becomes:

$$L = L_1 - 4\pi^2 f^2 M^2 L_2 / (R_2^2 + (2\pi f L_2)^2) \quad (2.10)$$

From (2.10) it is clear that the inductance is a function of the distance between the coil and the target. These sensors are frequently used in AMBs. They have good linearity and have a relatively large measurement range.

2.6 Self-sensing magnetic bearing

In a nuclear fusion reactor, some kind of diagnostic system should be implemented to control the rotor of the turbine. Since the sensors are used under severe irradiation circumstances, the number of diagnostic sensors is desired to be small or zero. Therefore, it is important to develop a sensorless or self-sensing control system in which there are no sensors of the controlled object and only monitors of the actuator.

High frequency voltage injection method

The first method is the high frequency injection method (HFVIM) [14]. This sensor works on the same principle as the inductive sensor, but there is no coil present.

The bearing's coils are used for sensing and controlling of the rotor. An additional high frequency component is superimposed on the control voltage. This control voltage imposes a high frequency current component through the coil. This high frequency current component contains a modulated position signal. The position of the rotor is extracted by demodulating the current. Figure 2.10 shows the HFVIM block diagram.

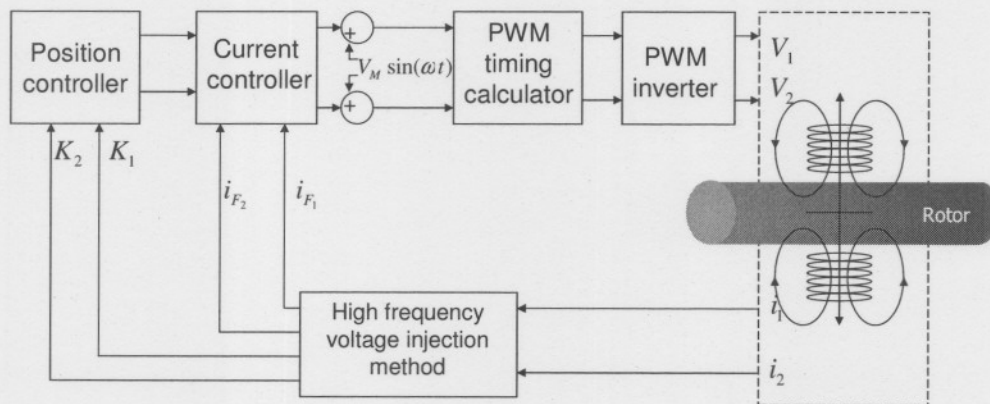


Figure 2.10: Block diagram of HFVIM

The HFVIM block in Figure 2.10 is expanded to the sub-block diagram in Figure 2.11, which illustrates the demodulation method.

The high frequency component is passed through a band-pass filter (BPF). This high frequency component is subtracted from the input current. The result is the real current without the high frequency component, which is the reference current to the amplifier.

The band-passed current signal is AM modulated in terms of the changing air gap. Demodulation of the signal is obtained by multiplying this signal by $-2 \cos \omega t$ and then passing it through a low-pass filter. The output of the HFVIM is proportional to the air gap.

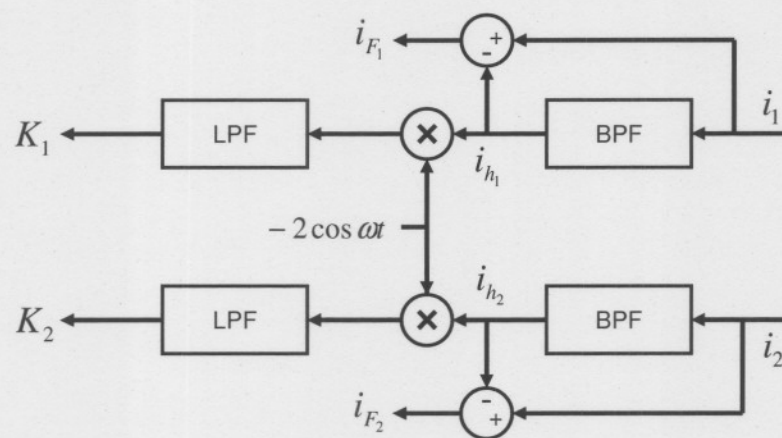


Figure 2.11: HFVIM position demodulation

PWM self-sensing method

The second method is to sense the current response of the switching amplifier [15]. The slope and amplitude of the current changes according to the air gap. This changing amplitude could then be used to determine the position of the rotor. Figure 2.12 shows the block diagram of a PWM self-sensing method.

Voltage and current measurements are taken externally from the coils of the AMB. These voltages and currents are inputs to the self-sensing sensor. The current signal is passed through a forward path filter (FPF). This filter constitutes of 1) a high-pass filter, 2) a full wave rectifier (FWR) and 3) a low-pass filter. The output of the estimator is subtracted from the FPF output. The estimator constitutes of 1) an analog multiplier 2) a non-ideal integrator and 3) a FPF. The difference between the estimator and FPF output is passed through a PI controller. The output of the PI is the output of the sensor. This method will be discussed in great detail in chapter 5, chapter 6 and chapter 7.

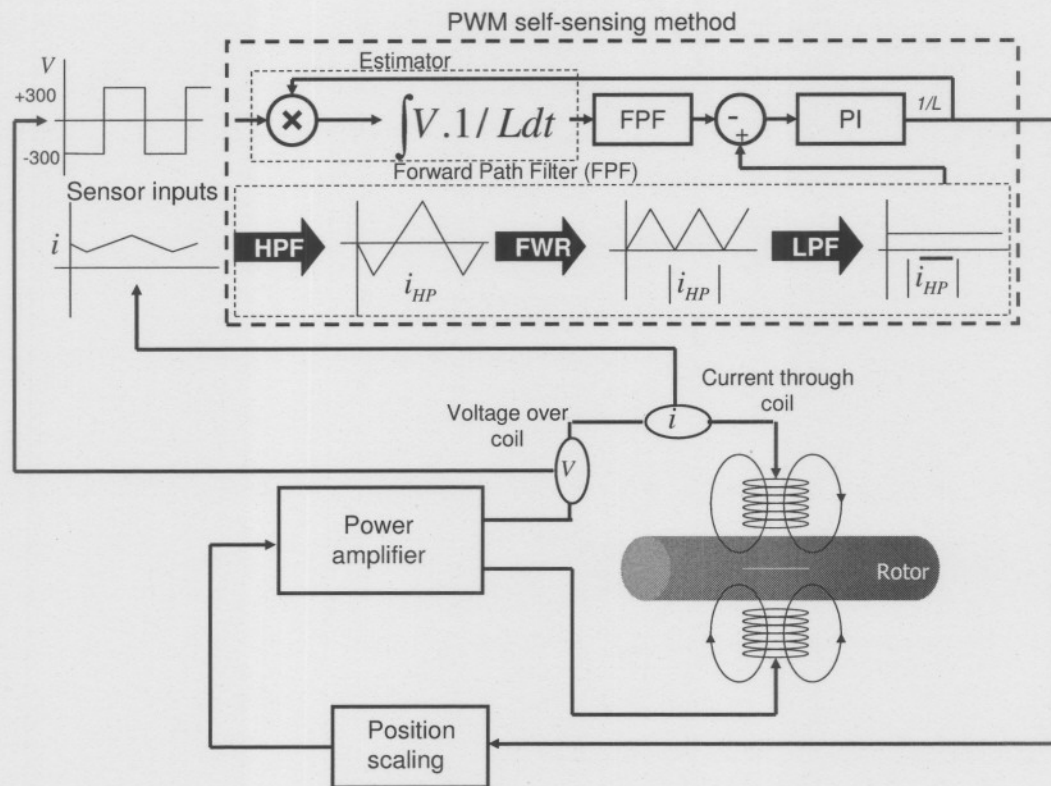


Figure 2.12: PWM self-sensing method

Advantages of self-sensing methods

Self-sensing methods have the benefit of removing most of the costs and potential sensor failures. Other potential advantages of self-sensing methods are that they use amplifier switching noise and eliminate the effect of sensor-actuator noncollocation [16]. Noncollocation is the placement of the sensor with regards to the centre position of the stator. In the case of contactless sensors with probes, the probes cannot be placed in the centre of the AMB coil, thus a small error is made when the position is measured.

2.7 Conclusion

From the literature it was found that different sensors are used in different application areas. In extreme operating conditions, where the temperature is extremely high, the self-sensing method is more appealing. The reason for this is that the self-sensing sensor is situated outside the AMB, thus the integrity of the AMB is increased. The self-sensing method reduces maintenance and wiring cost. The sensor itself is also not as expensive compared to the other sensors mentioned.

Optical Sensor

In this chapter the barrier optical sensor will be discussed. The sensor was shortly introduced in chapter 2. In this chapter it will be analysed and developed. The sensor is constructed from general parts, but in the design the following sensor properties and characteristics must be taken into consideration 1) measurement range, 2) light intensity, 3) the placement of the phototransistor and infrared LED and 4) the characteristics of the phototransistor and infrared LED.

3.1 Basic layout

The barrier type sensor relies on the obstruction of the light source [2]. The ball or rotor is the medium that blocks a percentage of the light. Figure 3.1 shows how the phototransistor illuminates infrared light and how the ball blocks a percentage of the light.

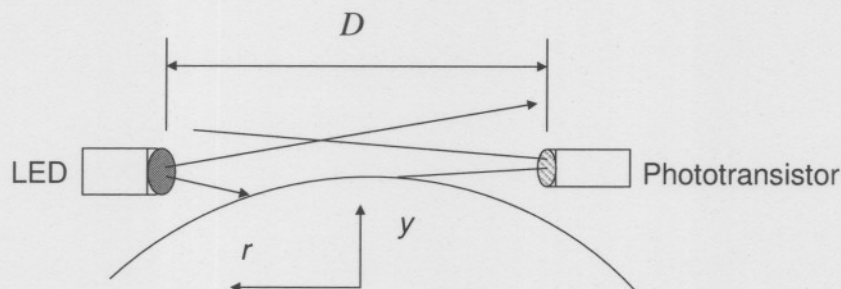


Figure 3.1: Barrier principle optical sensor

When the ball moves upwards the light will decrease and the sensor output will decrease.

3.2 Advantages and disadvantages

By choosing a compatible infrared optical transistor and LED an excellent linear relation can be obtained between the rotor's position and the measurement range. The bandwidth of the sensor is high compared to other sensors. High bandwidth is ideal in high vibration or fast rotating applications like AMBs. The sensor is sensitive to dust as this will decrease the performance of the sensor.

Other infrared sources also influence the sensor's performance. If the sensor is subjected to a constant infrared source, the sensor will have a dc offset. If the source varies over time the sensor's output will also vary.

3.3 Electromagnetic spectrum

Figure 3.2 shows the electromagnetic spectrum. In the design of the infrared optical sensor the wavelengths in the vicinity of infrared light must be taken into consideration.

If Figure 3.2 is inspected closely, it is obvious that red light is the nearest adjacent wavelength to infrared light, thus red light is also a source of noise.

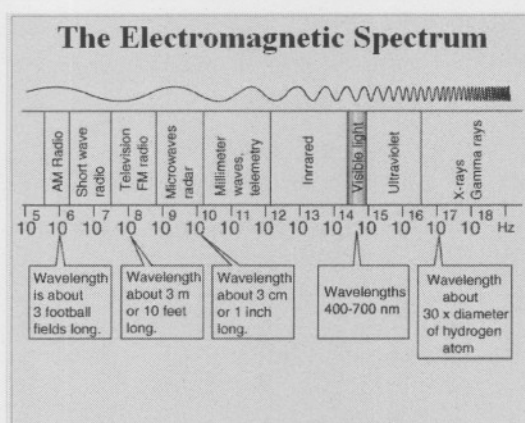


Figure 3.2: Electromagnetic spectrum

From the observation above the sensor must compensate for these adjacent wavelengths.

3.4 Design phase

The block diagram in Figure 3.3 illustrates the basic principle of operation of the optical sensor. The cost and the availability of infrared LEDs and phototransistors make it the most intriguing choice. The model developed is not of high accuracy, thus low cost.

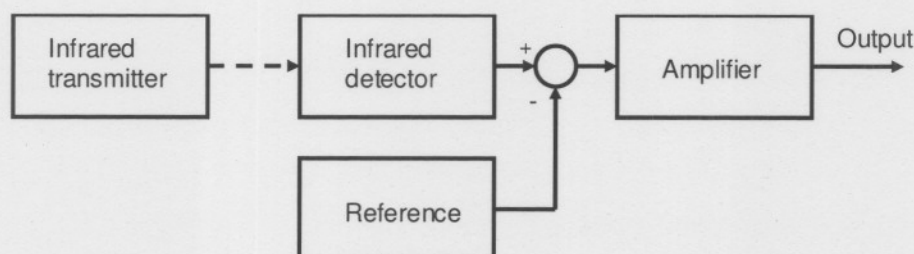


Figure 3.3: Basic Infrared sensor block diagram

The first block shows the infrared source which transmits the light to the detector. The infrared source consists of an infrared LED which is operated in the vicinity of its maximum operating point. This will ensure that the noise generated by other sources is relatively small compared to the emitter. This high operating point ensures a high signal to noise ratio.

The second block consists of an infrared detector. The infrared detector consists of an infrared phototransistor. The phototransistor is the vital component of the optical sensor. The response time and the linearity of the phototransistor determine the bandwidth and linearity of the optical sensor. As mentioned before, the intensity of the infrared noise plays an important role, because the infrared detector is very sensitive to red light. This can be seen from the electromagnetic spectrum [17]. Some of these detectors have filters to prevent noise disturbances, like daylight.

The third block determines the point of reference. The amount of infrared light emitted from other sources determines the reference point. This reference voltage is subtracted from the infrared detector to reduce the effects of external infrared sources. This external source of infrared light could be viewed as noise.

The fourth block is the amplifier. The output of the phototransistor is small and is not scaled. When the output of the phototransistor is amplified to a specific value then it is in a more usable format.

3.5 Basic infrared sensor circuit design

The circuit design is based on Figure 3.3. Each component will be developed separately.

3.5.1 Infrared emitter

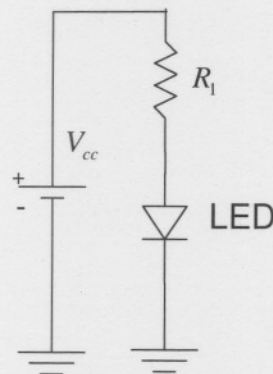


Figure 3.4: Infrared emitter

Figure 3.4 shows the circuitry for the infrared light emitter. The current through the LED determines the emitted light intensity. By using ohms law the current through the LED is determined by using (3.1), where V_f is the voltage drop across the LED.

$$I_D = \frac{V_{cc} - V_f}{R_1} \quad (3.1)$$

By varying resistance R_1 the optimal light intensity could be found. Great care must be taken not to drive the LED over its critical voltage point or it will self-destruct.

Figure 3.5 shows the relationship between the current and the radiation intensity. The continuous current must not be more than 100 mA for the LD274 infrared LED or it will self-destruct. If the infrared light intensity of the LED must be increased and the dc current through the LED is at its maximum operating current (100 mA), then the current through the LED must be pulsed. The current limit of the LED is drastically increased by using this pulse current technique. Figure 3.5 shows that the current could be increased up to 3 A if the pulse time is 20 μ s. The radiant intensity is defined as a function of the current.

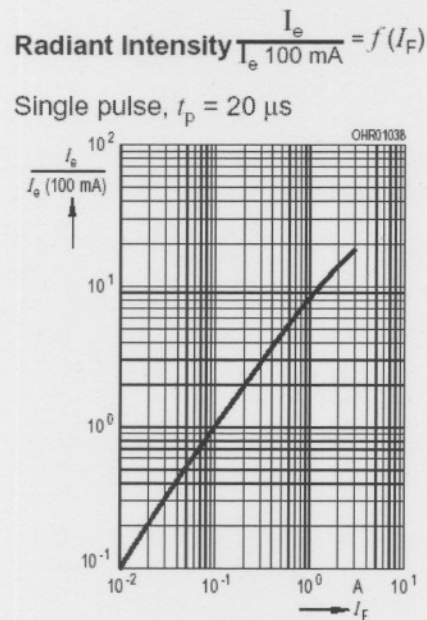


Figure 3.5: Radiant intensity versus current

3.5.2 Infrared detector

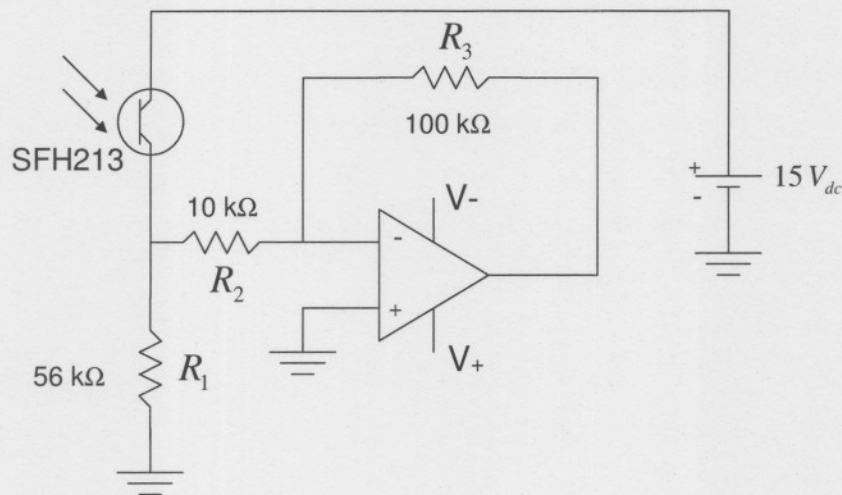


Figure 3.6: Infrared detector

The characteristics of the infrared phototransistor determine the resistor value R_1 . The method works on the change in the current through the resistor R_1 . When the infrared light increases or decreases the voltage across the resistor R_1 will increase or decrease accordingly.

The reference circuit is a duplicate of the circuit in Figure 3.6, but has no infrared LED. The noise is the external infrared sources. In this design the TL082 operational amplifier is used.

An important point to mention is to operate the infrared phototransistor in the linear point, as shown in Figure 3.7 [18]. The phototransistor (SFH213) could be viewed as a bipolar transistor, in which the light adjusts the base current.

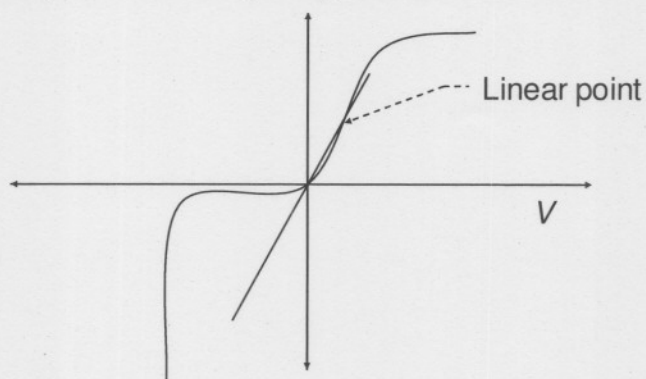


Figure 3.7: Linear point

3.5.3 Inverting amplifier

The inverting amplifier is used to amplify the voltage across R_I in Figure 3.6. The output of the reference circuit is also amplified by the same factor. The inverting amplifier has the ability to amplify the signal with values smaller than one, but has the disadvantage that the signal is inverted.

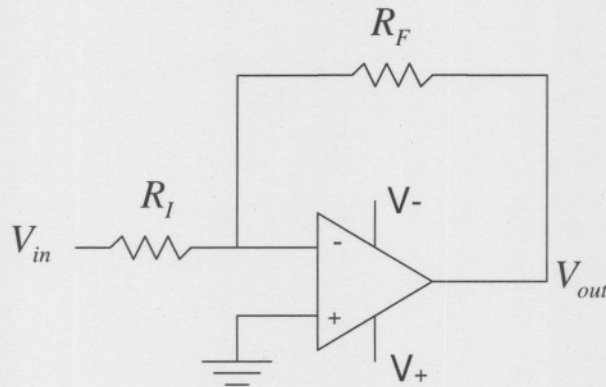


Figure 3.8: Inverting amplifier

Figure 3.8 shows the circuitry of an inverting amplifier [3]. The amplification factor is a function of resistors R_F and R_I as described by (3.2).

$$A = -\frac{R_F}{R_I} \quad (3.2)$$

The right combination of R_F and R_I must be chosen to prevent the saturation of the operational amplifier.

3.5.4 Difference amplifier

Sometimes it is desirable to obtain the difference between two signals. Figure 3.9 shows a difference amplifier [3]. The outputs of the two inverting amplifiers are subtracted from each other. The gain of the difference amplifier could be adjusted as shown in (3.3).

$$V_{out} = \frac{R_f}{R_{in}}(V_2 - V_1) \quad (3.3)$$

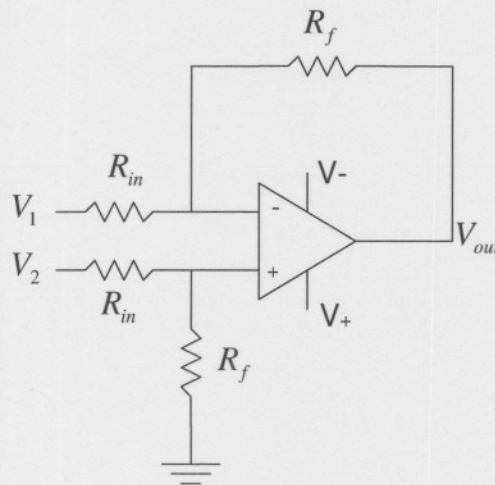


Figure 3.9: Difference amplifier

If V_1 is the phototransistor output and V_2 is the reference voltage, the sensor output will be positive.

3.5.5 Infrared sensor

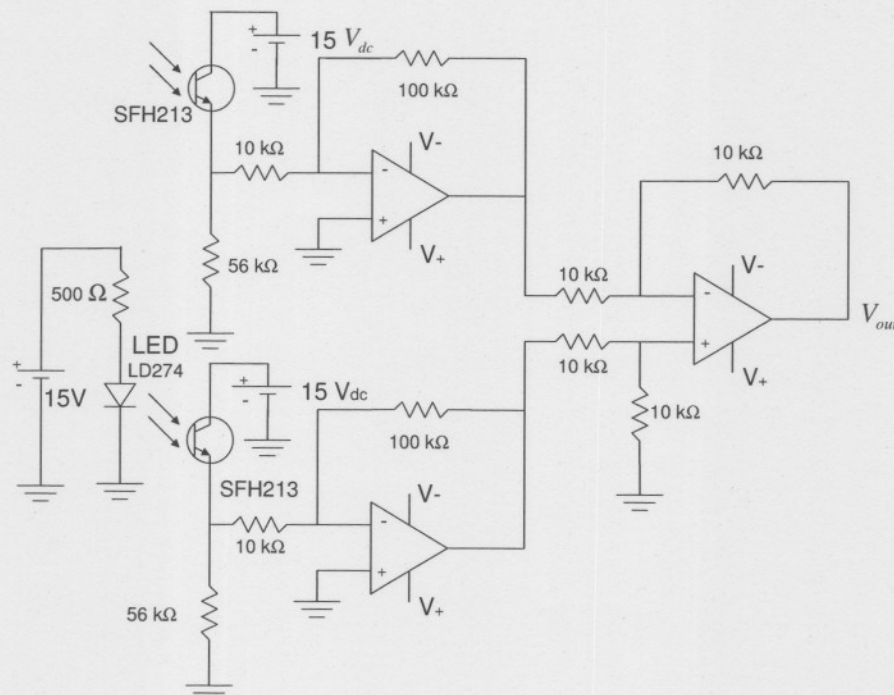


Figure 3.10: Infrared sensor circuit

Figure 3.10 shows the integrated circuit of the optical infrared detector when all the components are combined.

3.6 Linearity

The circuit in Figure 3.10 shows only one infrared LED and phototransistor pair. By using more than one pair of LEDs and phototransistors, the linearity range could be increased. The two LEDs are placed parallel to each other and the same for the phototransistors. The pair placement is shown in Figure 3.11. The phototransistors are electrically connected in parallel. The LEDs are also connected in parallel, but the resistor value R_i in (3.1) is halved. The smaller resistor increases the current, thus the infrared output intensity of the individual LEDs will remain the same.

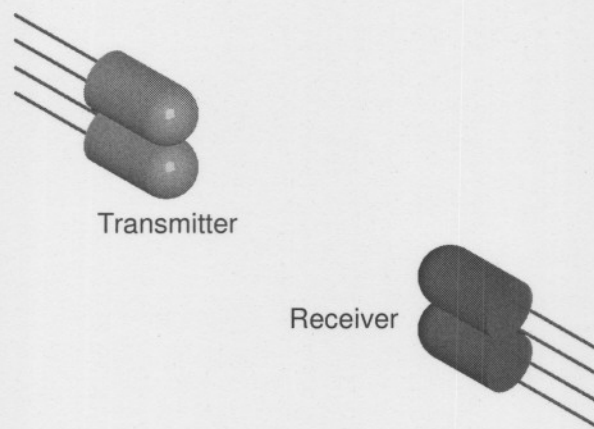


Figure 3.11: Double layer layout

3.7 Results

All the results that follow are based on the sensor in Figure 3.10, but with variations in phototransistor and LED layout. These numerical values are results of practical experiments in the electronic laboratory (with neon lights on, thus with a noise source).

3.7.1 3 mm LED and phototransistor

The following results are found from an optical sensor that uses only one infrared LED and one phototransistor. The dimensions of the LED and the phototransistors are 3 mm in diameter. The practical layout is as shown in Figure 3.12 where the grey instrument is a micro-meter.

The LED and phototransistor are placed 3 cm apart. The shaft of the micro-meter obstructs the infrared light. By turning the dial of the micro-meter the light is measurably obstructed.

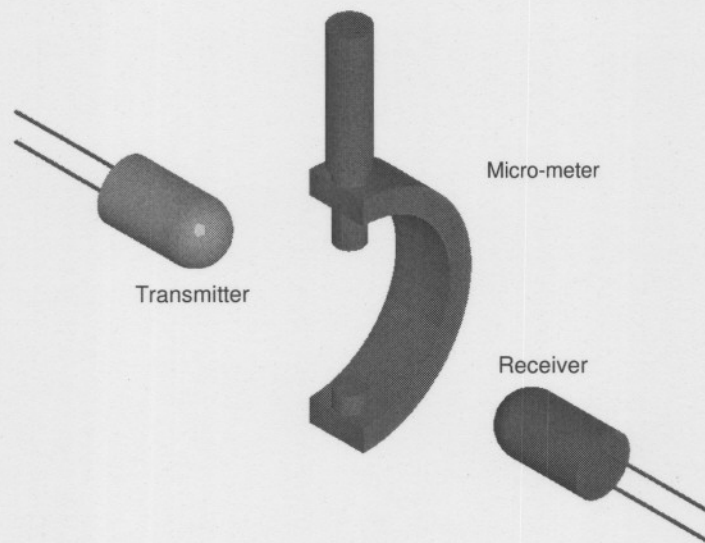


Figure 3.12: Practical test

Figure 3.13 shows the response of the sensor as the micro-meter is adjusted. The usable measurement distance in this case is very small.

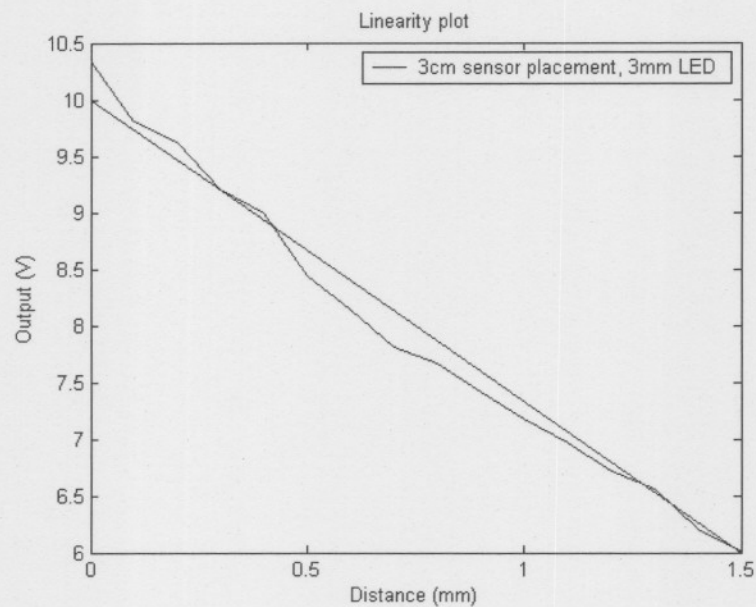


Figure 3.13: 3 mm LED sensor response

Figure 3.14 shows the deviation from the ideal line. The maximum fault in this case is 3 %, that is very large compared to the small measurement distance available.

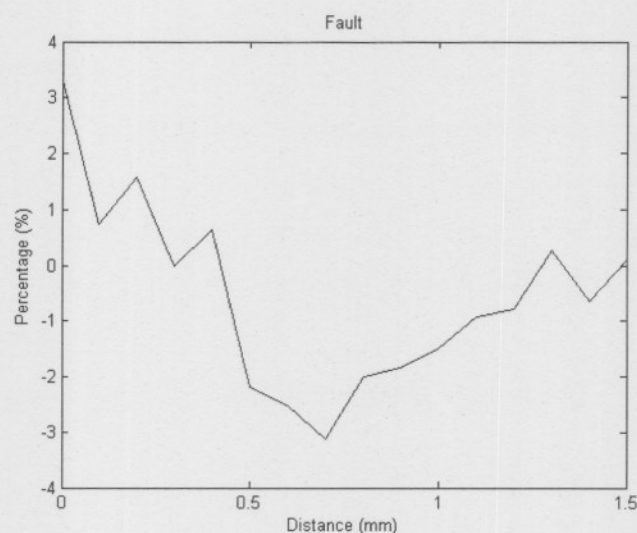


Figure 3.14: Error of 3 mm sensor

3.7.2 Three 5 mm LEDs and two phototransistors

The following results are found from an optical sensor that uses three infrared LEDs and two phototransistors. The dimensions of the LED and the phototransistors are 5 mm diameter. The practical layout is as shown in Figure 3.15.

The test was implemented on the ball of a magnetic bearing demonstrator project. The LEDs and phototransistors are placed 4 cm apart.

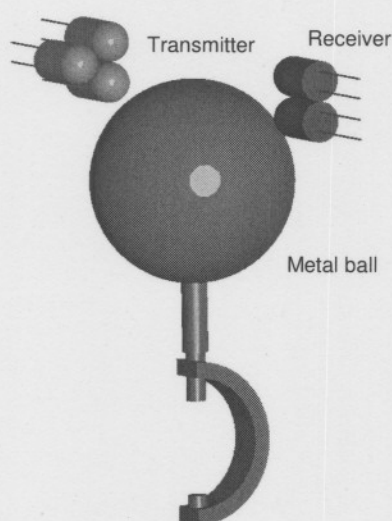


Figure 3.15: Ball setup

Figure 3.16 shows a relatively good response from the ball setup. When the ball was turned the ball wobbled. The ripple in the graph is caused by the three LEDs which projects a 3D component on the two phototransistors, thus the wobble.

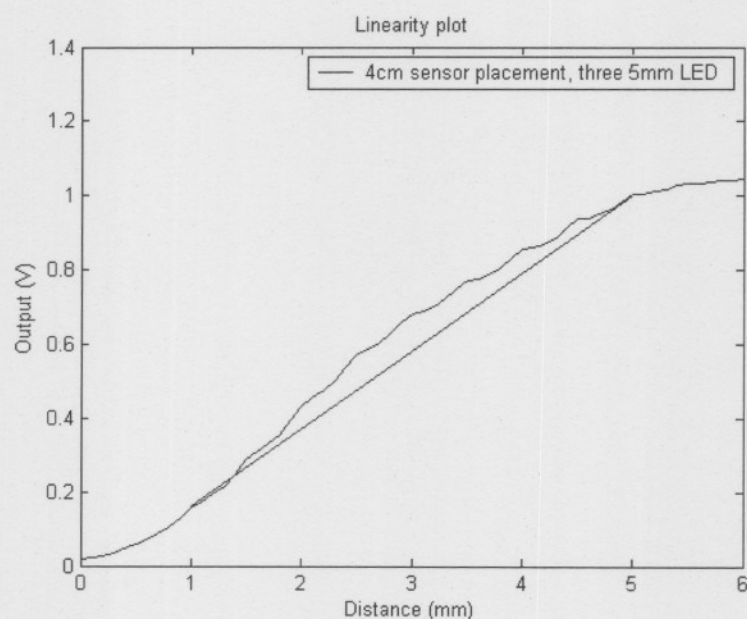


Figure 3.16: Ball setup sensor response

The fault is relatively small over the sensing range.

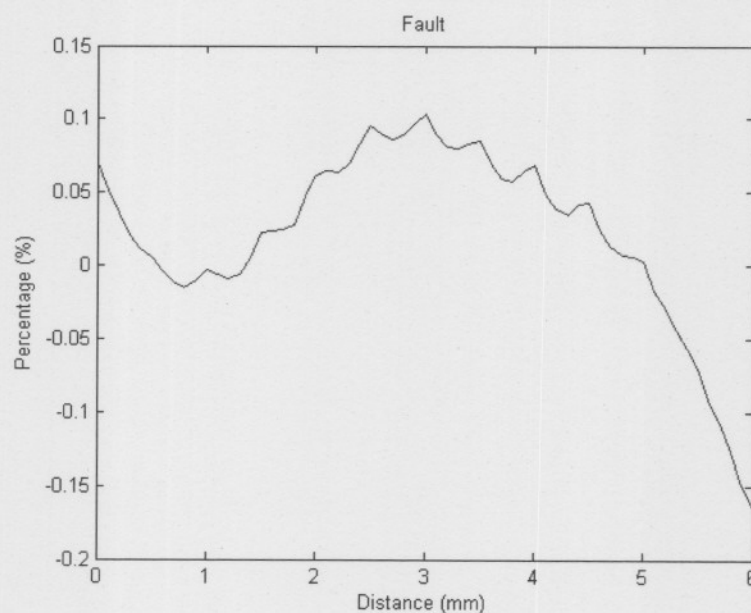


Figure 3.17: Ball setup fault in percentage

3.7.3 Two 5 mm LEDs and two phototransistors

The following results are found from an optical sensor that uses two infrared LEDs and two phototransistors in parallel.

The dimensions of the LED and the phototransistor are 5 mm diameter. The practical layout is as shown in Figure 3.18. The LEDs and phototransistors are placed 4 cm apart.

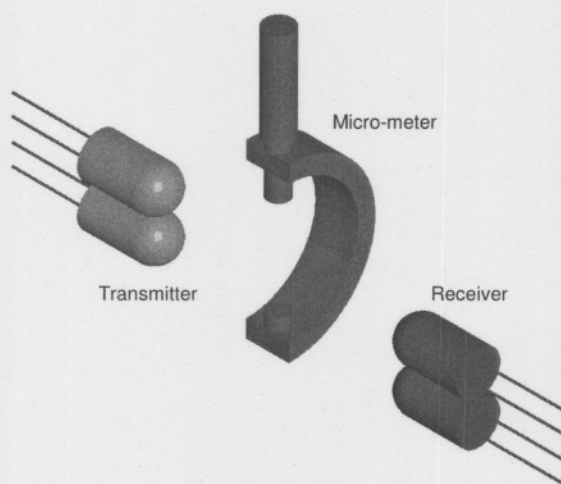


Figure 3.18: Double LED and phototransistor layout

Figure 3.19 shows the best response at all the combinations of components and setup. The sensing range is large and the linearity is relatively constant over the wide range.

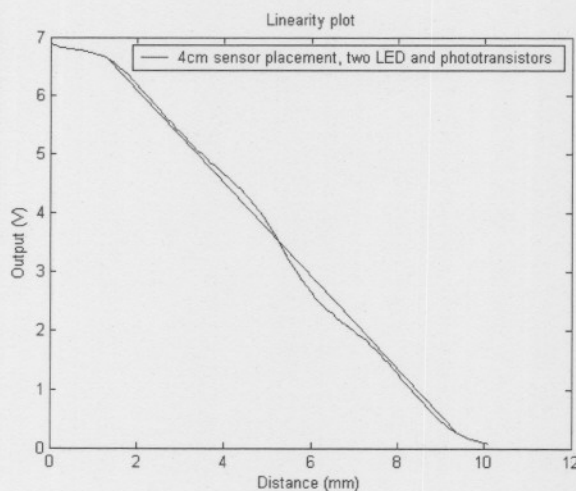


Figure 3.19: Parallel phototransistor sensor response

The fault is relatively small over the wide sensing range. The results were good enough for the sensor to be used in the magnetic bearing demonstrator project.

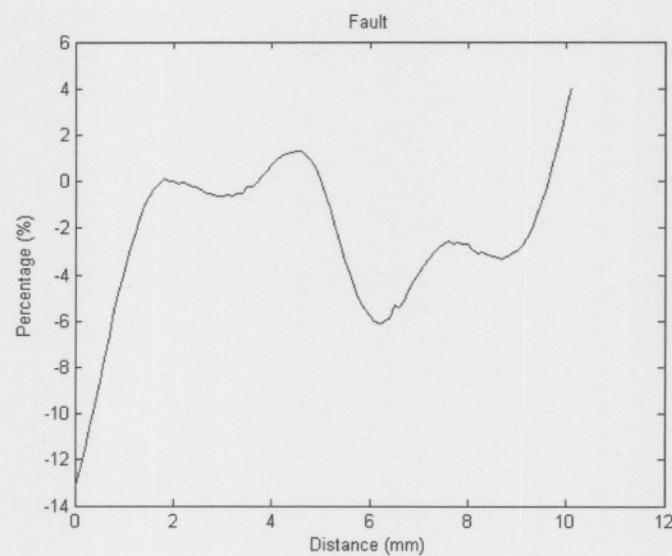


Figure 3.20: Parallel phototransistor fault in percentage

3.8 Optical sensor implementation

The double layer layout was implemented on the AMB demonstrator as shown in Figure 3.21 and Figure 3.22.

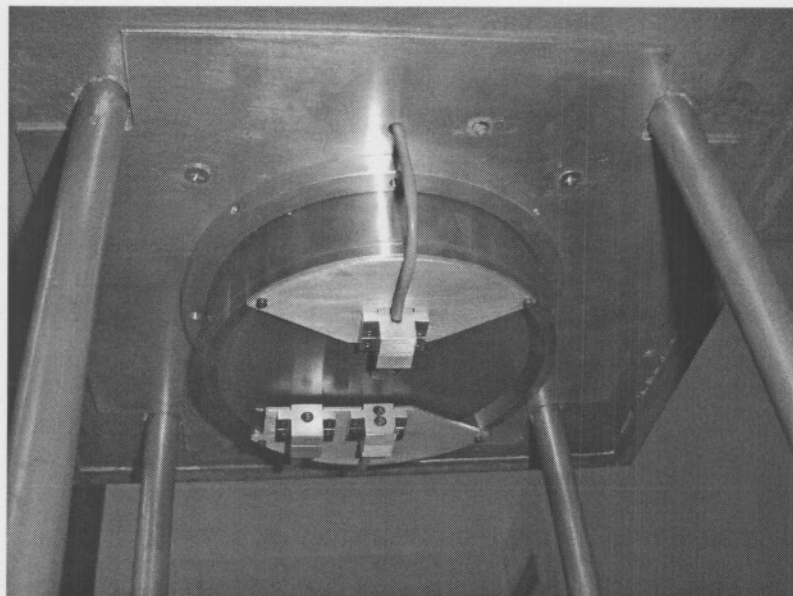


Figure 3.21: Double layer optical sensor

The two phototransistors, which are directed at the two infrared LEDs, is the position sensor component. The single phototransistor on the left in Figure 3.21 measures the external infrared noise levels.



Figure 3.22: Double layer optical sensor

The LEDs and phototransistors are connected to the analog circuitry through shielded cables. These shielded cables reduce the noise susceptibility.

3.9 Conclusion

The infrared sensor circuit is sensitive to external infrared noise. The reference infrared detector decreases the effect of external noise. It is assumed that the infrared detectors are in close proximity and the sensor detects the same amount of noise.

The distance between the LEDs and phototransistors plays an important role in the linearity. An increase in distance decreases the linearity.

The linear range of the sensor is also dependent on the characteristics of the infrared phototransistor and the dimensions of the phototransistor and LED. The larger the diameter of the phototransistor and LED, the more linear the sensor becomes.

The sensitivity ($V/\mu m$) of the detector can be adjusted by the amplification factor for the specific application.

The sensor range can be increased by placing two phototransistors in parallel. The parallel combination has a small effect on the sensor response, but the sensor range is dramatically increased.

Inductive Sensor

The inductance sensor is a relatively high precision sensor, thus the accuracy of the sensor is ideal for AMB applications. The inductance sensors are frequently used in smaller scale AMBs. When an inductive sensor is developed for AMBs the following design factors must be addressed: 1) the bandwidth of the sensor 2) minimizing of magnetic noise and 3) the influence of different ferromagnetic materials. This chapter is based on these design factors.

4.1 Inductance

The inductive sensor is based on the proportionality of the air gap to the inductance. If a coil is placed as shown in Figure 4.2, the inductance of the coil will change according to the size of the air gap [2], [6]. The coils are subjected to a high frequency voltage. This ac voltage with a constant frequency f will result in a constant reactance if the air gap is kept in a static position. But the distance of the air gap changes continuously, thus the core's properties change over time. The reactance is proportional to the distance as described in (2.1).

The distance could be determined by using the voltage difference between the two series coils. The changing reactance of the two series coils is shown in Figure 2.5.

These two series coils make use of ferrite cores. These cores are constructed of the halves of small ferrite transformers. The core is shown in Figure 4.1.

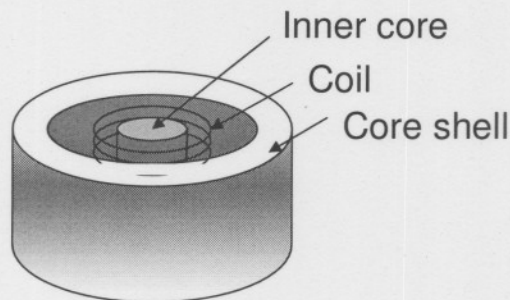


Figure 4.1: Sensor core

Only one half of the transformer is used for each coil. These cores must have a high permeability. This will increase the accuracy of the reading because the reactance is mainly dependent on the air gap.

An advantage of the ferrite core is that it reduces the susceptibility to other magnetic fields, since the core is nearly a closed shell except at the air gap.

4.2 Inductive sensor layout

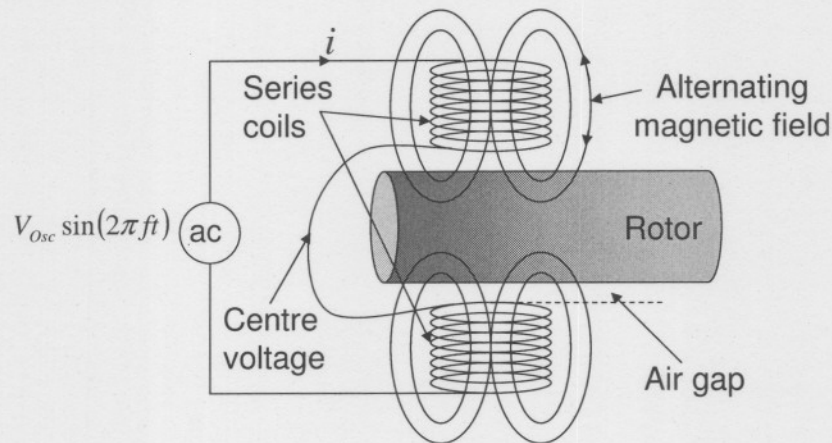


Figure 4.2: Inductive sensor

If the rotor's position is shifted upwards the inductance of the top coil will increase and the bottom coil will decrease, thus the sensor output will decrease. This continuous change in air gap has some effects that must be taken in consideration. The measurement distance between the sensor's core and the rotor must be carefully considered, because the linearity of the sensor is a function of the distance. If the distance is too large or too small the linearity of the sensor will be poor. Thus the measurement range must be calibrated for the specific core and rotor material. If the frequency f of the injected voltage $V_{osc} \sin(2\pi ft)$ is too high the eddy current effect will increase and the linearity of the sensor will be decreased. If the air gap of one of the sensor's coils is much smaller than the other one the eddy current effect will be predominantly in the coil with the small air gap. Again it is important to calibrate the measurement range of each coil and ensure it is the same for both the coils or the sensor will become unsymmetrical and it will decrease the linearity of the sensor. Another point is to ensure the two coils are identical. This will also increase the linearity of the sensor.

If the rotor is in the centre position, the centre voltage of the two series coils will be $\frac{1}{2}V_{Osc}$. This is due to the equal impedance of the series coils. The impedances of the two coils act like a voltage divider.

4.3 Advantages and disadvantages

The sensors linearity and the measurement range are relatively good.

The main disadvantage of the sensor is that it is sensitive for any magnetic noise. The noise could be classified into two categories, internal and external noise.

The internal noise is generated by the sensor components itself which is already a main concern. The other concern is that a magnetic bearing is one large nonlinear magnetic field.

The external noise is generated by any system in the vicinity of the sensor, like motors and generators which relies on magnetism. Great care must thus be taken to minimize the effect of exposure to magnetic fields and to situate coils at shielded spaces.

4.4 Design phase

Figure 4.3 shows the block diagram of the inductive sensor. In this section each of the blocks will be discussed and developed with analog electronics.

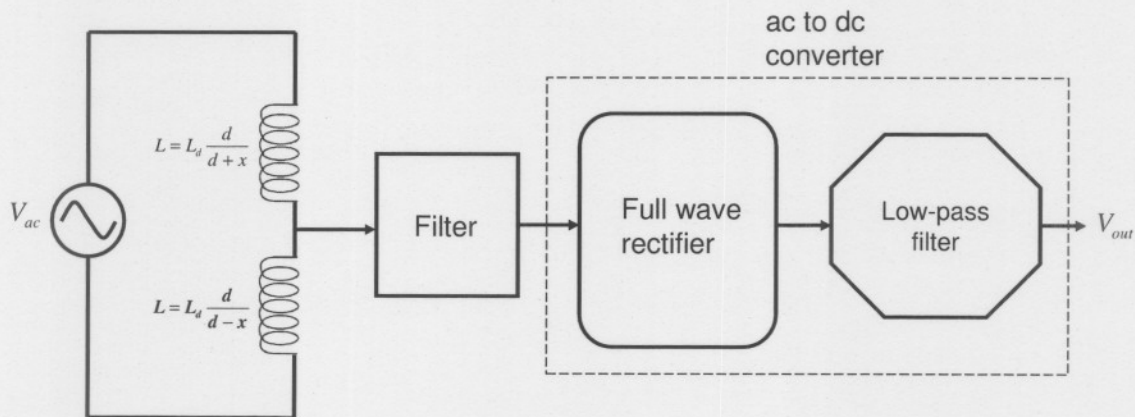


Figure 4.3: Inductive sensor block diagram

The first block after the two series coils shows a filter. As mentioned before an ac supply ($V_{Osc} \sin(2\pi ft)$) with a specific frequency f is applied to the sensor coils. This ac voltage is generated by a Wien-bridge oscillator.

All the other frequencies are defined as noise and could be eliminated by using a band-pass filter. The order of the filter is determined by the amount of noise which are near the injected frequency f .

The second block is a full wave rectifier. The purpose of the rectifier is to convert the negative voltages to positive, since only the modulated amplitude is of interest. A full wave rectifier is used, since it will increase the accuracy of the sensor. When a half wave rectifier is used some of the information is lost.

The third block shows an ac to dc converter. The purpose is to convert the rectified ac voltage to a dc voltage. The result is a changing dc voltage proportional to the outer band of the rectified ac voltage. Figure 4.4 illustrates the demodulation process. There is always a ripple component present. This ripple must be as small as possible or it will interfere with the control of the AMB.

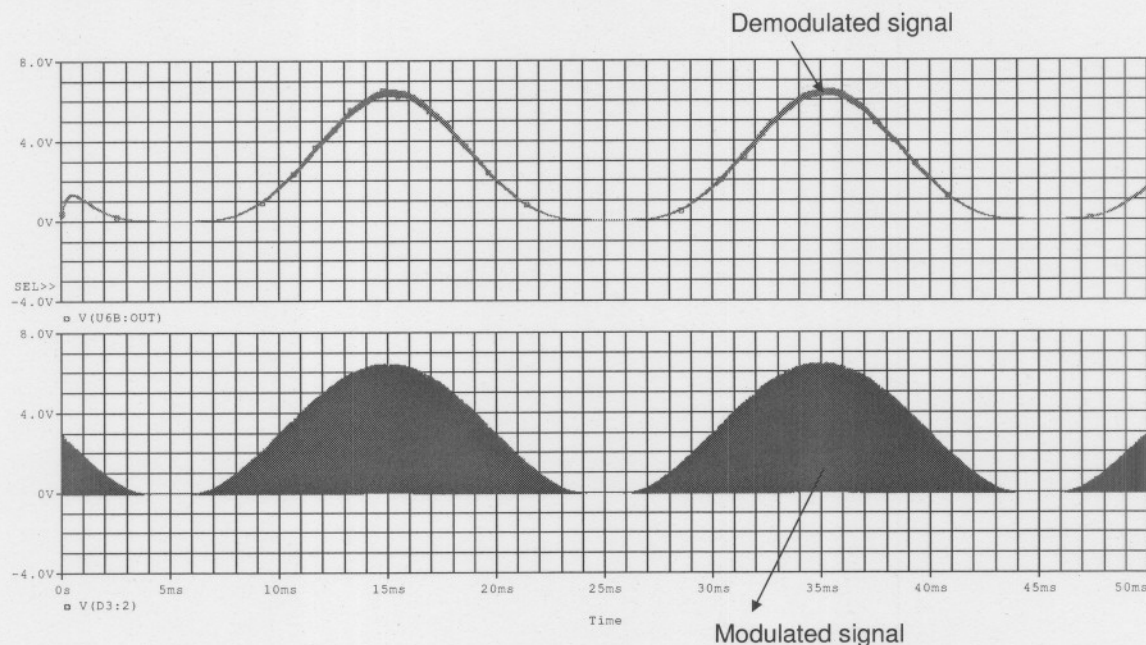


Figure 4.4: Rectified ac to dc output

4.5 Inductive sensor circuit design

The circuit design is based on Figure 4.3. Each component will be developed individually and then integrated to form the sensor.

4.5.1 Wien-bridge oscillator

The function of the Wien-bridge oscillator is to generate a pure sine wave. This sine wave is injected into the sensor coils. The frequency f is chosen relatively high because the inductance of the sensor coils are relatively low. Equation (4.1) shows how the reactance changes as the frequency and inductance changes.

$$Z_L = j2\pi fL \quad (4.1)$$

In this design the air gap is relatively large so the eddy current effect is reduced. The injection frequency f is chosen at 36 kHz. The Wien-bridge oscillator is shown in Figure 4.5. The signal generating IC 8038 may be used instead of the Wien-bridge oscillator, but it also makes use of external capacitors [3]. These external capacitors are greatly affected by temperature changes, thus a change in temperature will increase or decrease the oscillation frequency.

The oscillation frequency is defined as:

$$f = \frac{1}{2\pi RC} \quad (4.2)$$

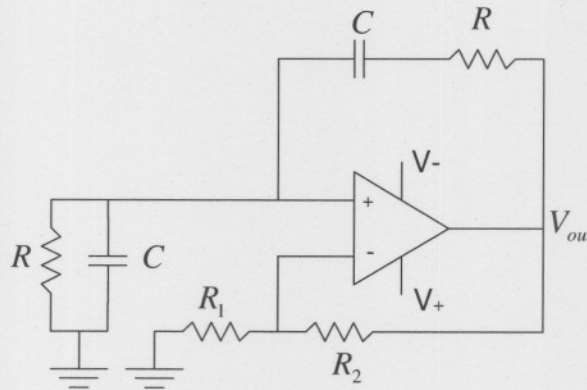


Figure 4.5: Wien-bridge oscillator

In this case f is known, thus C must be calculated [3]. The resistor R is arbitrarily chosen. The Wien-bridge oscillator requires that $R_2 / R_1 = 2$.

4.5.2 Band-pass filter

The band-pass filter as shown in Figure 4.6 was designed by a CAD program, FilterSolutions.

A third order filter was chosen because the second order cut off gradient is not steep enough and the third order is still implementable with general parts.

In this case a third order high-pass and low-pass filter were used in series to reduce the effect of tolerance of standard parts. All the operational amplifiers used in the circuit are TL084 ICs. The TL082 ICs are supplied with a ± 15 V supply.

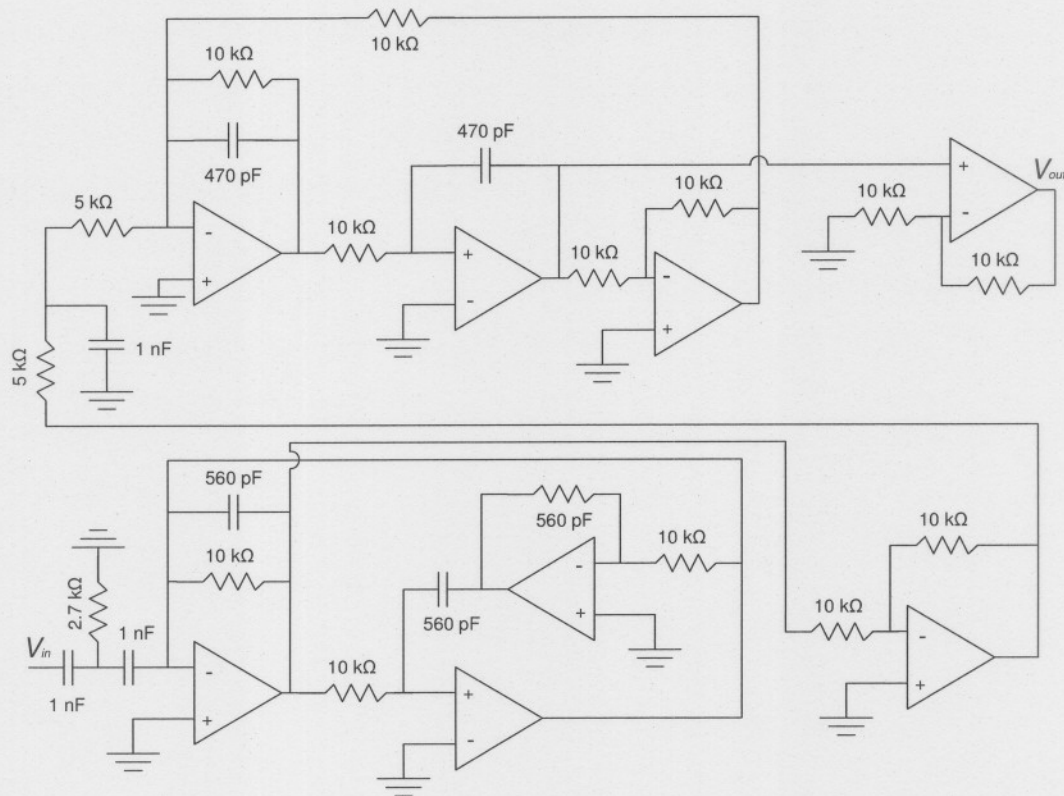


Figure 4.6: Band-pass filter circuit

The response of the filter is shown in Figure 4.7. The pass band of the filter is at the frequency f of the injected voltage which in this case is about 36 kHz.

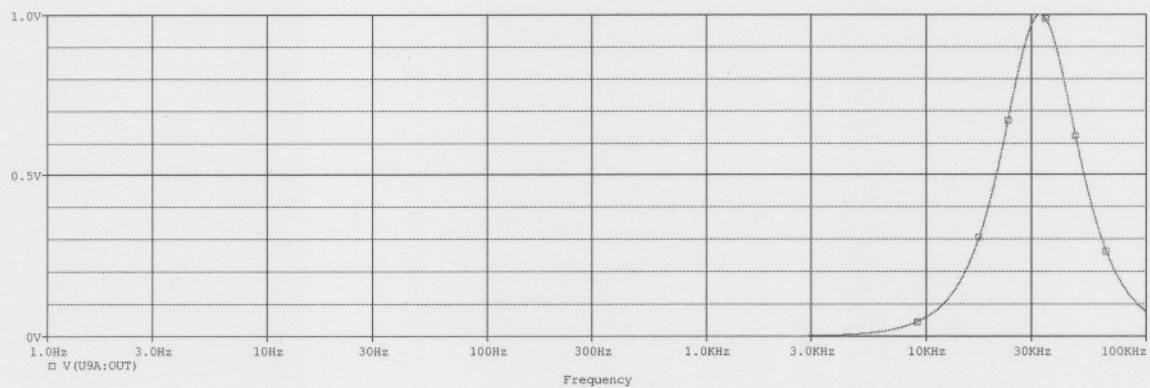


Figure 4.7: Band-pass filter response

4.5.3 Full wave rectifier

Different types of full wave rectifier (FWR) circuits are available. The four diode circuit as shown in Figure 4.8 gives an excellent output, but by using the ac to dc converter shown in Figure 4.10 the output is not as stable as desired. The output of the full wave rectifier combined with the ac to dc converter as shown in Figure 4.12 is more accurate over a wide range of frequencies. Both of these cases will be discussed.

The output of the four diode full wave rectifier is shown in Figure 4.9. If a pure sine wave is taken as input, the output of the rectifier at the zero crossing point is relatively near zero. In the case of the two diode configuration rectifier the output at the zero crossing point is not always near zero. The operational amplifier used in the full wave rectifier is the TL082 IC. 1N4148 signal diodes are used.

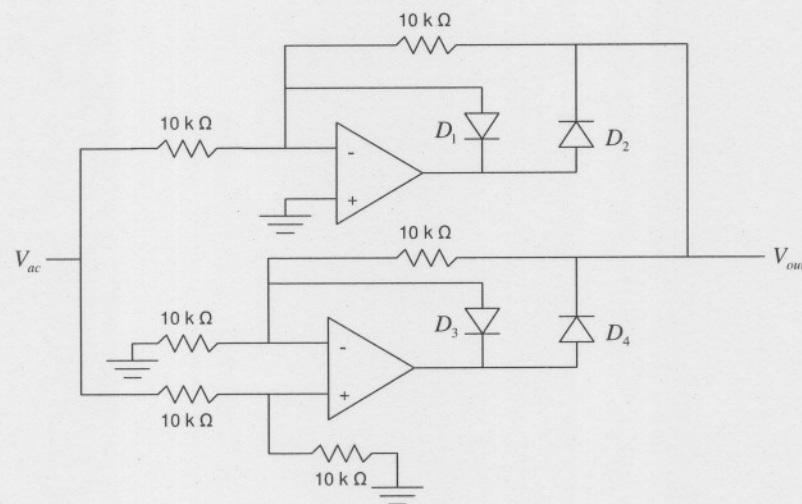


Figure 4.8: Full wave rectifier circuit

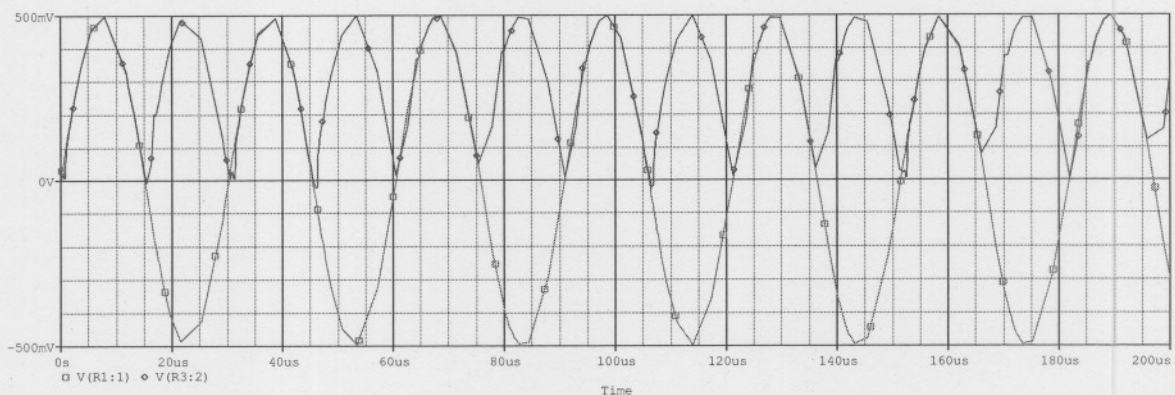


Figure 4.9: FWR input and output voltage

4.5.4 Ac to dc converter with four diode rectifier

The ac to dc converter in this section is designed to work with the four diode rectifier and low-pass filter. The low-pass filter consists of a first order RC network. The low-pass filter will always have a residue ripple component because of the high bandwidth requirement. The low-pass filter circuit is shown in Figure 4.10.

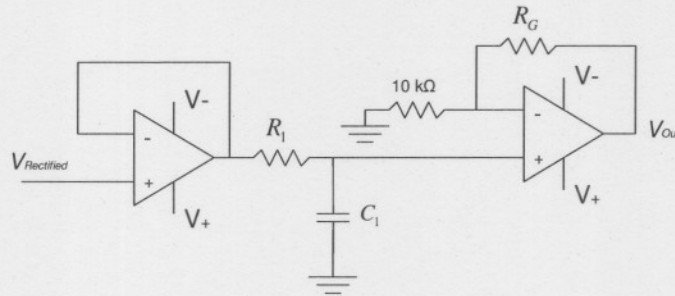


Figure 4.10: Ac to dc converter

This ripple component can be defined as noise. The maximum allowable ripple of the sensor is chosen to meet the specification of the controller and signal to noise ratio. The operating frequency f of the sensor coils is known (36 kHz). The resistor R_1 and capacitor C_1 determine the cut off frequency of the low-pass filter [19].

The cut off frequency of the low-pass filter determines the sensor's bandwidth in Hz.

$$BW_{sensor} = \frac{1}{2\pi R_1 C_1} \quad (4.3)$$

Figure 4.11 shows the relationship between the capacitance and the bandwidth.

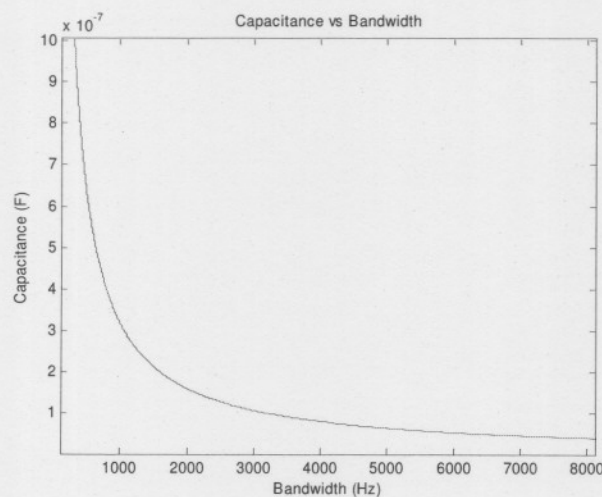


Figure 4.11: Capacitance versus bandwidth

It is evident that there is an inverse relationship between the bandwidth and the ripple, determined by the capacitor.

4.5.5 Ac to dc and full wave rectifier combination

The circuit in Figure 4.12 is a combination of a full wave rectifier and ac to dc converter [3]. The circuit is more stable than the previous design and is easier to implement. The main advantage of this ac to dc converter circuit is that it can operate across a wide range frequency.

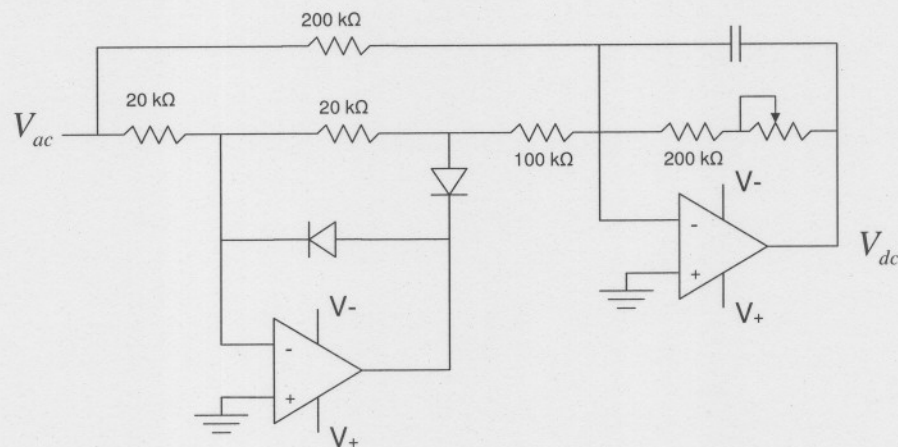


Figure 4.12: Ac to dc converter

The capacitor C_1 in Figure 4.12 determines the amplitude of the ripple component. The oscillation frequency f is constant, thus the capacitor C_1 is calculated as follows.

$$C_1 \gg \frac{1}{4\pi(R_1 + R_{POT})f} \quad (4.4)$$

The gain is adjusted by resistor R_{POT} .

4.5.6 Linearity evaluation on a homopolar AMB

The sensor was evaluated using a homopolar AMB. The maximum deviation of the rotor from the centre is 0.5 mm. The inductive sensors, used in conjunction with the homopolar AMB, are shown in Figure 4.13.

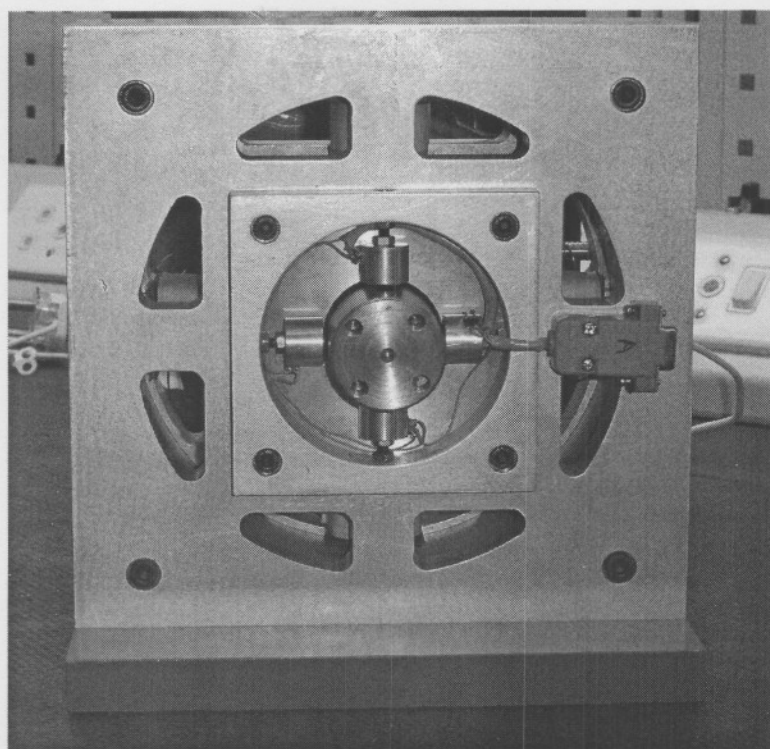


Figure 4.13: Inductive sensors for homopolar AMB

A low frequency sine wave is given as reference voltage to the horizontal and vertical control system. The position of the rotor changes according to the amplitude of the reference voltage. Figure 4.14 shows the linear relation between the two reference voltages for both the vertical and horizontal inductive sensors and their outputs [20].

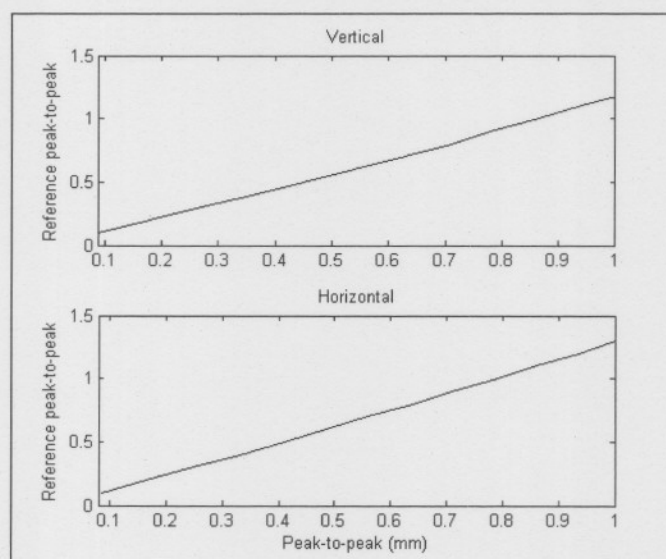


Figure 4.14: Linearity test

4.5.7 Sensor circuit

The inductive sensor circuit diagram is included on the CD in appendix C.4.

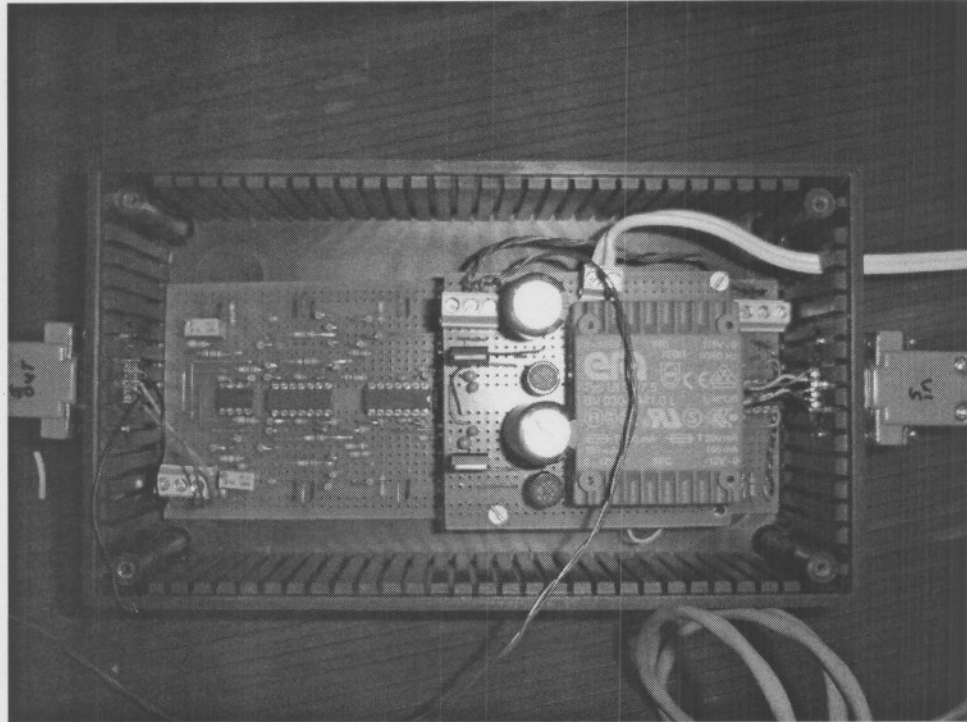


Figure 4.15: Inductive sensor analog circuit

Figure 4.15 shows the analog circuitry for horizontal and vertical sensor with the internal power supply.

4.6 Design problems

An important part of the design is to minimize the effect of noise. The sensor's coils must be shielded to minimize the effect of electric fields. This is done by covering the top and side of the coil with an aluminium casing.

It is usually not possible to place the coils and the electronic components of the sensor in close proximity, thus the cable which connects the coils and electronics must be shielded.

Another problem is to minimize the internal noise. This could be done by placing decoupling capacitors to ground at every supply point. Another precaution is to minimize the length of all connections and proper circuit layout.

4.7 Conclusion

The static output of the sensor is reliable, but as the rotation speed of the rotor increases the deformation of the signal increases. This is due to some frequency components near the bandwidth of the sensor.

The bandwidth of the sensor is above 500 Hz which was the requirement of the homopolar AMB.

The static linearity test shows that the sensor has a fine linear response. The sensor is not sensitive to lubricants and normal temperature changes at the sensor coils. This advantage makes the inductive sensor more likely to be used in dirty conditions. The change in temperature at the electronics may have an effect on the oscillation frequency of the Wien-bridge oscillator. The 8038 function generator IC may be used, but the IC also makes use of an external capacitor which will also change as the temperature changes.

Self-Sensing modelling

A more complicated but intriguing method for sensing the position of an active magnetic bearing (AMB) is the sensorless sensor or also known as the self-sensing method. In the case of a nuclear fusion reactor, some kind of sensor must be used to control the rotor. These sensors must be able to withstand the harsh conditions. Since the sensors are used under severe irradiation circumstances, the number of sensors is desired to be small or zero. Therefore, it is important to develop a sensorless or self-sensing control system.

The self-sensing method which makes use of the current response of the switching amplifier will be introduced. [15] The self-sensing method relies on the high frequency current slope and amplitude of the amplifier voltage. These current components change according to the air gap, thus the changing current could then be demodulated to determine an estimation of the position of the rotor. Figure 5.1 illustrates the basic operating principle of the self-sensing method.

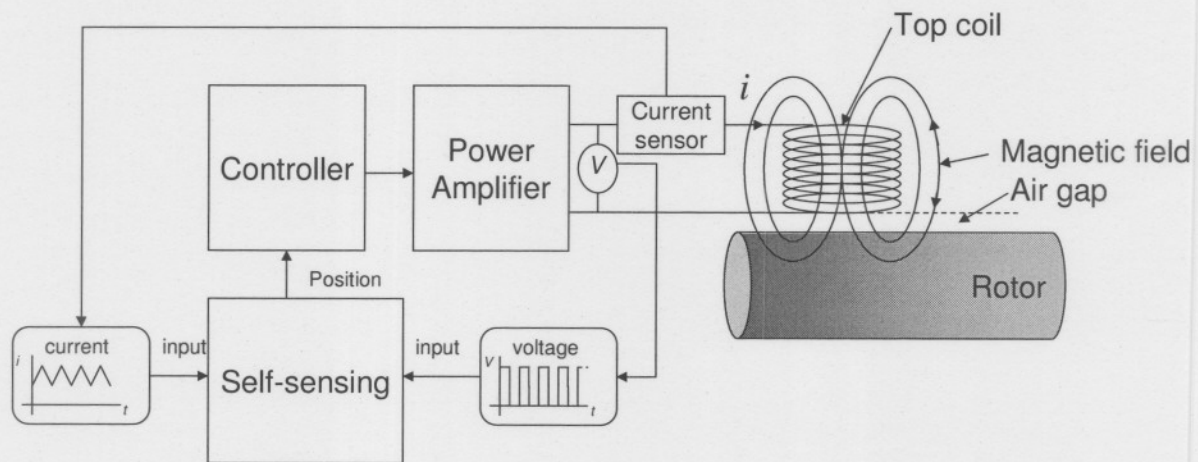


Figure 5.1: Self-sensing method

The self-sensing method has the advantage that the voltage and current measurements could be taken externally from the AMB. The sensor could be situated outside the AMB, making it ideal for high temperature applications.

This method has the added benefits of lowering the cost and reducing potential failure of the sensor. The wiring between the bearing and the controller is also minimized.

These economic benefits and improved reliability makes the sensor more appealing for future applications.

Before the position estimation could be derived or implemented on the active magnetic bearing, the AMB model and the amplifier used to drive the AMB must be discussed in detail.

5.1 Active magnetic bearing model

In this section the model of the AMB's coil will be derived from fundamental equations [15]. The primary point to start is to derive a mathematical model of the coil.

As mentioned before the current through the coil is the fundamental component of the self-sensing method. By using Faraday's law the voltage across the coil is [21]:

$$V = N \frac{d\Phi}{dt} + iR \quad (5.1)$$

By using Ampere's loop law, (5.2) is derived.

$$\oint \mathbf{H} \cdot d\mathbf{l} = Ni \quad (5.2)$$

An approximation of the Ampere loop in (5.2) could be derived, if it is assumed that the coil is long enough (length of the core is large) [21].

$$H_c l_c + 2H_g (g \pm x) = Ni \quad (5.3)$$

In this case it is stated that g is the nominal air gap and that x is the changing air gap. Equation (5.3) applies for the transverse pair of coils.

A linear relationship exists between the magnetic induction and the flux density. This relationship is dependent on the type of material. The magnetic field intensity through the gap is given by (5.4) and the intensity through the core is given by (5.5), [21].

$$H_g = \frac{B_g}{\mu_0} \quad (5.4)$$

$$H_c = \frac{B_c}{\mu_0 \mu_r} \quad (5.5)$$

where μ_0 is the permeability of air and μ_r is the relative permeability of the material. In (5.4) and (5.5) all the nonidealities are neglected.

Furthermore if it is assumed that the cross-sectional area of the air gap is constant and no leakage or fringing takes place. The flux density in the core and the air gap is the same.

$$B_c \approx B_g \quad (5.6)$$

Substituting (5.4), (5.5) and (5.6) into (5.3) the results in (5.7).

$$\Phi = \frac{Ni \cdot \mu_0 A_g}{2(g \pm x) + \frac{l_c}{\mu_r}} \quad (5.7)$$

The inductance of the coil could be determined by the following equation.

$$L = N \frac{\partial \Phi}{\partial i} \quad (5.8)$$

By substituting the flux through the core, (5.7) into (5.8) a mathematical model of the inductance of the AMB's coil is derived.

$$L = N \frac{\partial \Phi}{\partial i} = \frac{N^2 \mu_0 A_g}{2(g \pm x) + \frac{l_c}{\mu_r}} \quad (5.9)$$

If it is assumed that the change of the air gap is zero, the nominal inductance of the coil is defined as follows:

$$L_0 = \frac{N^2 \mu_0 A_g}{2g + \frac{l_c}{\mu_r}} \quad (5.10)$$

Substituting (5.9) into (5.1) the model of the coil current could be written as:

$$\frac{di}{dt} = \frac{2(g \pm x) + \frac{l_c}{\mu_r}}{N^2 \mu_0 A_g} (V - Ri - i \frac{dL}{dt}) \quad (5.11)$$

In (5.11) the last term is known as the back electromotive force (EMF). In most cases the EMF is small, thus it could be neglected. Neglecting the EMF in (5.11), the coil model is written as follows [15].

$$\frac{di}{dt} = \frac{2(g \pm x) + \frac{l_c}{\mu_r}(V - Ri)}{N^2 \mu_0 A_g} \quad (5.12)$$

Now that the model of the coil is defined the amplifier must be understood. In the next section the types of amplifiers are discussed.

5.2 Amplifiers

The AMB makes use of some control technique to maintain the rotor in a stable position. This control output is a small voltage which increases or decreases the current through the AMB to maintain a stable system. This current is generated by an amplifier. In this section the linear amplifier and switching amplifier will be discussed.

5.2.1 Linear Amplifier

The linear amplifier makes use of some input voltage from the control system and the current is directly related to the voltage [15]. Equation (5.13) shows the relation.

$$I_{out} = \beta V_{in} \quad (5.13)$$

The linear amplifier is the easiest amplifier to implement, but is not effective to use in the case of the AMB. The efficiency of the linear amplifier is about 5 to 10 percent. The linear amplifier consists of MOSFETs or IGBTs which is operated in its linear range. An advantage of linear amplifiers is that they generate nearly no noise. Figure 5.2 shows a linear amplifier.

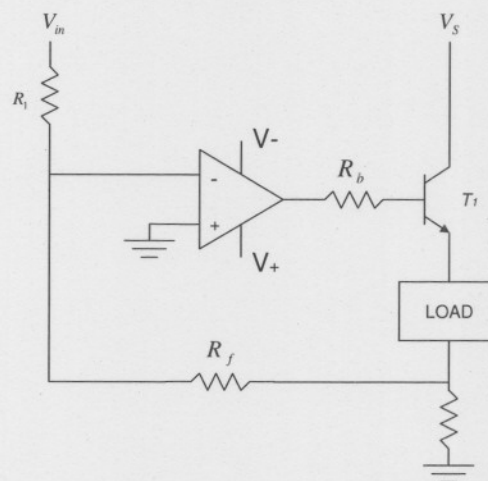


Figure 5.2: Linear amplifier

The linear amplifier will not be used thus no further detail will be discussed.

5.2.2 Switching Amplifier

Switching amplifiers are very efficient (80 to 90 percent) compared to linear amplifiers [15]. The switching amplifiers are commonly used to drive AMBs.

The most known switching amplifier is the H-bridge amplifier. The H-bridge is shown in Figure 5.3.

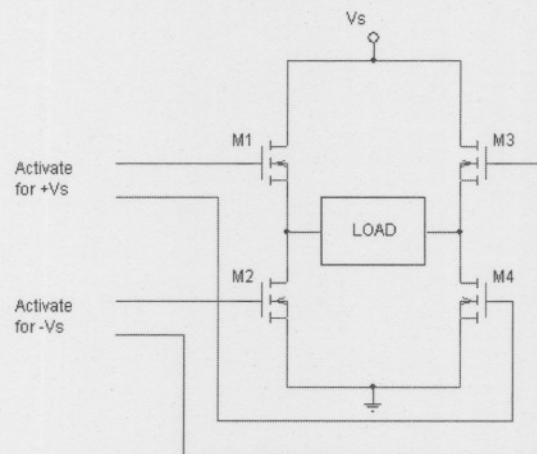


Figure 5.3: H-bridge

Switching amplifiers have constant voltage, with a varying duty cycle. A typical PWM wave is shown in Figure 5.4.

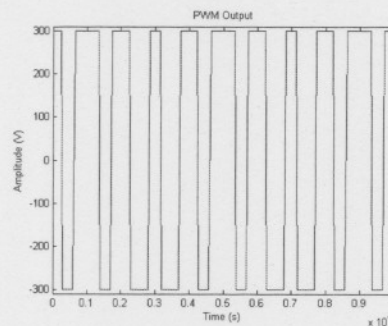


Figure 5.4: PWM output

The input voltage of the amplifier controls the duty cycle of the switched period. The switching amplifier has some disadvantages.

The output contains harmonic distortion and the surrounding electronic components are contaminated with switching noise. The high efficiency of the switching power amplifier makes it the obvious choice in AMB applications.

5.3 Simulated AMB model output

In this section the current output of the active magnetic bearing model will be discussed. These current outputs can be divided in three case studies. Case 1: constant air gap and duty cycle. Case 2: 50% duty cycle and varying air gap. Case 3: varying duty cycle and air gap.

5.3.1 Case 1

In the first case the current through the AMB model will be modelled at a constant PWM duty cycle and constant air gap. This AMB model is easily modelled in Simulink®, because it is a linear time-invariant system [22].

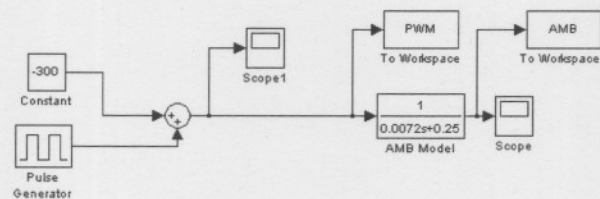


Figure 5.5: Simulink® model

A typical output of the Simulink® AMB model is shown in Figure 5.6.

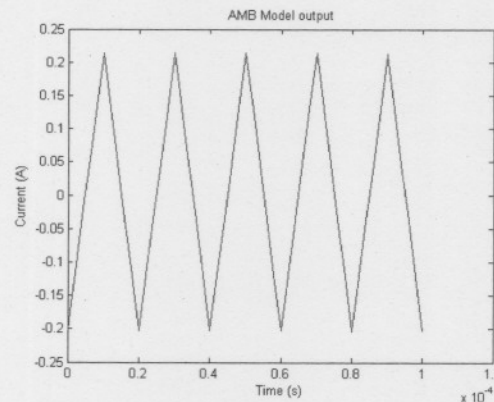


Figure 5.6: Simulink® model output

The output is a triangular wave with a constant frequency. This frequency is the same as the switching frequency of the switching amplifier.

5.3.2 Case 2

Case 2 is more complex because the AMB model is now time-varying. The equations in section 5.1 become important. By defining a varying air gap as: $x = 0.01e - 004\sin(2\pi f_{gx} t)$. The varying air gap is shown in Figure 5.7.

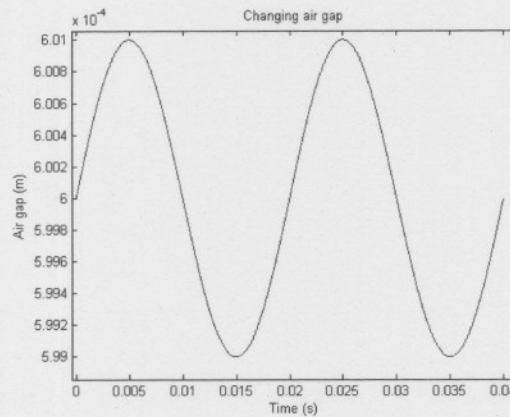


Figure 5.7: Varying air gap

By using (5.9) the corresponding inductance is calculated and is shown in Figure 5.8.

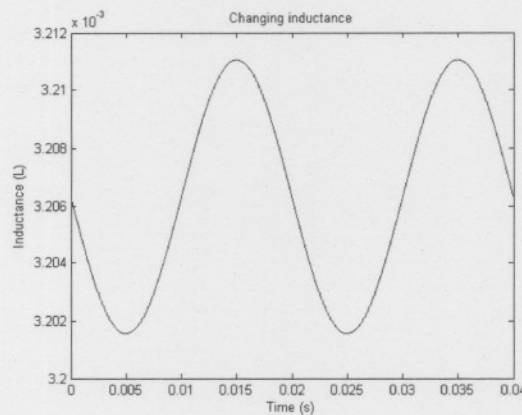


Figure 5.8: Varying inductance

The changing air gap and inductance have a large effect on the AMB's current output. The output of the current is shown in Figure 5.9. The time axis is plotted from 0.05 s to 0.08 s to show that the current is still in triangular form. The modulation effect on the current, due to the changing inductance is also clear.

The modulation of the current is directly related to the air gap. The output current of the AMB model resembles an AM signal. The Matlab® program used is on the CD in appendix B.1.7.

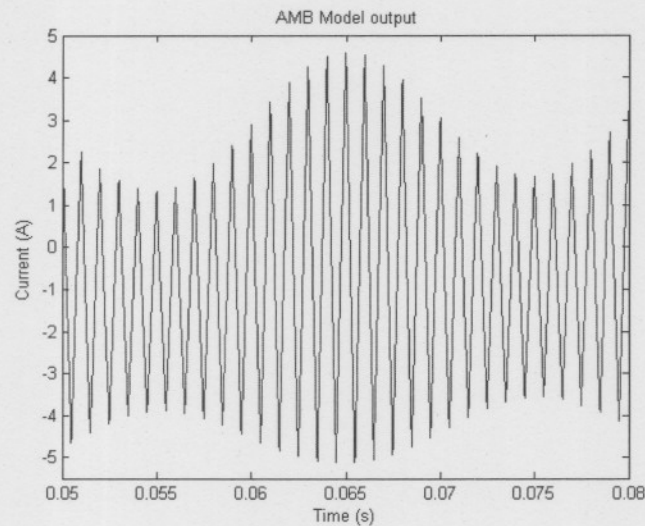


Figure 5.9: Varying current

5.3.3 Case 3

In this case a PID controller is used to stabilize the rotor at a specific position. The contribution of the controller to the PWM is a varying duty cycle. A disturbance could be inserted into the system. This contributes to a varying air gap. The output of the AMB model is shown in Figure 5.10.

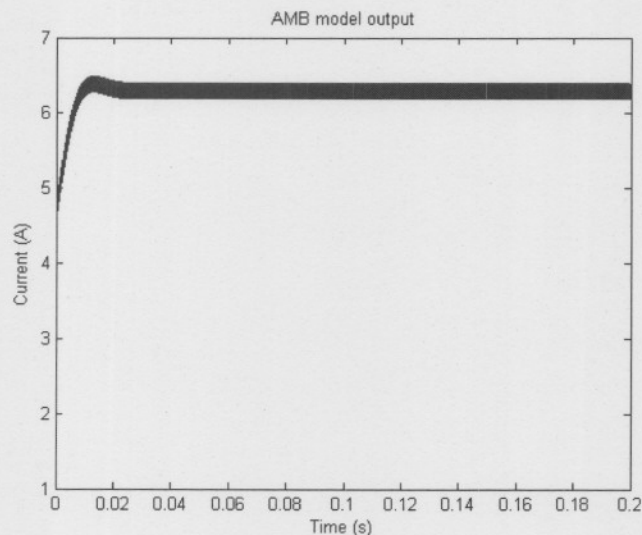


Figure 5.10: AMB model current output

Figure 5.11 shows that the current through the system is still of triangular form.

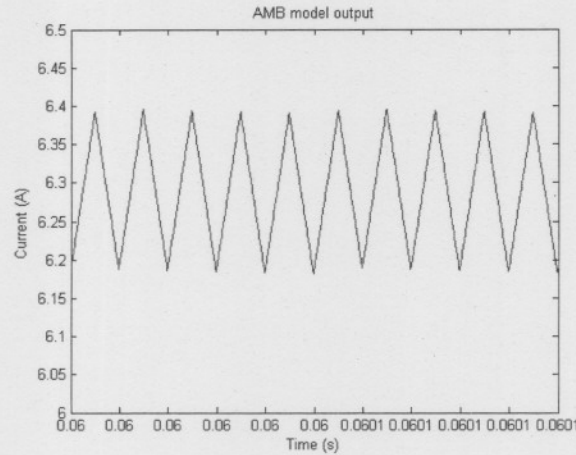


Figure 5.11: AMB model current output

In the next section the sensor will be mathematically analysed to be able to measure changing duty cycle and air gap.

5.4 Mathematical sensor development

In this section the self-sensing method will be derived mathematically. The continuous sensor will be converted to digital form to implement it in Matlab®. This section will make use of the results in section 5.3. The first obvious starting point is to develop a model for the demodulator.

5.4.1 AMB current demodulator

As mentioned in section 5.3 the output of the AMB model is directly related to the changing air gap, thus the inverse is also true. Remember that the duty cycle of the PWM was constant at 50%. The obvious choice is to build an AM demodulator. The demodulator makes use of a high-pass filter, full wave rectifier and a low pass filter. The demodulator will be proved in the continuous time domain by using harmonic analysis [20].

Harmonic analysis

The demodulator will be derived from the result of (5.12), [15]:

$$\frac{di}{dt} = \frac{2(g \pm x) + \frac{l_c}{\mu_r}(V - Ri)}{N^2 \mu_0 A_g} \quad (5.14)$$

If it is assumed that the reluctance of the core is neglect able and only one coil is considered to simplify the mathematics then (5.14) becomes:

$$\frac{di}{dt} = \frac{(g-x)}{gL_0}(V-Ri) \quad (5.15)$$

where L_0 is defined by (5.16).

$$L_0 = \frac{N^2 \mu_0 A_g}{2g} \quad (5.16)$$

If the resistance of the coil is neglected, the average current over a time t is given by (5.17) where ξ is determined by the switching logic.

$$i(t) = \int_0^t \frac{V(\xi)}{L_0 g} (g-x(\xi)) d\xi \quad (5.17)$$

Switching amplifiers make use of PWM voltages, thus the voltage $v(t)$ is constant between $+V_s$ and $-V_s$. The Fourier series of a rectangular wave is given by (5.18).

$$v(t) = \sum_{n=1}^{\infty} (a_n \cos n\omega_s t + b_n \sin n\omega_s t) \quad (5.18)$$

where ω_s is the switching frequency $\frac{2\pi}{\tau}$

The voltage output of the power amplifier for one switching period is shown in Figure 5.12.

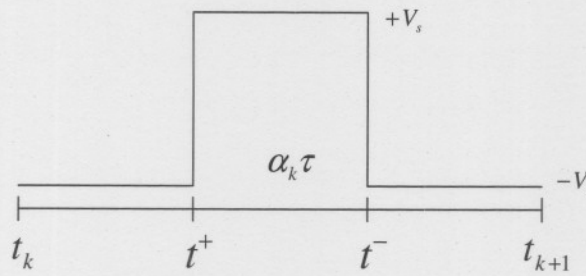


Figure 5.12: Voltage across coil

Equation (5.18) is an even function, is approximated as follows:

$$v(t) \approx -\frac{4V_s}{\pi L_0 g} \cos(\omega_s t) \quad (5.19)$$

The current is approximated by substituting (5.19) into (5.17):

$$i(t) = -\frac{4V_s}{\pi L_0 g} \int_0^t (g - x_0 \sin(\omega \xi)) \cos(\omega_s \xi) d\xi \quad (5.20)$$

Again it is assumed that the air gap changes sinusoidally with a frequency of ω and a position amplitude of x_0 . By solving (5.20) the current is as follows:

$$i(t) = -\frac{4V_s}{\pi L_0 g \omega_s} \left[g \sin(\omega_s t) - \frac{x_0}{1 - (\omega/\omega_s)^2} \left\{ -\frac{\omega}{\omega_s} + \frac{\omega}{\omega_s} \cos(\omega t) \cos(\omega_s t) + \frac{\omega}{\omega_s} \sin(\omega t) \sin(\omega_s t) \right\} \right] \quad (5.21)$$

High-pass filter

The purpose of the high pass filter is to remove all the low frequency components from the current output. If it is assumed that $\frac{\omega}{\omega_s} < 0.1$ and the pass band of the high-pass filter is chosen at $f_{pass} = 0.1\omega_s$ (5.21) becomes:

$$i(t) \approx -\frac{4V_s}{\pi L_0 g \omega_s} (g - x_0 \sin \omega t) \sin \omega_s t \quad (5.22)$$

From (5.22) it is clear that the current is related to the air gap and the current is amplitude modulated. Now the current could be demodulated by a full wave rectifier and a low-pass filter. The low-pass filter follows the envelope of the rectified current.

Amplitude demodulator

The conventional method to demodulate the current is by rectifying the output and then feeding the rectified output through a low pass filter. The output of the rectifier is shown in (5.23).

$$|i(t)| \approx \frac{4V_s}{\pi L_0 g \omega_s} (g - x_0 \sin \omega t) |\sin \omega_s t| \quad (5.23)$$

By evaluating the Fourier transform on the high frequency components of the rectified output, (5.24) is found as [23]:

$$|\sin(\omega_s t)| = \frac{1}{2} a_0 + \sum_{n=1}^{\infty} (a_n \cos(n\omega_s t) + b_n \sin(n\omega_s t)) \quad (5.24)$$

If (5.24) is convoluted with a low-pass filter time function with cut-off frequency just higher than the AMB bandwidth, the only component left is the dc component. Where a_0 is calculating as follows:

$$a_0 = \frac{1}{\pi} \int_0^\pi |\sin \theta| d\theta = \frac{4}{\pi} \quad (5.25)$$

The output of the filter is:

$$\begin{aligned} \text{Envelope} &= \frac{8V_s}{\pi^2 L_0 g w_s} (g - x) \\ \text{Envelope}_{\text{switching_period}} &= \frac{4V_s \tau}{\pi^3 L_0 g} (g - x) \end{aligned} \quad (5.26)$$

Forward path filter

Now that it is proved that the output of the AMB model is a function of the air gap, the reluctance of the coil is investigated in detail in terms of switching cycles. The current through the AMB model is defined as [15]:

$$i(t) = \begin{cases} I_0 - \frac{V_s}{L(x)} t, & @ 0 < t < t^+ \\ I_0 - \frac{V_s}{L(x)} (1 - \alpha) \tau + \frac{V_s}{L(x)} t, & @ t^+ < t < t^- \\ I_0 + \frac{2V_s}{L(x)} \alpha \tau - \frac{V_s}{L(x)} t, & @ t^- < t < \tau \end{cases} \quad (5.27)$$

where k starts at 0 (Figure 5.12) $\alpha_k = \alpha$, $\alpha = \frac{t^- - t^+}{\tau}$, $t^+ = \frac{\tau}{2}(1 - \alpha)$, $I_k = I_0$,

$$t^- = \frac{\tau}{2}(1 + \alpha), \quad t = t - t_k, \quad t_0 = 0$$

The first component of the forward path filter is the high-pass filter. If it is assumed that the high frequency components are much higher than the low frequency components the output of the filter is as follows:

$$i_{HP}(t) = i(t) - \left[I_0 + \frac{V_s}{L(x)} (2\alpha - 1)t \right] \quad (5.28)$$

The second component of the forward path filter is a rectifier. If the output of the high-pass filter is feed through a full wave rectifier the result is as follows:

$$|i_{HP}(t)| = \begin{cases} \frac{V_s}{L(x)} \cdot 2\alpha t, & @ 0 < t < t^+ \\ \frac{V_s}{L(x)} (1 - \alpha)(\tau - 2t), & @ t^+ < t < \tau/2 \\ \frac{V_s}{L(x)} (1 - \alpha)(2t - \tau), & @ \tau/2 < t < t^- \\ \frac{V_s}{L(x)} \cdot 2\alpha(\tau - t), & @ t^- < t < \tau \end{cases} \quad (5.29)$$

The third component of the forward path filter is a low-pass filter. When the rectified output is passed through the low-pass filter, then the output of the forward path filter is a function of the air gap.

$$u = \frac{1}{\tau} \int_0^\tau |i_{HP}(t)| dt = \alpha(1 - \alpha) \left(\frac{V_s \tau}{\mu_0 N^2 A_g} \right) \left(g \pm x + \frac{l_c}{2\mu_r} \right) \quad (5.30)$$

The output of the forward path filter (FPF) is shown in Figure 5.13. From simulation it is determined that the FPF output can not be used as feedback variable due to the delay. The AMB simulation is only stable over a very small range of cut off frequencies of the FPF. The Matlab[®] program is given in appendix B.1.2 and B.1.9.

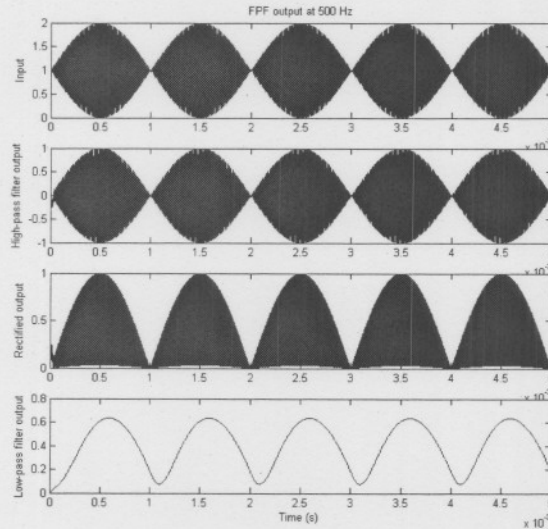


Figure 5.13: Forward path filter output

5.4.2 Parameter estimator and controller

The forward path filter (FPF) is not sufficient to use as a position predictor. From the AMB and self-sensing sensor simulation it was determined that AMB is stable over a very small region. This region is determined by the cut off frequencies of the high-pass and low-pass filter. In practice it is imposible to implement such filters. And the PID constant of the AMB model is largely affected. In this section the parameter estimator and controller will be developed [15]. The block diagram in Figure 5.14 shows the parameter estimator.

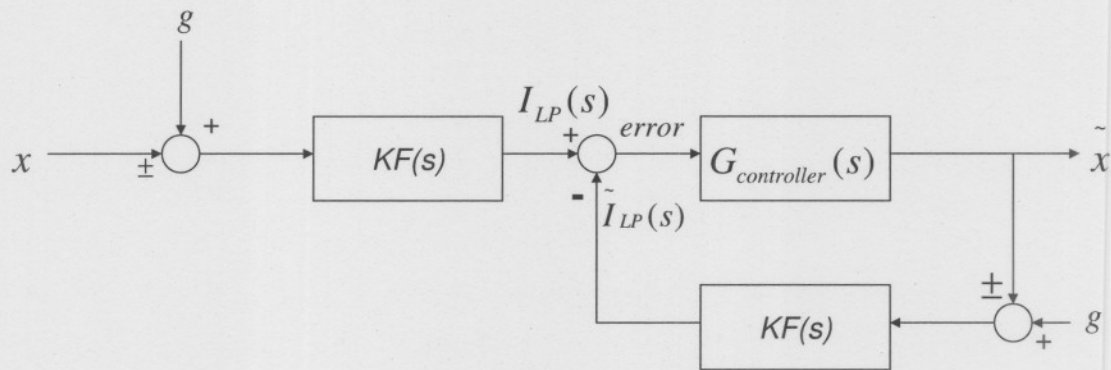


Figure 5.14: Parameter estimator block diagram

Equation 5.12 will be used again, but in this case the estimated current will be calculated as follows. The output of the controller is placed through a model of the bearing coils that give an estimated current of the bearing model. This AMB model is implemented by using an integrator to derive the current from the power amplifier voltage.

$$\frac{d\tilde{i}}{dt} = \frac{2(g \pm \tilde{x}) + \frac{l_c}{\mu_r}(V - Ri)}{N^2 \mu_0 A_g} \quad (5.31)$$

The output of the simulated current is placed through the same forward path filter as the real current. This simulated current is then subtracted from the real value to calculate the error. This error is the input to the PI controller. The error is defined as:

$$x^* = \arg \min_{\tilde{x}} \left[\left| \tilde{i} - \tilde{i}(\tilde{x}) \right| \right] \quad (5.32)$$

By using (5.30) and neglecting the coils reluctance, the output of the forward path filter is defined as:

$$u = \frac{1}{2}\alpha(1-\alpha)\left(\frac{V_s\tau}{\mu_0 N^2 A_g}\right)(g \pm x) = K(g \pm x) \quad (5.33)$$

where $K = \frac{1}{2}\alpha(1-\alpha)\frac{V_s\tau}{gL_0}$ and L_0 defined in (5.16).

If the output of (5.33) is investigated in the frequency domain then the output is in terms of K and the transfer function of the low-pass and high-pass filter and the air gap.

$$u = KF(s)(g \pm x) \quad (5.34)$$

From (5.33) it is found that K is a gain which is a function of the switching interval. The transfer function $F(s)$ is the transfer function of the forward path filter (FPF).

Now that it is established that K is not constant over time, the boundaries of K must be defined. When the duty cycle is 50% then the maximum of K will be reached. This is derived from (5.33).

$$K_{\max} = \frac{1}{2} \cdot \frac{1}{2} \left(1 - \frac{1}{2}\right) \frac{V_s\tau}{gL_0} = \frac{V_s\tau}{8gL_0} \quad (5.35)$$

The minimum of K is defined by the minimum of the allowable duty cycle of the power amplifier. Let the minimum duty cycle of the power amplifier be $\alpha_{\min}$ then the minimum K is $K_{\min} = \frac{1}{2} \cdot \alpha_{\min} (1 - \alpha_{\min}) \frac{V_s\tau}{gL_0}$. Note that the K is symmetrical at the 50% duty cycle point, thus duty cycles larger than 50% will also give a smaller value for K .

$$K_{\min} < K(\alpha) < K_{\max} \quad (5.36)$$

If the simulated current is passed through the FPF (5.34) then the output in the frequency domain is as follows.

$$\tilde{u} = KF(s)(g \pm \tilde{x}) \quad (5.37)$$

The error is the simulated current passed through the FPF $\tilde{u}(s)$ minus the real current passed through the FPF.

The error is given in (5.38).

$$error = u - \tilde{u} = KF(s)(x - \tilde{x}) \quad (5.38)$$

5.5 Simulation results and stability evaluation of filters

Now that the controller and the estimator are included into the self-sensing technique, the results of the sensor can be evaluated. Figure 5.15 shows the sensor output and the real position versus time. The estimator and PI controller improves the stability of the sensor. This improved stability also enlarges the cut-off frequency range of the filters used in the self-sensing technique.

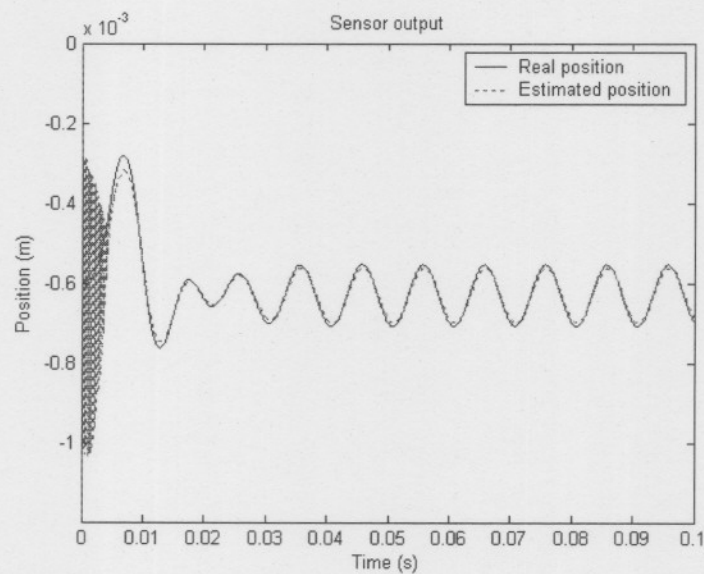


Figure 5.15: Self-sensing technique output

The stable region of the cut-off frequency of the low-pass filter is shown in Figure 5.16.

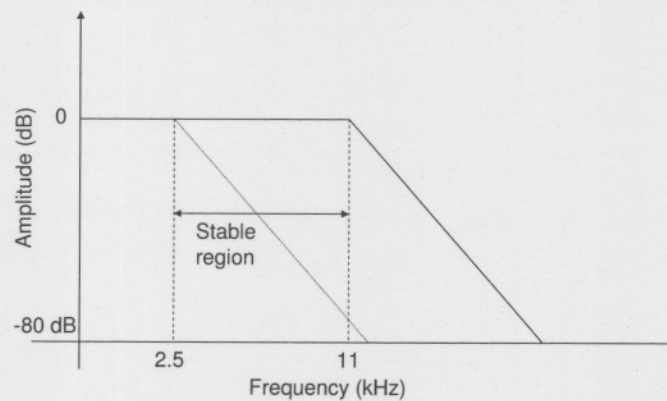


Figure 5.16: Stable cut-off frequency range for low-pass filter

The stable region of the cut-off frequency of the high-pass filter is shown in Figure 5.17.

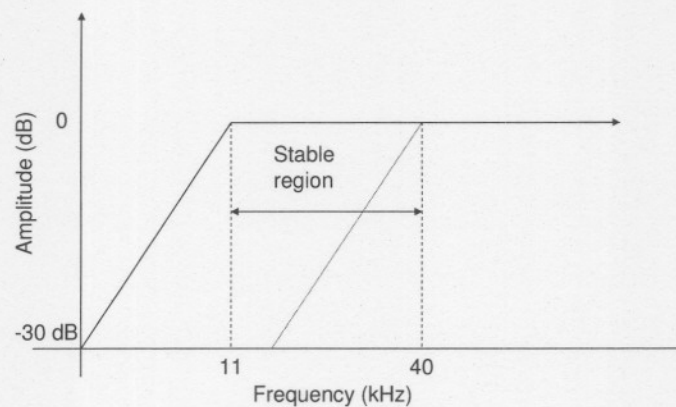


Figure 5.17: Stable cut-off frequencies range for high-pass filter

In the simulation it was found that the cut-off frequencies of the filters could be adjusted. When the low-pass filter cut-off frequency is 2.5 kHz and the high pass filter cut-off frequency is 40 kHz the sensor settles near the real position in a shorter time interval. The settling time is shown in Figure 5.18.

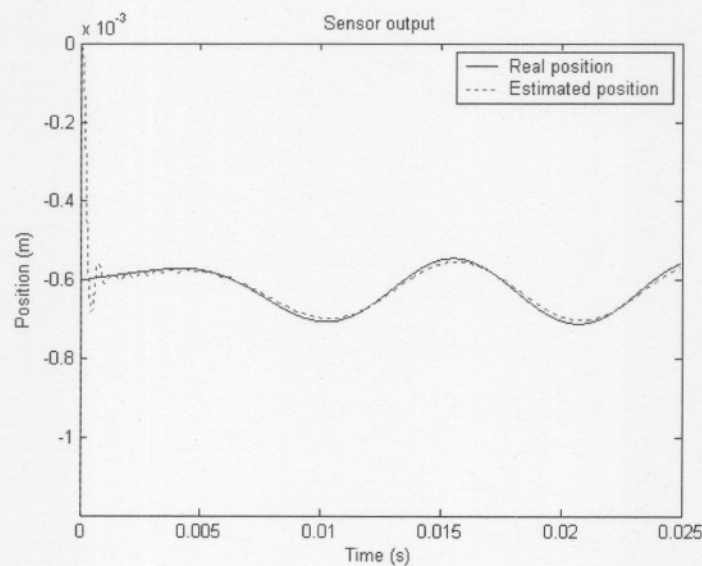


Figure 5.18: Faster settling time

In the case when the low-pass filter cut-off frequency is 10 kHz and the high pass filter cut-off frequency is 15 kHz the sensor takes a longer time to settle and the output overshoots the real position a number of times. The settling time is shown in Figure 5.19.

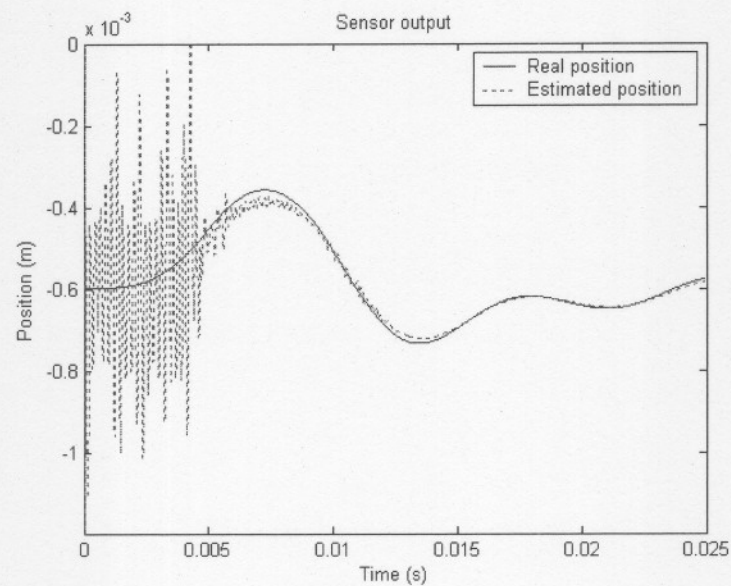


Figure 5.19: Slow settling time and overshooting

Figure 5.20 shows how the cut-off frequencies of the filters influence the settling time and the overshooting estimated position.

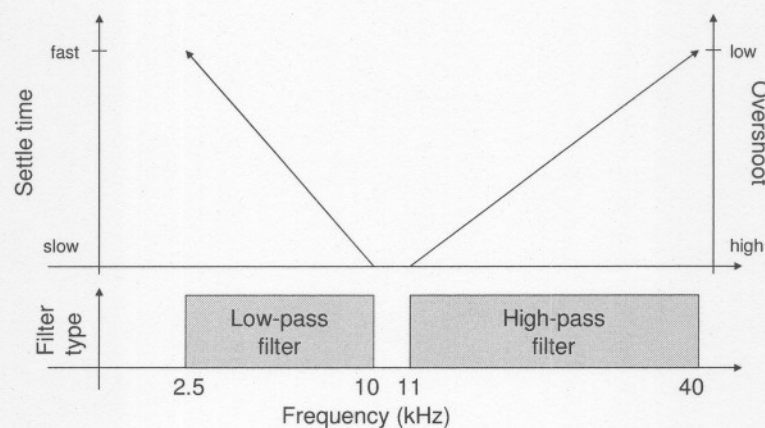


Figure 5.20: Settling time and overshoot versus cut-off frequency

The self-sensing technique with the parameter estimator was found to be more stable. The increased stability will make it easier to implement the sensor with analog electronics. The components used to implement the hardware are not all the same, thus the FPFs would not be exact duplicates. This increased stability will compensate for the unmatched components.

5.6 PI controller

From the simulation it was found that the PI controller variables are of utmost importance. The simulated model of the sensor is only stable over a specific region. This region is shown in Figure 5.21. These PI variables, shown in Figure 5.21, was determined from the simulation.

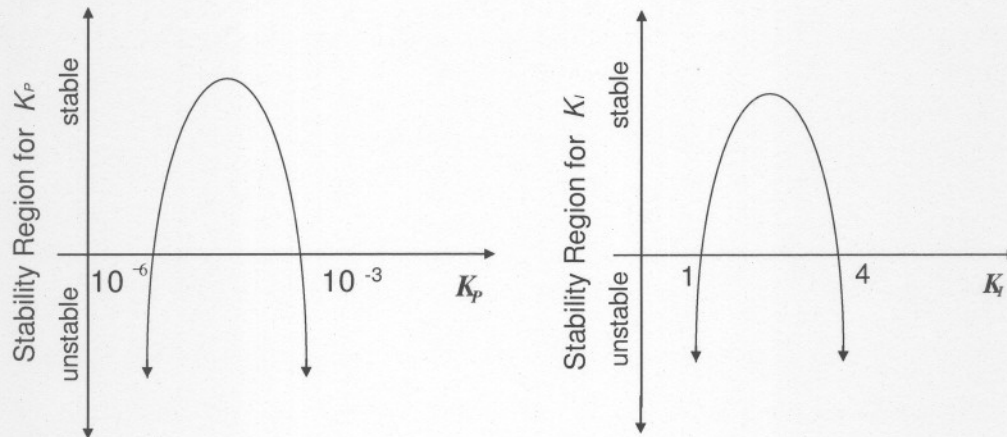


Figure 5.21: Stability region

5.7 Conclusion

The self-sensing method based on PWM current makes use of the high frequency triangular wave of the coil current and the switching voltage of the power amplifier. By filtering all the low current frequencies and demodulating the high frequencies the position is estimated. By adding a parameter estimator the stability of the sensor is increased and the cut-off frequency ranges of the filters are increased. The improved stability makes the sensor also more feasible to implement in hardware form.

AMBs used in extreme conditions will be more reliable and the sensors require no maintenance inside the AMB.

Self-sensing Sensor Implementation

In Chapter 5 the mathematical model of the self-sensing sensor was derived [15]:

This model was simulated and demonstrated to be stable. These mathematical blocks from Chapter 5 will be used as guidelines for the hardware development. All the components of the self-sensing method will be implemented in analog electronics. The reason for the analog implementation is that the self-sensing sensor makes use of high frequency components. In digital electronics the sampling rate of the digital signal processing must be very high. This chapter focuses on the design of the following main components 1) current sensor, 2) high-pass filter, 3) full wave rectifier, 4) low-pass filter, 5) downscaling, 6) analog multiplier, 7) integrator, 8) difference amplifier and 9) PI controller. The integration of these components is shown in Figure 6.1. The inputs to the sensor are the current through the coil and the voltage across the coil. The voltage input is converted to an estimated current. The current through the coil and the estimated current are passed through two identical FPFs. The outputs of these FPFs are passed through a difference amplifier. The output of the difference amplifier is an error signal. The error is the input to the PI controller. The output of the PI controller is proportional to the position of the rotor.

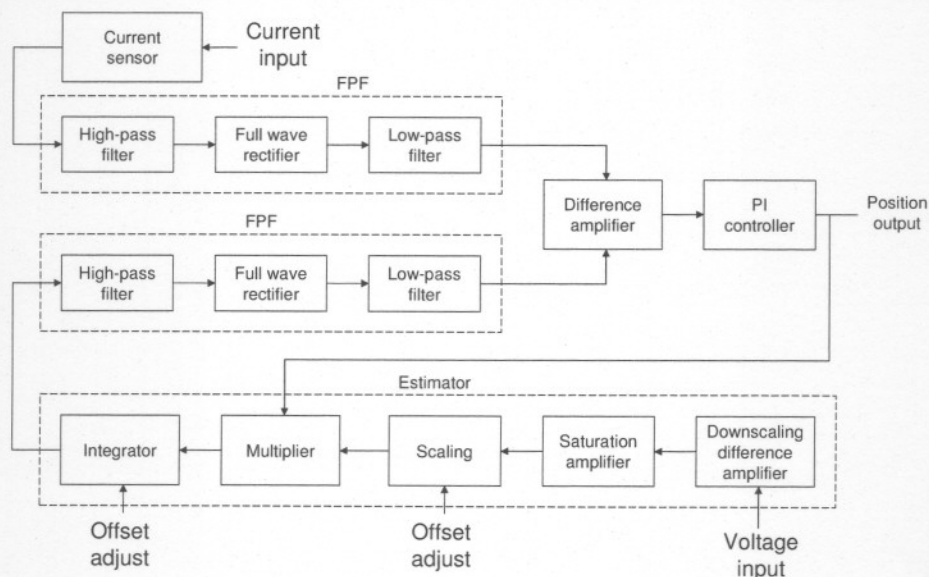


Figure 6.1: Integrated self-sensing method

6.1 Current sensor

The self-sensing method relies on the changing current through the electromagnet coils, thus a high bandwidth current sensor must be developed or purchased. The first consideration was a Hall Effect current sensor, but it is relatively expensive and the bandwidth of the sensor is too low for this application. The second consideration was to put a low resistance sense resistor in the path of the current and measure the voltage across the resistor. A high common mode rejection (CMR) operational amplifier is required, because of the high square wave voltage across the resistor. The sensed signal is also very small compared to the switching voltage, thus the signal is very noisy. The third option was to use a current transformer. The current conducting wire of the coil is passed through a ferrite core as the primary winding. The secondary coil of the transformer constitutes 200 turns that are terminated in a sense resistor R . Figure 6.2 illustrates the current transformer.

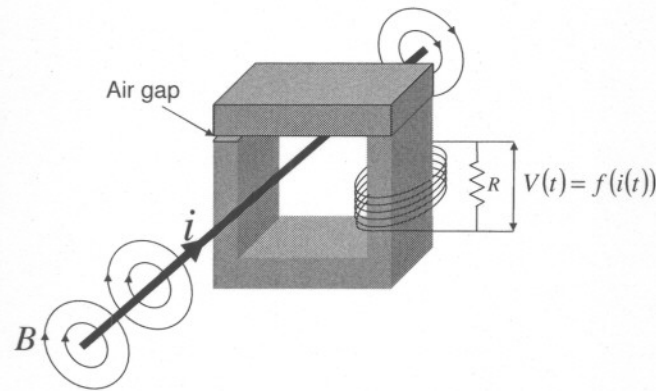


Figure 6.2: Current sensor

The changing current component through the wire induces a changing flux in the core of the current transformer. From Lenz law, the secondary windings will oppose the change in flux.

$$i_s = \frac{1}{N} i_p \quad (6.1)$$

where i_p is the alternating current through the primary winding and N is 200 turns.

The output of the current sensor is across the terminating resistor R .

The ferrite core must have a small air gap to prevent it from saturating due to the dc component of the current.

The output of the current sensor is satisfactory for self-sensing applications. It is important to note that the dc current component is lost. Figure 6.3 shows the hardware implementation of the current sensor. The white wire is the primary winding.

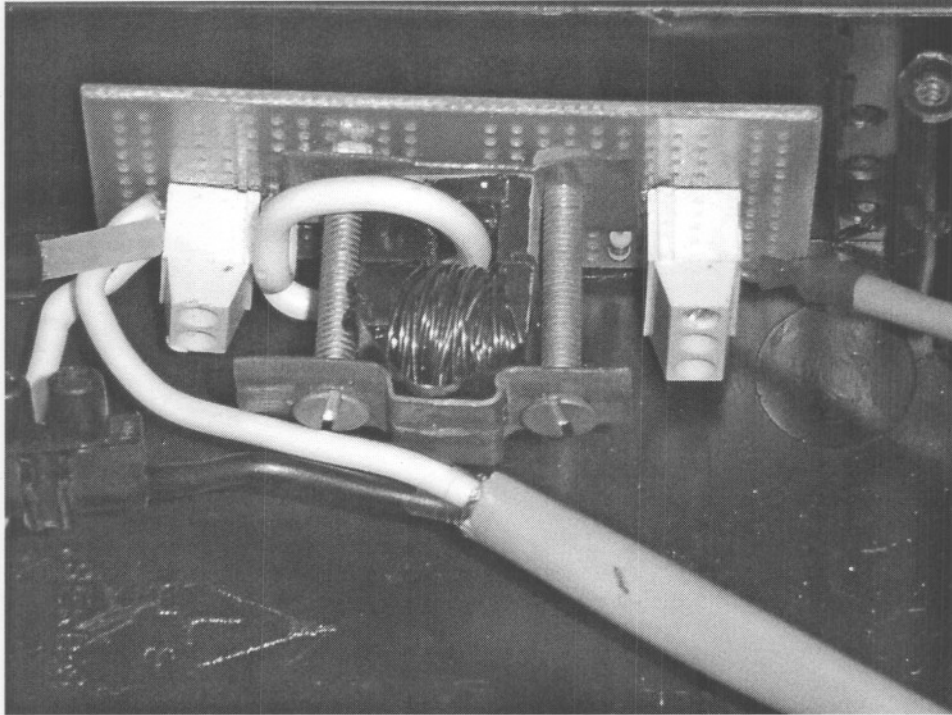


Figure 6.3: Current sensor implementation

6.2 High-pass filter

The purpose of the high-pass filter is to pass only the high frequency current components. From chapter 5 it was found that the cut-off frequency must be larger or equal to a tenth of the power amplifier switching frequency. In this case the switching frequency 100 kHz, thus the filter must be designed to pass all frequencies above 10 kHz. In this design cut-off frequency is chosen at 15 kHz.

The hardware used to implement the filter is the UAF42 integrated circuit (IC), which is a second order filter. This IC has trimmed internal capacitors (C_1) of 1000 pF and 50 k Ω (R_1) resistors. These internal components reduce the component drift when the temperature changes over time and minimise the number of external components. This IC can be operated as a high-, low- or band-pass filter.

Since the filter cut-off frequency is defined and the hardware is chosen, the filter can be developed. A Butterworth filter is used because of its zero ripple at the pass band. The realization of the UAF42 filter is shown in Figure 6.4.

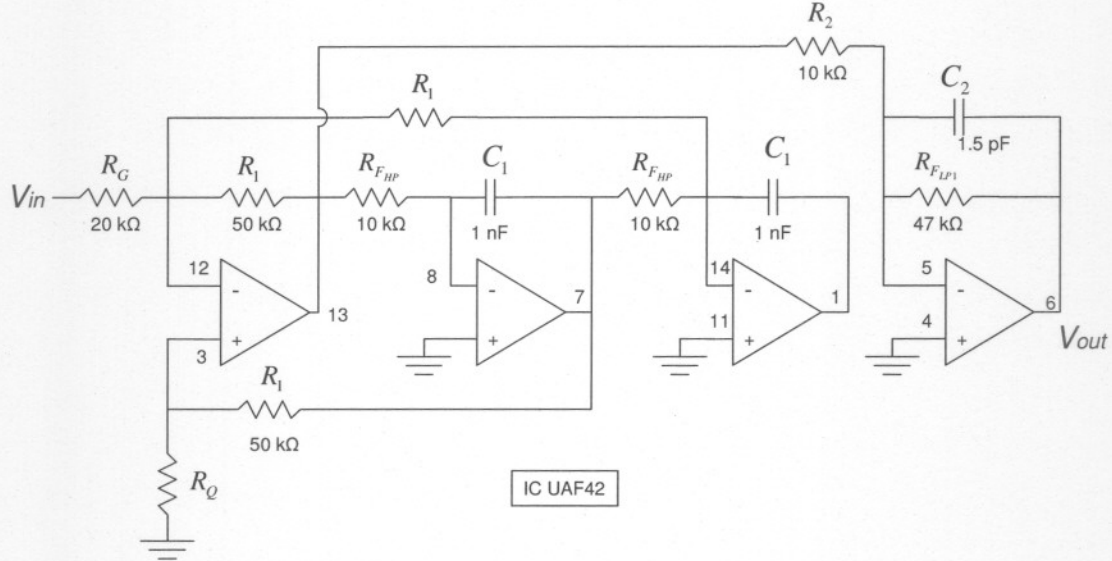


Figure 6.4: UAF42 High-pass filter

If it is assumed that x_1 is the output of the first operational amplifier on the left and x_2 is the output of the second operational amplifier, the state-space representation is as follows:

$$\begin{bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{bmatrix} = \begin{bmatrix} -2\zeta_{HP}\omega_{HP} & \omega_{HP} \\ -\omega_{HP} & 0 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} K_{HP}\omega_{HP} \\ 0 \end{bmatrix} \quad (6.2)$$

The cut-off frequency of the filter is determined by the external resistor $R_{F_{HP}}$, thus easily adjustable. The cut-off frequency calculation is shown in (6.3).

$$\omega_{HP} = \frac{1}{R_{F_{HP}} C_1} \quad (6.3)$$

With the cut-off frequency chosen to be 15 kHz and the internal capacitor C_1 constant at 1 nF. The value of $R_{F_{HP}}$ is determined to be 10.6 kΩ. The nearest standard value is 10 kΩ. The cut-off frequency with the standard value is at 15.92 kHz. From the simulated results in Figure 5.20 the sensor will still be stable at 15.92 kHz.

The damping factor ζ_{HP} from (6.4) determines the type of the filter. In this case a Butterworth filter is desired.

For a Butterworth filter the damping factor is equal to 0.707.

The damping factor is determined as follows:

$$\zeta_{HP} = \frac{1}{2} \left(2 + \frac{R_1}{R_G} \right) \left(\frac{R_Q}{R_Q + R_1} \right) \quad (6.4)$$

From (6.4) it is clear that the damping factor is a function of the external resistors, thus adjustable. From literature [15] R_G is chosen as 20 k Ω , R_1 is the known internal resistor and R_Q is determined to be 22.9 k Ω . The nearest series combination of standard values are 15 k Ω and 8.2 k Ω which adds to 23.2 k Ω . The new damping factor is determined to be 0.713.

The forward gain K_{HP} of the filter is determined by the ratio of the resistors R_1 and R_G as shown in (6.5).

$$K_{HP} = \frac{R_1}{R_G} \quad (6.5)$$

The forward gain is 2.5.

The fourth operational amplifier on the right is the final gain of the filter and the purpose of the parallel capacitor ($C_2=1.5$ pF) is to attenuate frequencies (ω_{HF}) which are much higher than the switching frequency. From practical experiments this switching noise is found up to 10 MHz.

The total gain of the high-pass filter is defined in (6.6) and is set to 11.75.

$$K_{HP_{total}} = K_{HP} \frac{R_{F_{LPI}}}{R_2} \quad (6.6)$$

The purpose of the gain $K_{HP_{total}}$ is to scale the output to a usable level, because the current input is relatively small (± 250 mV_{p-p}).

The transfer function of the high-pass filter is derived from ω_{HP} , ζ_{HP} and $K_{HP_{total}}$. The transfer function is shown in (6.7).

$$H_{HP}(s) = K_{HP_{total}} \left(\frac{s^2}{s^2 + 2\zeta_{HP}\omega_{HP}s + \omega_{HP}^2} \right) \left(\frac{\omega_{HF}}{s + \omega_{HF}} \right) \quad (6.7)$$

The bode diagram of the transfer function is shown in Figure 6.5.

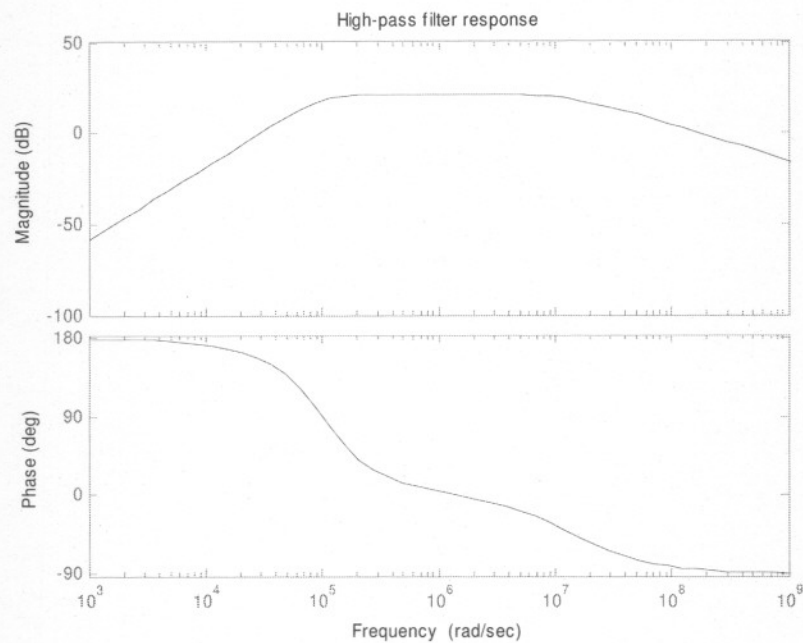


Figure 6.5: Bode diagram

Figure 6.6 shows the response of the high-pass filter simulated in Orcad. The Orcad simulation values are the same as determined above. The results from Figure 6.5 and Figure 6.6 are found to be satisfactory.

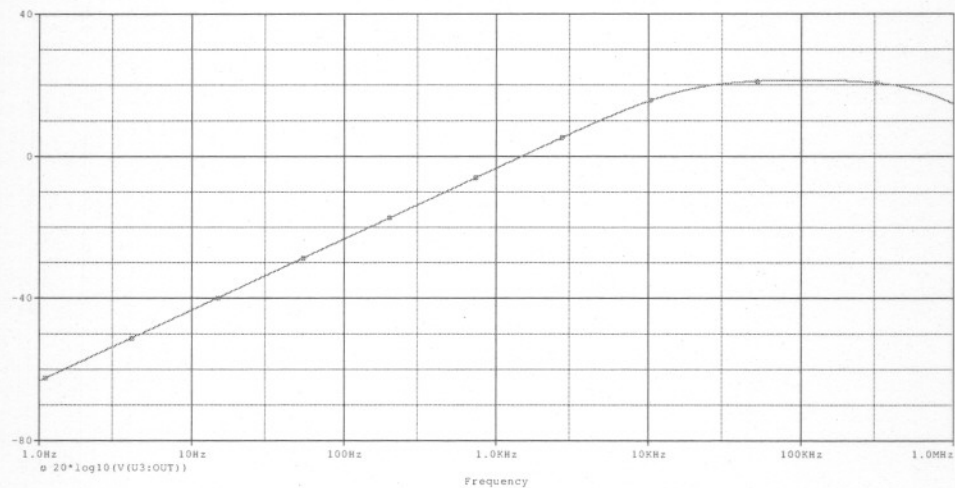


Figure 6.6: High-pass filter response

6.3 Full wave rectifier

The function of the full wave rectifier is to take the absolute value of the ac signal from the high-pass filter output. Different types of full wave rectifier circuits are available as discussed in chapter 4. In the self-sensing method the four diode circuit will be used. The circuit diagram in Figure 6.7 makes use of a TL082 IC.

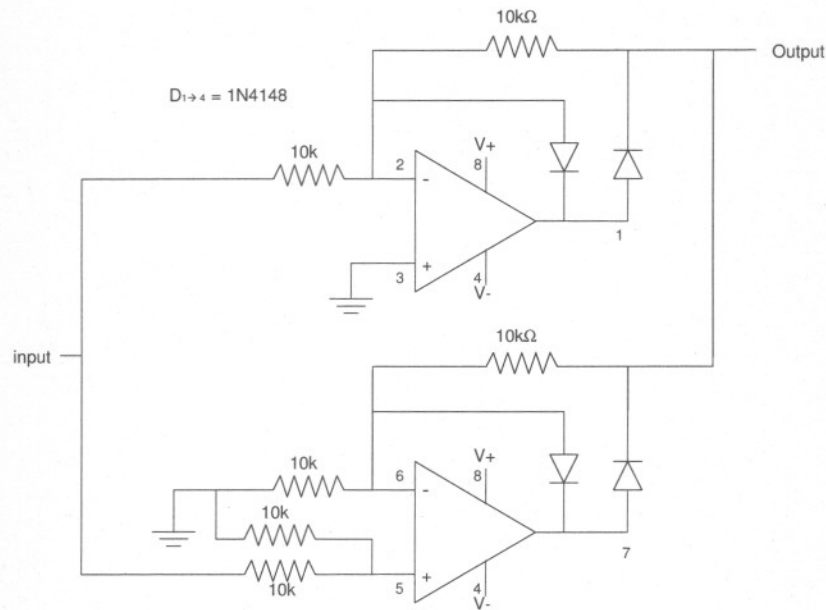


Figure 6.7: Full wave rectifier circuit

The output of the full wave rectifier is shown in Figure 6.8 with a sine wave as an input.

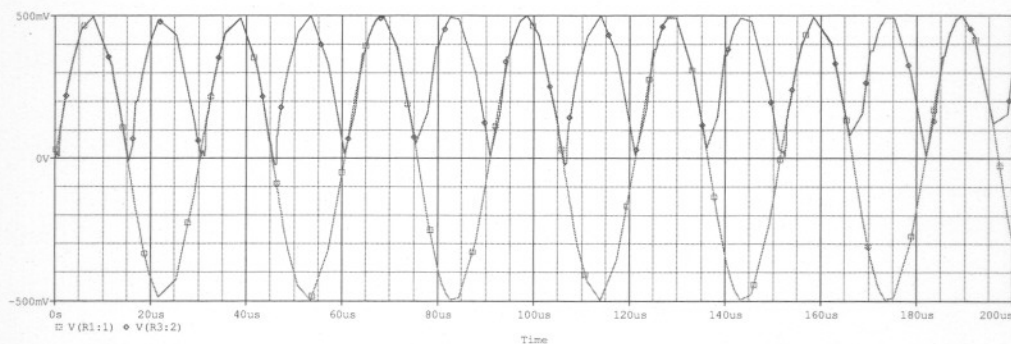


Figure 6.8: Full wave rectifier output

6.4 Low-pass filter

The purpose of the low-pass filter is to follow the peak values of the full wave rectifier. This has the result of a changing voltage at low frequencies which is a function of the peak values of the high frequency current.

From chapter 5 it was found that the cut-off frequency must be smaller or equal to a tenth of the power amplifier switching frequency. As mentioned before the switching frequency is 100 kHz, thus the filter cut-off must be designed at 10 kHz or below. The low-pass filter cut-off frequency is chosen at 6 kHz. The UAF42 IC will be used in the low-pass filter configuration.

Now that the filter cut-off frequency is chosen, the filter could be designed. The low-pass filter circuit diagram is shown in Figure 6.9.

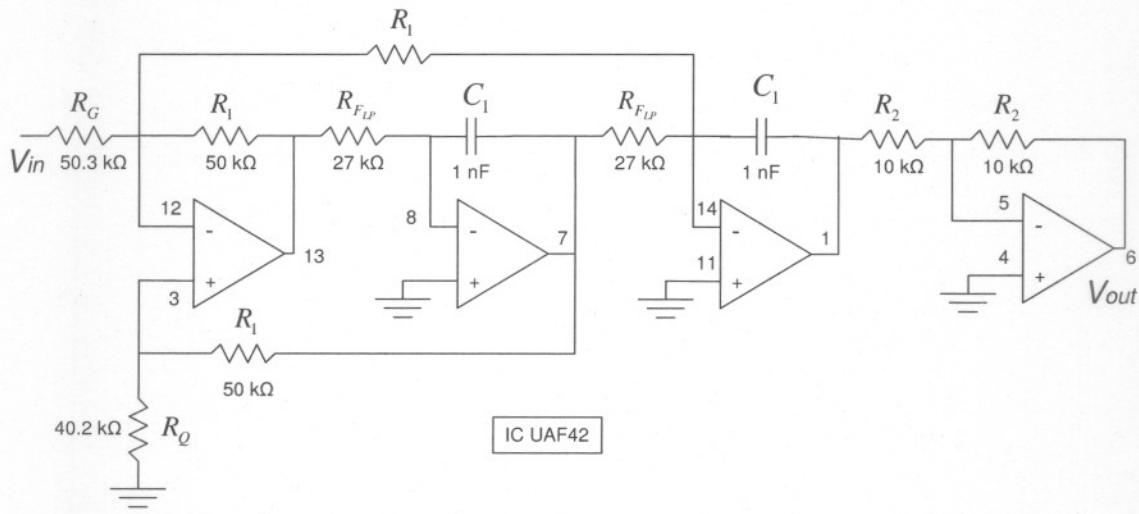


Figure 6.9: UAF42 Low-pass filter

If the same state-space is used as in high-pass filter case, then the cut-off frequency of the low-pass filter is also determined by the external resistors.

The cut-off frequency of the low-pass filter is shown in (6.8).

$$\omega_{LP} = \frac{1}{R_{FLP} C_1} \quad (6.8)$$

The value of R_{FLP} can now be determined, since C_1 (1 nF) is known. From (6.8) R_{FLP} is determined to be 26.52 kΩ. The nearest standard value is 27 kΩ. The cut-off frequency is at 5.89 kHz if the standard value is used. From the simulated results in Figure 5.20 the sensor will still be stable at 5.89 kHz. In this case the damping factor is set to 0.651 which makes the filter underdamped.

$$\zeta_{LP} = \frac{1}{2} \left(2 + \frac{R_1}{R_G} \right) \left(\frac{R_Q}{R_Q + R_1} \right) \quad (6.9)$$

The forward gain K_{LP} of the filter is also determined by the ratio of the resistors R_1 and R_G as shown in (6.10).

$$K_{LP} = \frac{R_1}{R_G} \quad (6.10)$$

The forward gain is set to 1.

The total gain of the high-pass filter is defined in (6.11) and is also set to 1. The total gain is set to 1 to prevent the saturation of the operational amplifier.

$$K_{LP_{total}} = K_{LP} \frac{R_{F_{LP}}}{R_2} \quad (6.11)$$

The transfer function of the low-pass filter is derived from ω_{LP} , ζ_{LP} and $K_{LP_{total}}$. The transfer function is shown in (6.12).

$$H_{LP}(s) = K_{LP_{total}} \left(\frac{\omega_{LP}^2}{s^2 + 2\zeta_{LP}\omega_{LP}s + \omega_{LP}^2} \right) \quad (6.12)$$

The bode diagram of the low-pass filter transfer function is shown in Figure 6.10.

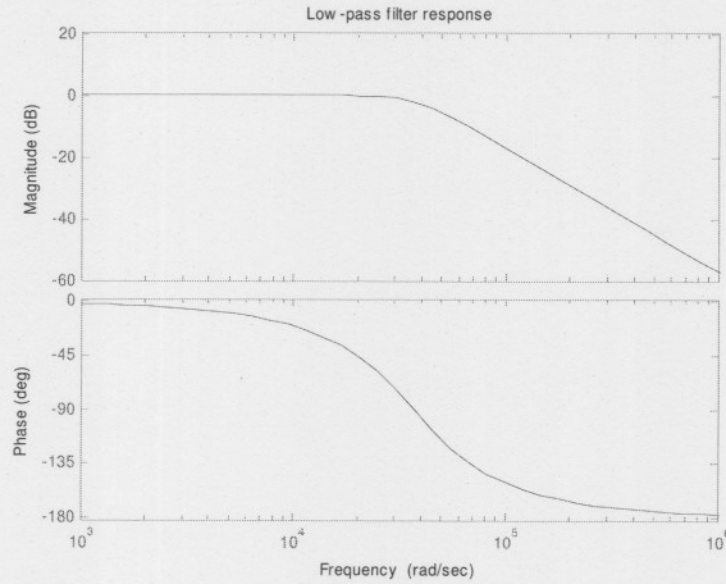


Figure 6.10: Bode diagram

Figure 6.11 shows the response of the low-pass filter. The external component values, which was calculated above, is used in the Orcad simulated.

The small gain just before the cut-off frequency should not be a problem, because the bandwidth of the mechanical model is about 500 Hz. From the bode diagram and the simulated result it is found that the low-pass filter performs to required specifications.

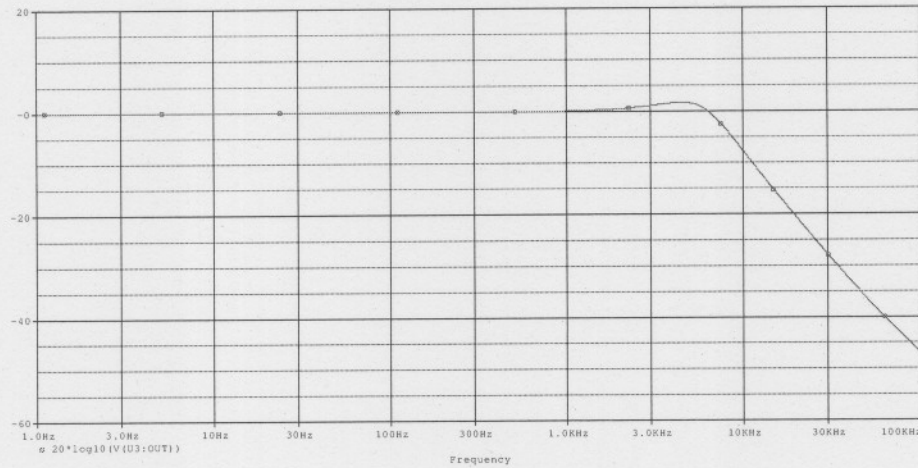


Figure 6.11: Low-pass filter response

6.5 Downscaling

The output of the power amplifier is a square wave with a 100 kHz frequency and amplitude of ± 310 V. This switching voltage of the power amplifier must be downscaled to be in the operating range of the operational amplifier. The operating voltage of the operational amplifier is ± 18 V. By dividing the ± 310 V voltage by 66, the operational amplifier operates normally. If the scaling factor is smaller, then common mode voltage becomes too large and the amplifier does not operate normally. This downscaling is done by a difference amplifier shown in Figure 6.12. The ± 310 V voltage is sensed across the coil of the AMB. The output of the difference amplifier is in the range of 5 V.

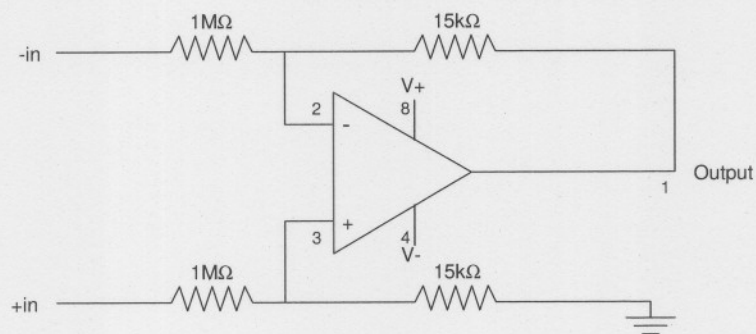


Figure 6.12: Scaling difference amplifier

Now that the power amplifier output voltage is in the operating range of the operational amplifiers the output voltage of the difference amplifier is now saturated, the amplification factor is at least three times. The reason for the saturation is that the amplifier voltage is not an ideal square wave. There is overshoot involved and the ideal ± 310 V output varies with 6-7 volts at 100 Hz. This 100 Hz component is caused by the rectified 50 Hz mains voltage which charges the power amplifier capacitor. The saturation circuit is shown in Figure 6.13.

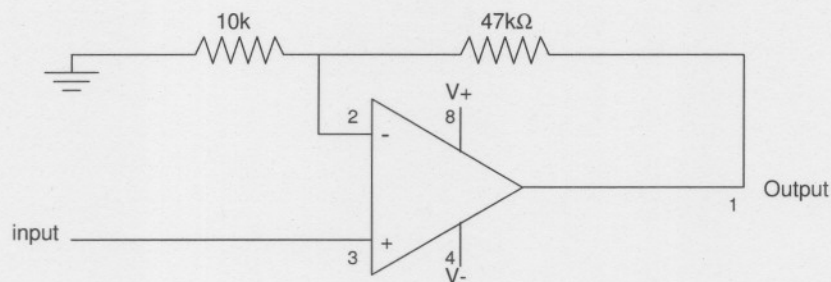


Figure 6.13: Saturation amplifier

The output of the saturation amplifier is larger than the input specification of the analog multiplier. This saturated voltage is divided to give an output below ± 10 V. This amplifier is constructed from a TL071 which has the additional function of a zero adjust. The reason for the zero adjust, is that the output of this operational amplifier determines the input of the integrator. If the square wave has offset and an ideal integrator is used the output of the integrator saturates. The adjustable offset amplifier is shown in Figure 6.14.

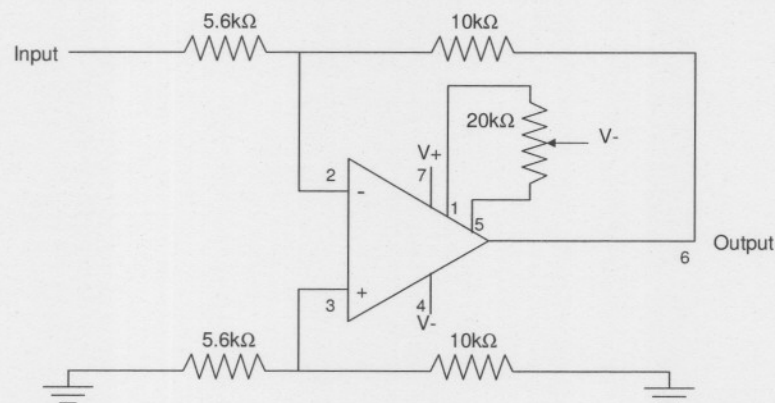


Figure 6.14: Null adjustable amplifier

6.6 Analog multiplier

The analog multiplier is required in the self-sensing technique to implement the parameter estimator as described in chapter 5. The MPY634 IC with a 10 MHz bandwidth will be used. As mentioned before the input requirements are that both of the inputs must be in the range of ± 10 V. The analog multiplier circuit diagram is shown in Figure 6.15.

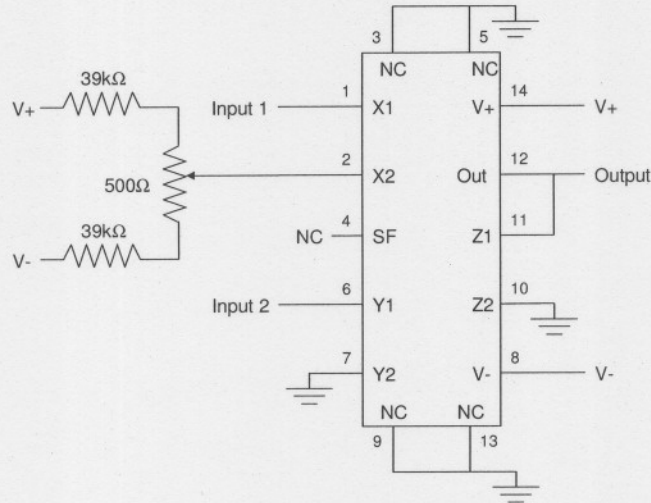


Figure 6.15: Analog multiplier

The transfer function of the multiplier is given in (6.13).

$$Out(t) = \frac{(X1(t) - X2(t))(Y1(t) - Y2(t))}{10} \quad (6.13)$$

All the no connection (NC) points are connected to ground to improve the stability of the multiplier.

6.7 Integrator

The integrator shown in Figure 6.16 is used to simulate the AMB model. The integrator transforms the square wave, which is directly related to the switching intervals of the power amplifier, to a triangular current wave. In this design the TL071 is also used to improve the zero offset problem as discussed.

The transfer function of the integrator is shown in (6.14).

$$H(s) = -\frac{R_2/R_1}{1 + sC_1R_2} \quad (6.14)$$

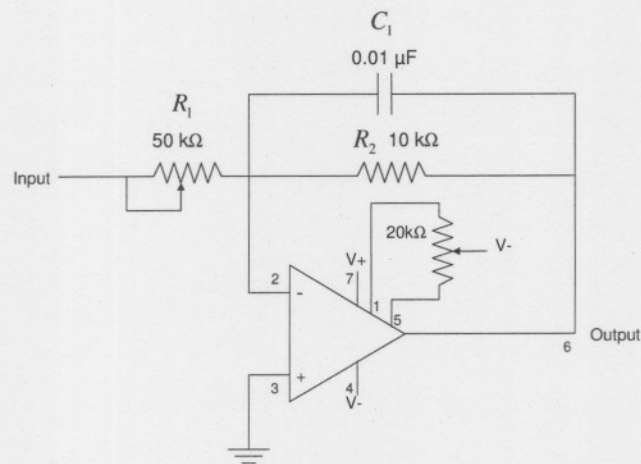


Figure 6.16: Integrator

The purpose of the adjustable resistor is to vary the gain of the integrator in order to match the real current.

Figure 6.17 shows what would happen if the 100 Hz component was not removed by the saturation procedure. The insert shows that the output of the integrator is of triangular form.

The output of the integrator is passed through the same FPF which consists of a high-pass filter, full wave rectifier and low-pass filter. This is shown in Figure 6.1.

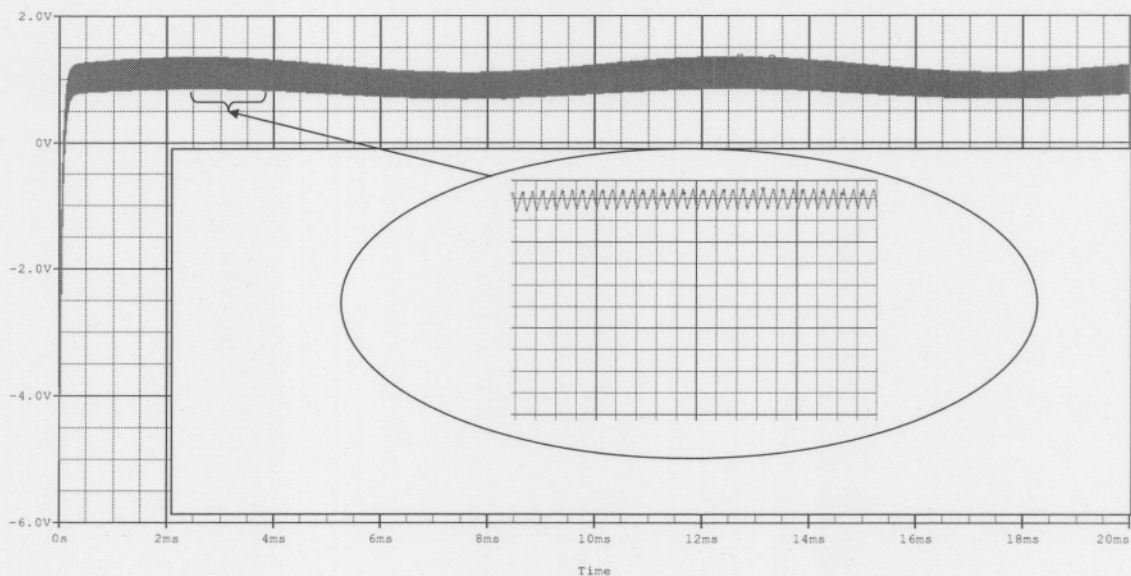


Figure 6.17: Integrator with 100 Hz component

6.8 Difference amplifier

The difference amplifier subtracts the output of the estimated current low-pass filter from the output of the real current low-pass filter. The difference amplifier is in the self-sensing method one of the most important parts. The self-sensing method relies on the error between the estimated current and the real current. Another reason for the requirement of a high precision difference amplifier is that the values that are subtracted are relatively small and the error output even smaller. It was determined from literature that the INA105 has high common mode rejection, unity gain and is relatively stable under temperature variations.

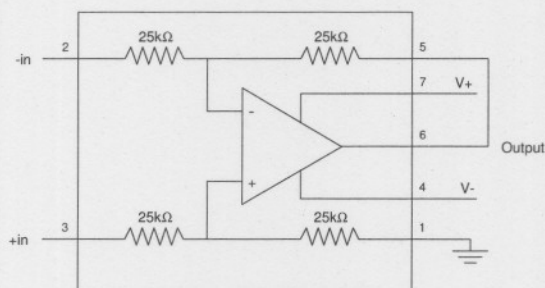


Figure 6.18: Precision difference amplifier

6.9 PI controller

The PI controller is derived from the simulated model. It was found from the simulation model that the K_p value must be smaller than 10^{-3} and the K_i value must be smaller than 4. It will be better to design the PI controller to be fully adjustable.

The circuit diagram of the PI controller is shown in Figure 6.19.

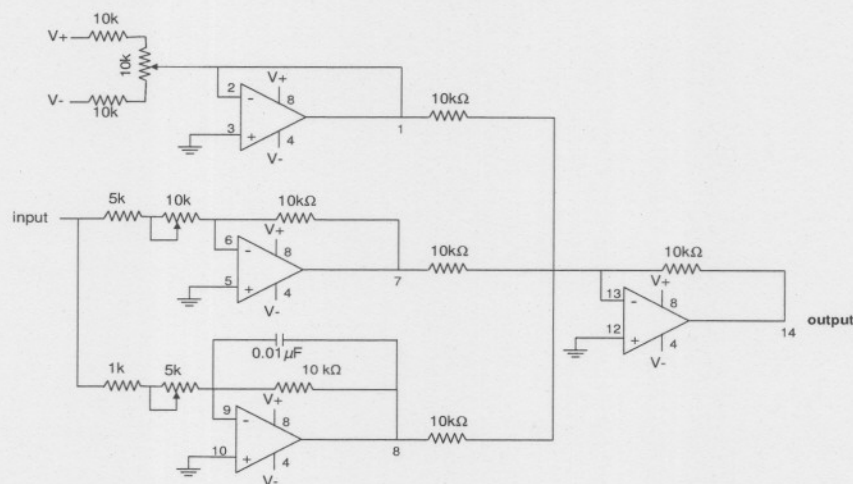


Figure 6.19: PI controller

The output of the controller is proportional to the position. By scaling the output of the controller the position is estimated.

6.10 Integrated self-sensing model

Figure 6.20 shows the scaling circuit. This circuit scales the ± 310 V voltage input from the power amplifier to the operating level of the operational amplifiers. The input of the current sensor is put through a voltage follower to prevent signal loading. This signal is then scaled to the desired voltage level by a noninverting amplifier.

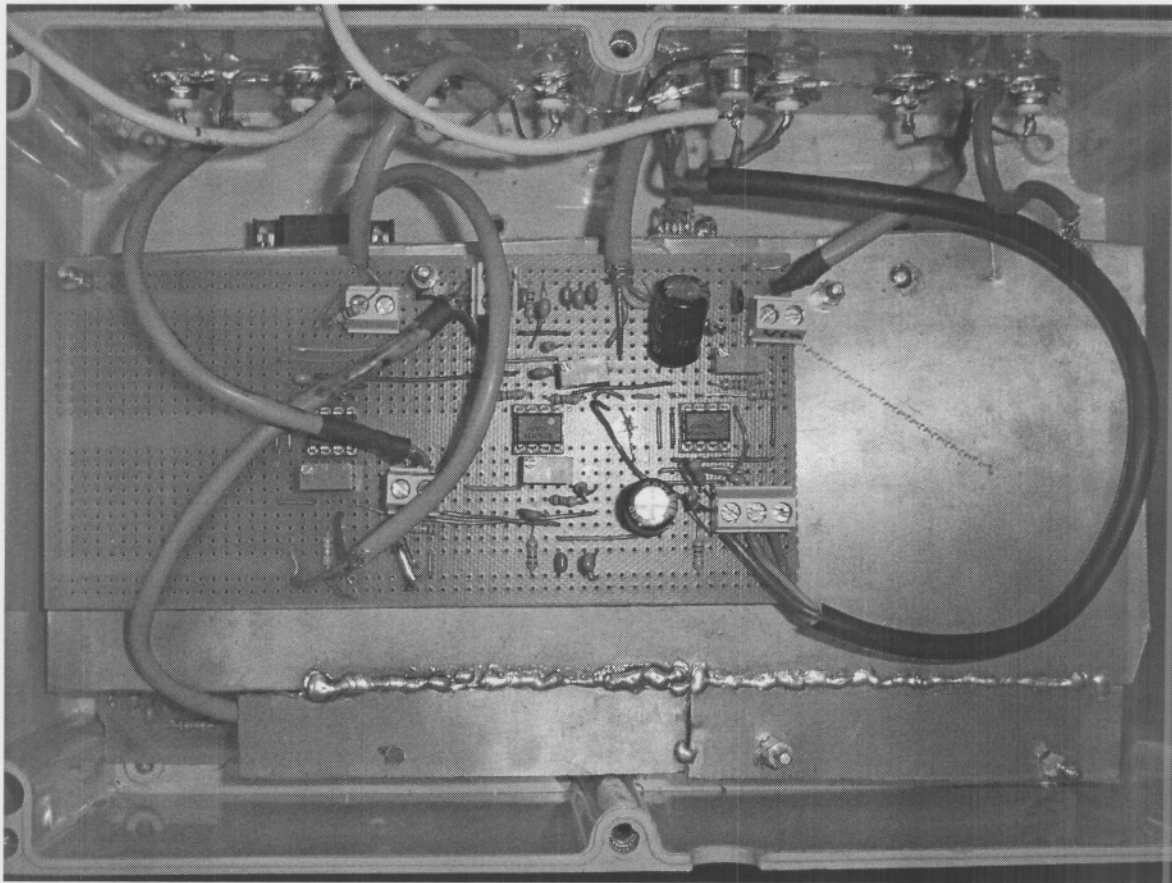


Figure 6.20: Scaling circuit

Figure 6.21 shows the self-sensing circuit. The left circuit board is the FPF for the real current input and the right circuit board is the FPF for the estimator. The circuit in the centre is the estimator and PI controller

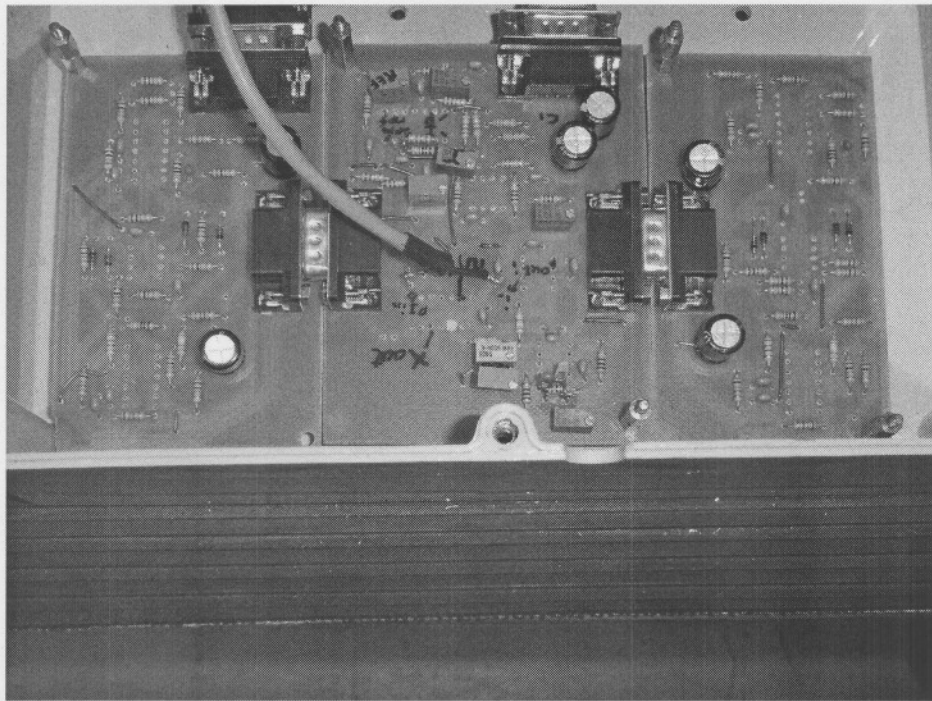


Figure 6.21: Self-sensing circuit

From the left in Figure 6.22, the power amplifier, current sensor and the self-sensing integration is shown.

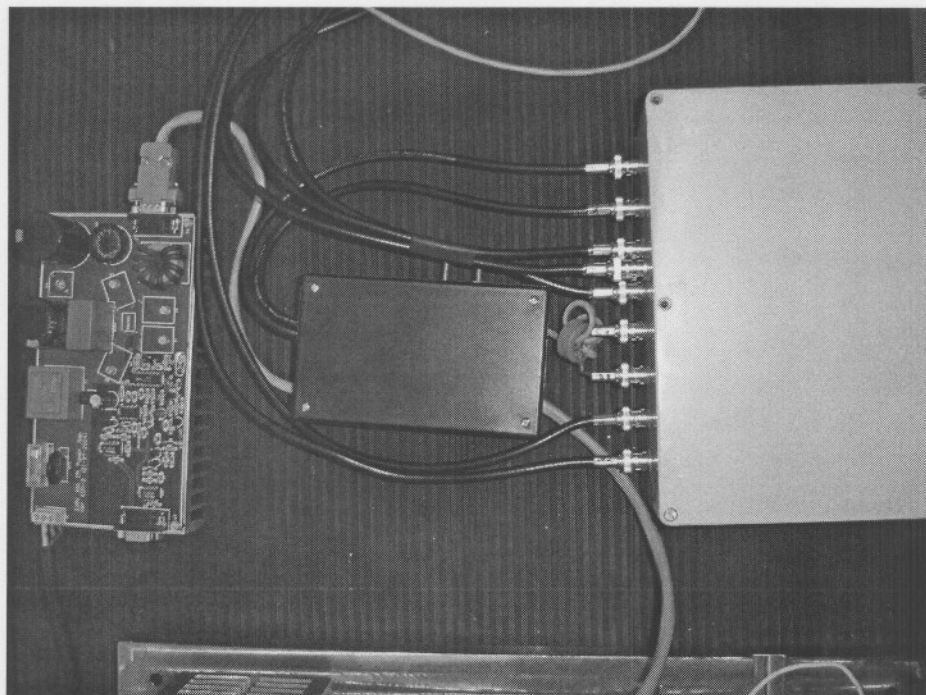


Figure 6.22: Power amplifier, current sensor and self-sensing sensor

Figure 6.23 shows the integration of the self-sensing sensor and the right stator of the heteropolar AMB.

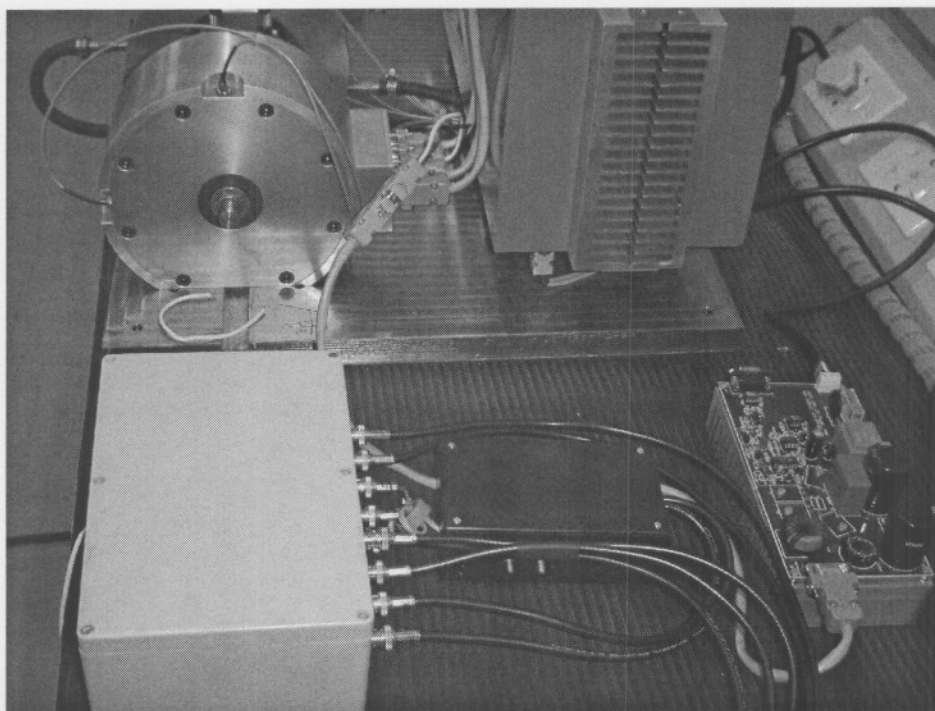


Figure 6.23: Self-sensing sensor and heteropolar AMB

6.11 Conclusion

The current sensor works well in the self-sensing application. In a case where the dc component is of importance the sensor will not work, because the dc current component is lost. The air gap is just a precaution to prevent the saturation of the ferrite core.

In the case of the high-pass filter it is important to remember that the gain is 11.75, thus the maximum input must be between ± 1 V. The attenuation of the high-pass filter, shown in Figure 6.5, looks poor but keep in mind that filter has a gain of 21.5 dB.

The downscaling difference amplifier must be further investigated to understand the irregular operation if the scaling factor is smaller than 66.

Further research must be done to evaluate the effect that the slew rate of the operational amplifiers has on the performance of the self-sensing method.

The integrator is supplied with a variable gain. This has the advantage to scale the estimated current to the same level as the real current component.

The integration of the different components was successful. The self-sensing method could now be evaluated in the next chapter.

Evaluation of Self-sensing Hardware

In this chapter the hardware of the self-sensing method is evaluated. These results determine the performance and quality of the sensor. Future work will be derived from the problems encountered in this system. The following components will be evaluated: 1) current sensor, 2) forward path filter, 3) integrator, 4) static sensor output, 5) dynamic sensor output and 6) nonlinearities.

7.1 Current sensor

The current sensor is evaluated by comparing the output of the current sensor to the current measured by a current probe. From the results shown in Figure 7.1 it is found that the current sensor's performance is adequate for this application. Again keep in mind that the dc component of the current is lost.

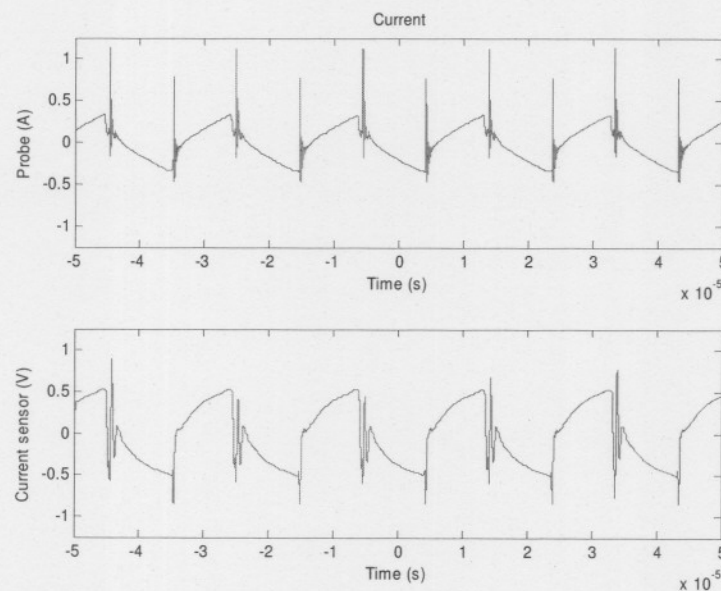


Figure 7.1: Current through coil

The output of the current sensor was put through a voltage follower to prevent signal loading. The output of the voltage follower is then put through a non-inverting amplifier to increase the current signal to a more convenient value.

7.2 Forward path filter

The forward path filter (FPF) is evaluated by passing a triangular wave of 100 kHz from the signal generator through the FPF. The output of the FPF was found to be satisfactory with the ideal signal passed through the filter. In the case where all the amplifiers are switched on and the current wave is generated by a non-ideal amplifier, the output of the FPF is very noisy. These noisy signals are shown in Figure 7.2. From Figure 7.2 it is determined that the noise amplitude increases as the air gap increases. This is due to the decrease in inductance of the coil. This decrease in inductance increases the current load on the power amplifier.

The solid line shows when the air gap is at its minimum and the dashed line is when the air gap is at its maximum.

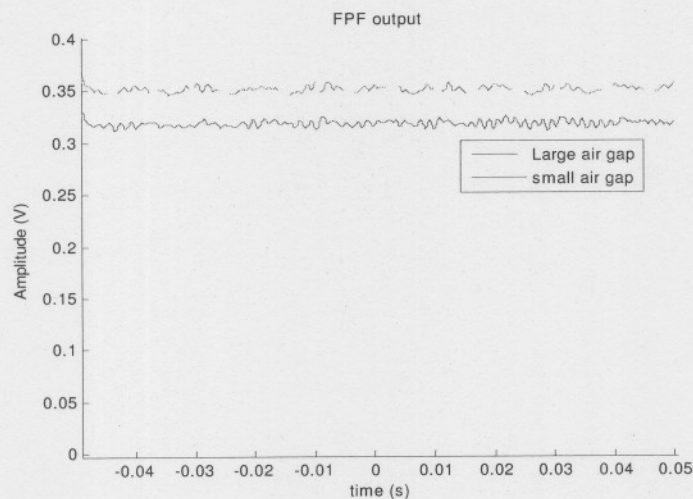


Figure 7.2: FPF output

This noise component cannot be filtered, because it is in the operating range of the control signal. If a band-stop filter is used, the control signal will have a phase shift and a small delay. This phase shift and delay in the control signal will result in instability. Subtraction could be used to eliminate the noise by subtracting the same noise component from the output. This method may work, but the signal which must be subtracted is a function of the air gap. The magnetic coupling between the horizontal and vertical coils increases the noise level. This magnetic coupling between the self-sensing coil and the two horizontal coils makes it even more impossible to implement this method due to the increase in noise.

7.3 Integrator

The integrator is evaluated by passing the scaled square wave of the power amplifier through the integrator. The output of the integrator is satisfactory with all the amplifiers switched on. This signal is shown in Figure 7.3. The capacitance of the analog integrator was refined to simulate the current more accurately.

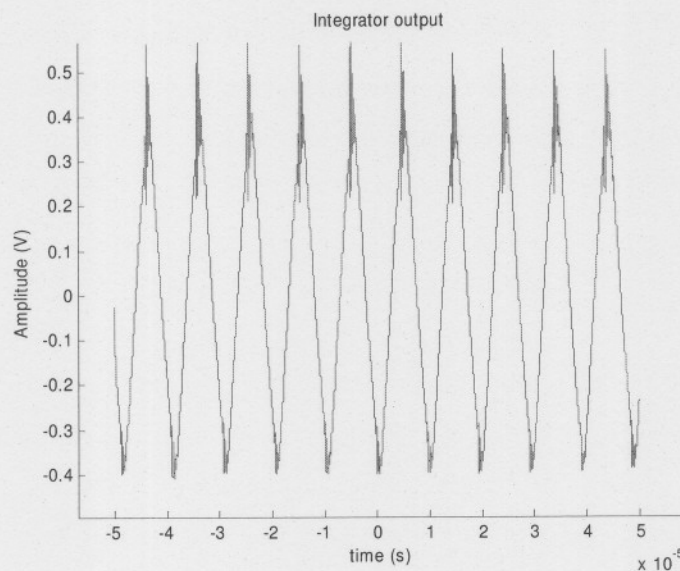


Figure 7.3: Integrator output

7.4 dSpace® implementation

The real position of the rotor is determined by high precision eddy current sensors. The outputs of the self-sensing sensor and eddy current sensor are not directly proportional to position. These outputs must first be rescaled to position signals. dSpace® is used as interface between Simulink® and the AMB model. The outputs of the eddy current sensor and self-sensing sensor are converted to digital form by using a 16 and 12 bit analog to digital converters (ADCs) respectively. The analog to digital conversion is done by the internal dSpace® board. This internal digital signal processing (DSP) board is linked to ControlDesk®. Figure 7.4 shows the graphical user interface (GUI) of ControlDesk®.

All the variables used in the AMB Simulink® model can be directly adjusted by using ControlDesk®. The outputs of the Simulink® model is converted to analog signals by the dSpace® board. This is done by 16 bit digital to analog converters (DACs). These outputs are fed to the control inputs of the power amplifiers.

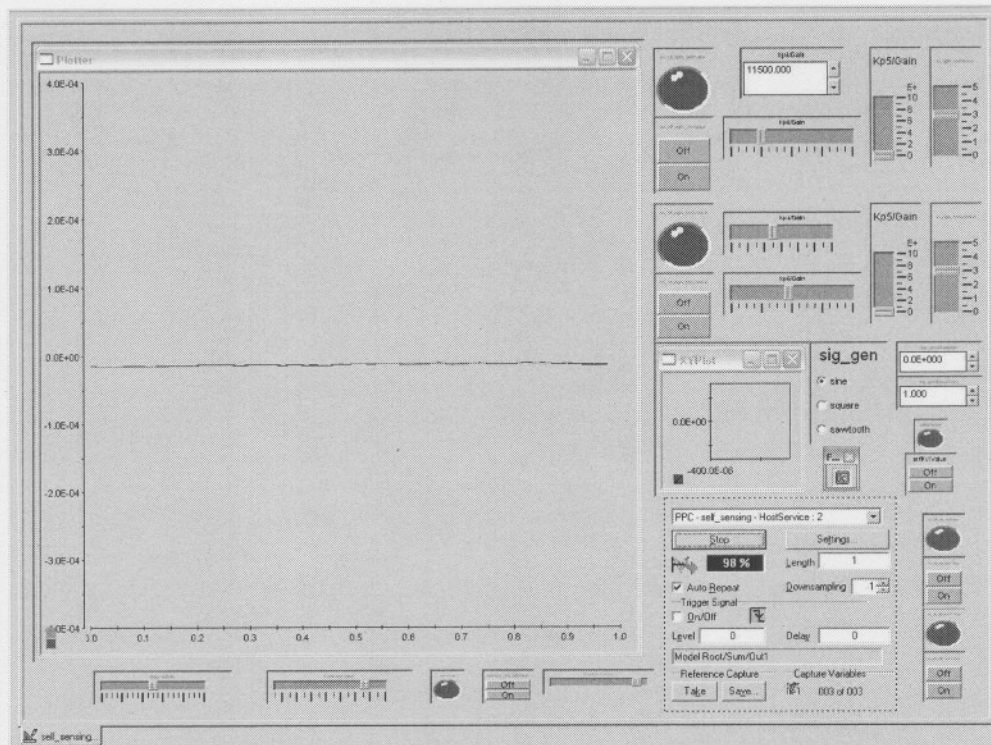


Figure 7.4: ControlDesk® GUI

7.5 Static response

In this section the output of the self-sensing method is evaluated. The PI controller is the output of the sensor, thus the PI controller is evaluated in conjunction with the sensor output. From practical results it was determined that the PI controller has an affect on the bandwidth. The resistor value R_f in Figure 6.19 was reduced from 50 k Ω to 10 k Ω . The bandwidth of the sensor was increased considerably.

The bottom coil of the AMB is used for position estimation and the top coil is used for control. The first method of evaluation is to power up the bottom coil's power amplifier with all the other amplifiers switched off. A constant current is passed through the coil without control feedback. The air gap is kept constant. The sensor output is shown in Figure 7.5. From the output of the sensor as shown in Figure 7.5 it is found that the output has a large noise component. From close inspection the noise is found to be a 700 mV, 100 Hz signal. It was at first thought that the sensor was not earthed correctly. With intense probing it was found that the power amplifiers have a 100 Hz component. The possible source of the noise has been located, but must be verified. This information was verified using two signal generators.

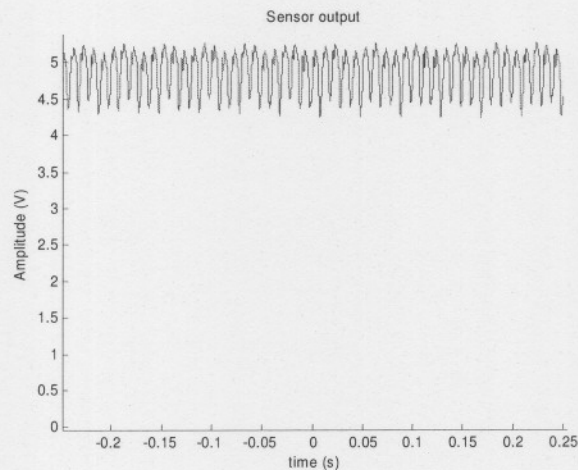


Figure 7.5: Sensor output with noise

The triangular waveform and the square waveform generated by the signal generators are used as inputs to the sensor. The results verified that the power amplifier was the source of the noise. The output of the sensor is shown in Figure 7.6 with ideal inputs. The noise is reduced by a factor nine.

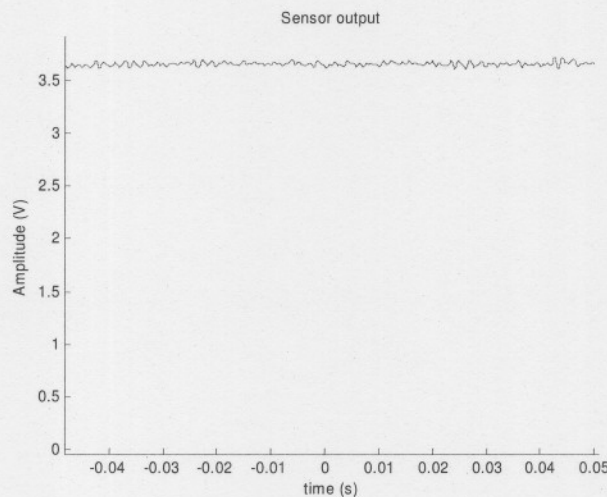


Figure 7.6: Sensor output with ideal inputs

The second method of evaluation is to switch on all the power amplifiers. Again the bottom coil is used for position estimation and the top coil, along with the two horizontal coils, are used for control. For the vertical control the self-sensing position is used as feedback. The power amplifier's triangular current and square voltage signals are used as inputs to the sensor. The air gap is kept constant. The sensor's output is shown in Figure 7.7. The results are scaled in dSpace.

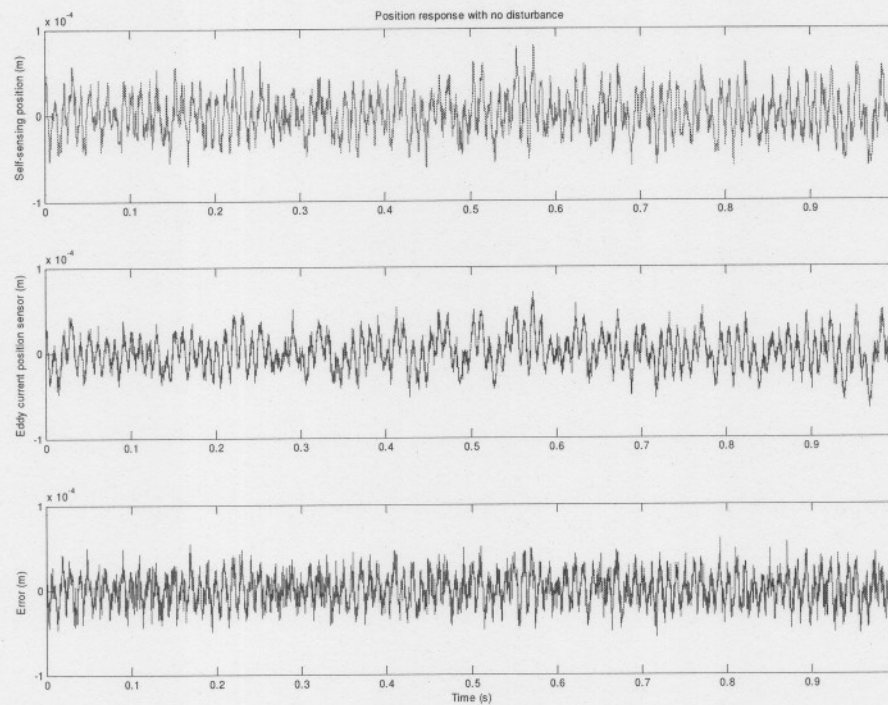


Figure 7.7: Sensor output with all amplifiers on

From Figure 7.7 it is clear that the output is even worse than with the first evaluation method. This increase in noise is the result of magnetic cross coupling from the horizontal coils to the bottom coil.

7.6 Dynamic response

In the previous section the noise was analysed. In this section the dynamic response is evaluated. All the following results have the same noise component as in section 7.5. The dynamic response is divided in two categories: 1) step response 2) sinusoidal response and 3) ramp response. It is important to note that the self-sensing output and the eddy current sensor's output is rescaled to position. This rescaling is done in dSpace[®]. For all the results that follow in section 7.6.1, 7.6.2, 7.6.3 and 7.6.4 the bottom coil is used for position estimation and the top coil is used for control. The bottom power amplifier is given a reference voltage. This reference voltage is just large enough to activate the power amplifier. This low reference voltage reduces the noise as discussed section 7.2.

7.6.1 Step response

The sensor's dynamics are best represented by a step response. Only low frequency step disturbances will be given to the system. The reason for this is that the mechanical bandwidth of the model limits the step response frequency. 1 Hz, 5 Hz and 10 Hz step responses are evaluated.

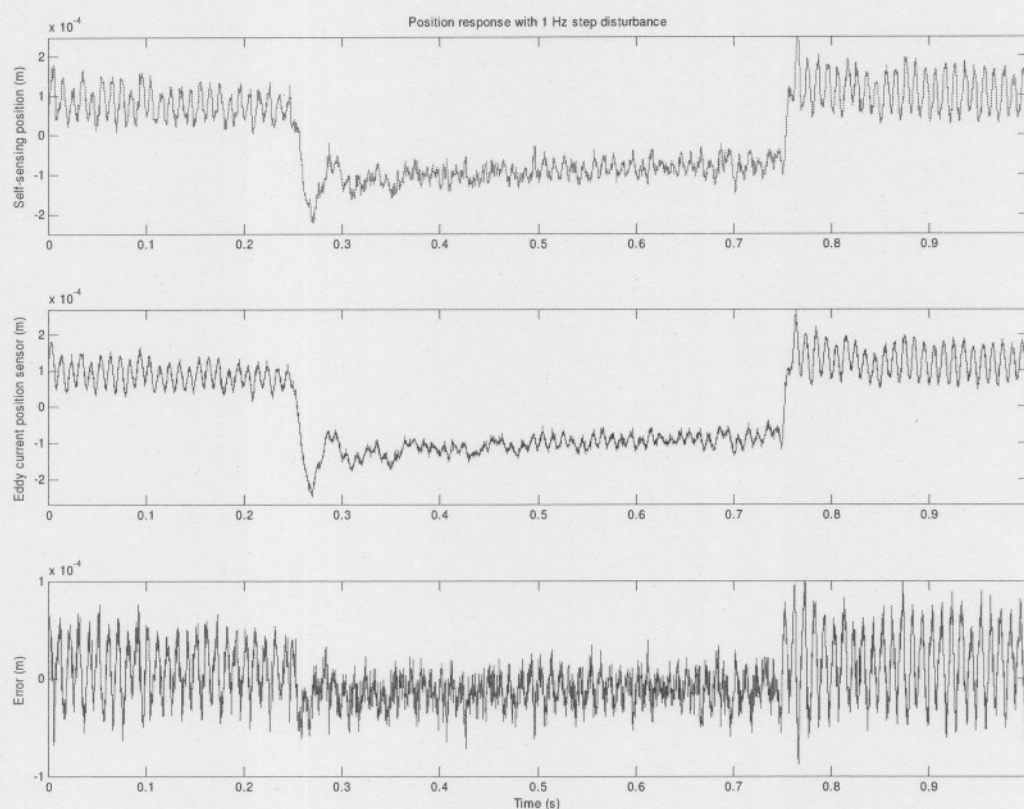


Figure 7.8: Sensor output with a 1 Hz step disturbance

Figure 7.8 shows a step disturbance of $30\text{ }\mu\text{m}$ at a frequency of 1 Hz. A relatively large noise component is present. The eddy current sensor signal also contains noise. The reason for this is that the whole system reacts on the noise of the self-sensor as if the noise is position. From the error signal in Figure 7.8 the noise is once more proven to be lower at small air gaps. In this case the air gap is smaller at time 0.28 s to 0.75 s.

Figure 7.9 shows a step disturbance of $30\text{ }\mu\text{m}$ at a frequency of 5 Hz.

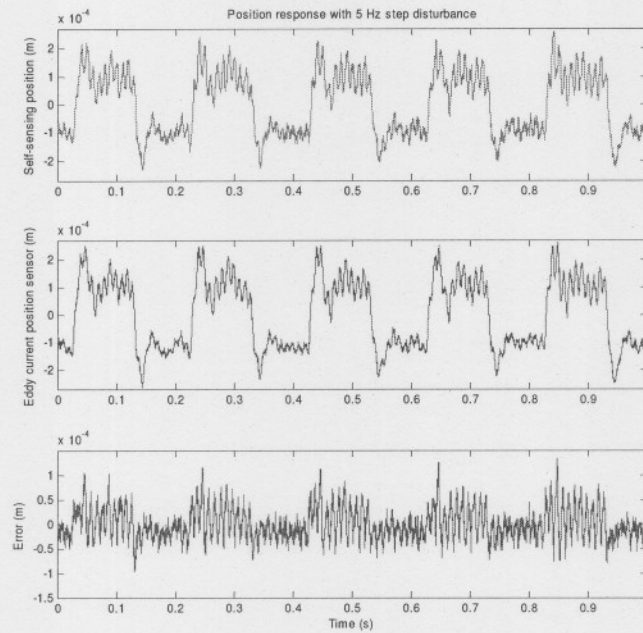


Figure 7.9: Sensor output with a 5 Hz step disturbance

Figure 7.10 shows a step disturbance of $30\text{ }\mu\text{m}$ at a frequency of 10 Hz. The bandwidth of the mechanical model starts to have an affect on the position response. Due to the bandwidth limitation a sinusoidal evaluation will be done in the next section.

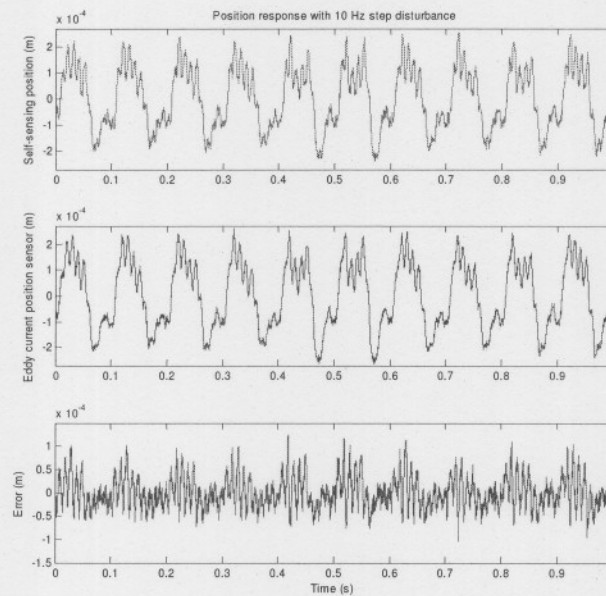


Figure 7.10: Sensor output with a 10 Hz step disturbance

It is important to note that the error is much smaller when the air gap is small.

These step response results confirm the noise effect at large air gaps. The bandwidth of the sensor was also proven to be large enough to suspend the rotor at steps up to 10 Hz.

7.6.2 Sinusoidal response

The purpose of the sinusoidal response is to verify that the sensor performs adequately at high frequency. Again the self-sensing sensor will be evaluated from low frequencies upwards.

Figure 7.11 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 10 Hz given to the system. Once more it is noticeable that the noise is larger where air gap is large.

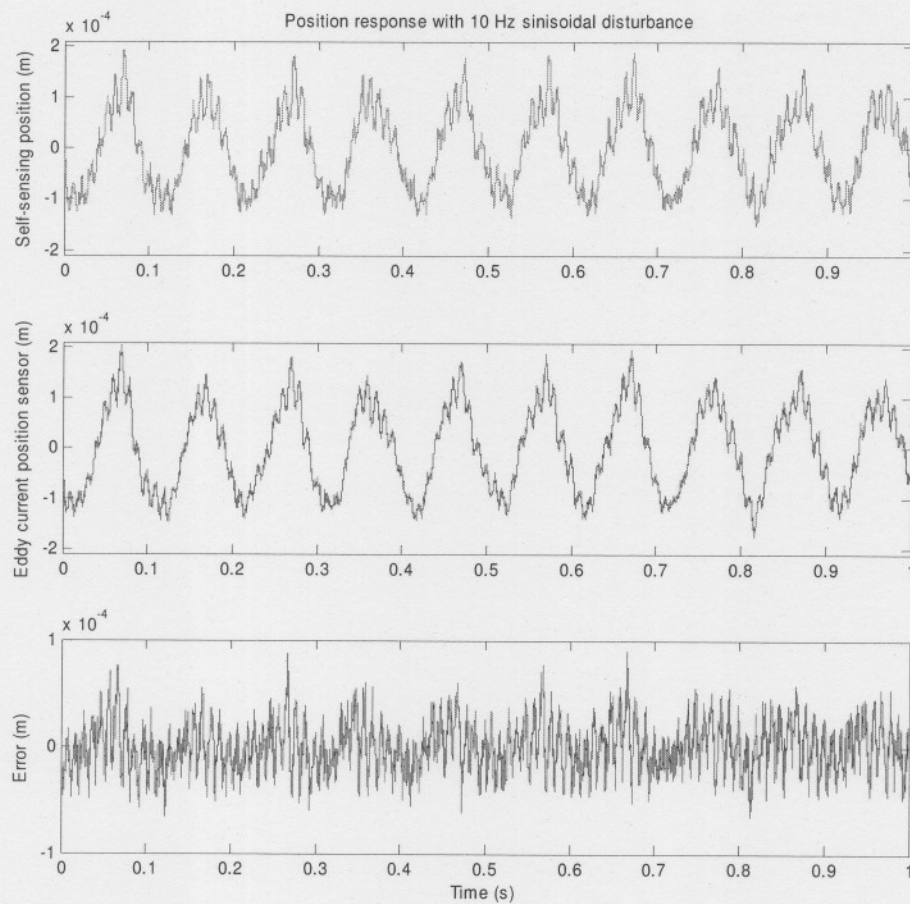


Figure 7.11: Sensor output with a 10 Hz sinusoidal disturbance

Figure 7.12 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 20 Hz.

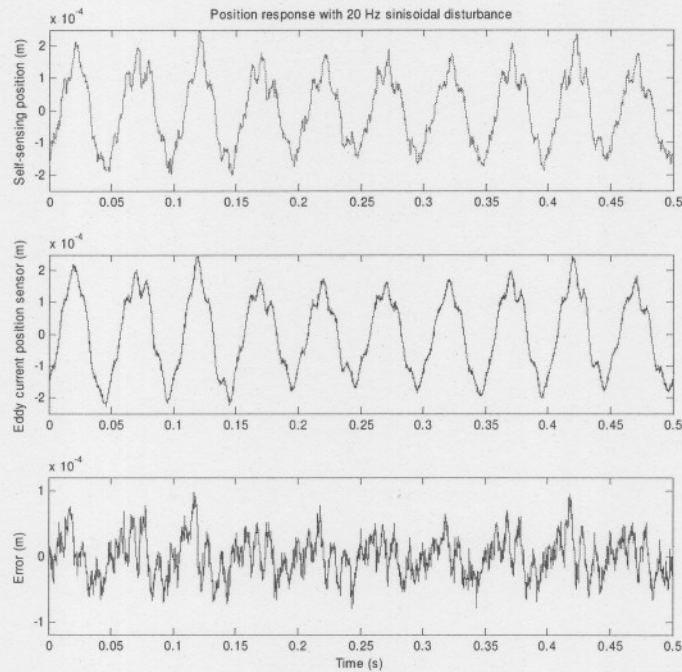


Figure 7.12: Sensor output with a 20 Hz sinusoidal disturbance

Figure 7.13 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 30 Hz.

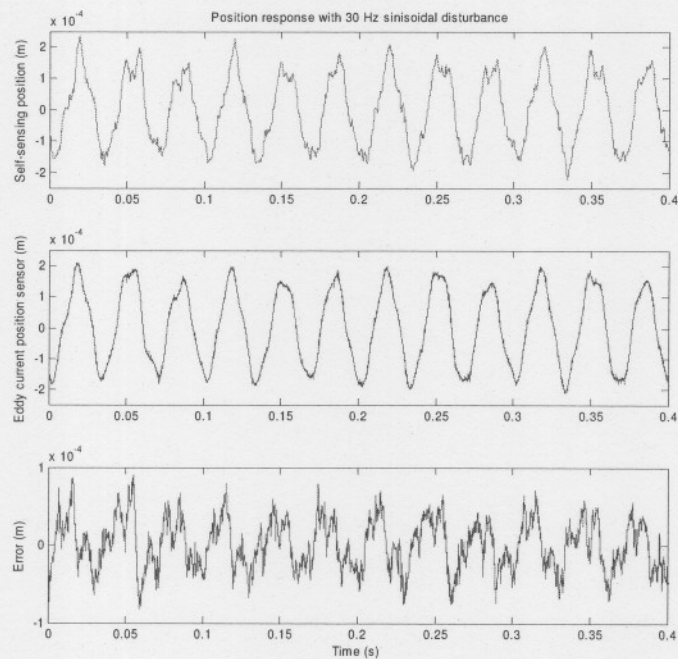


Figure 7.13: Sensor output with a 30 Hz sinusoidal disturbance

Figure 7.14 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 40 Hz .

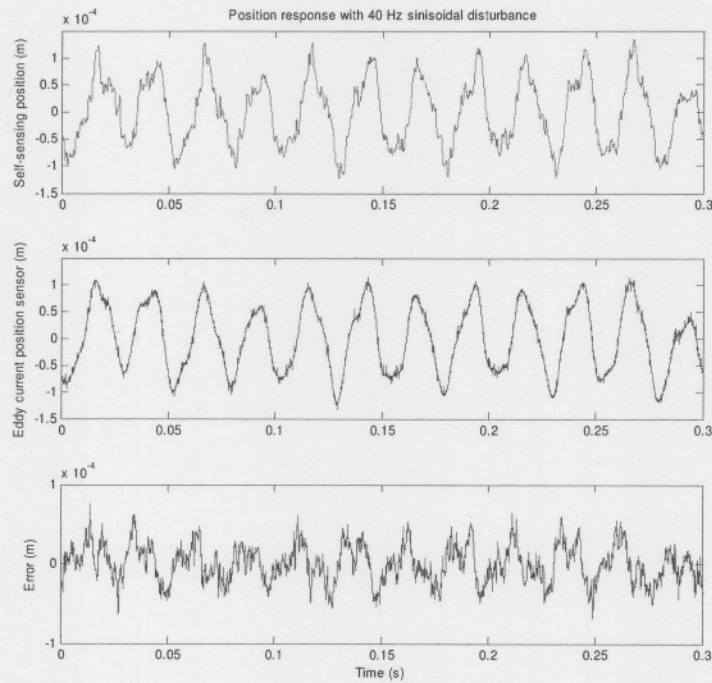


Figure 7.14: Sensor output with a 40 Hz sinusoidal disturbance

Figure 7.15 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 50 Hz .

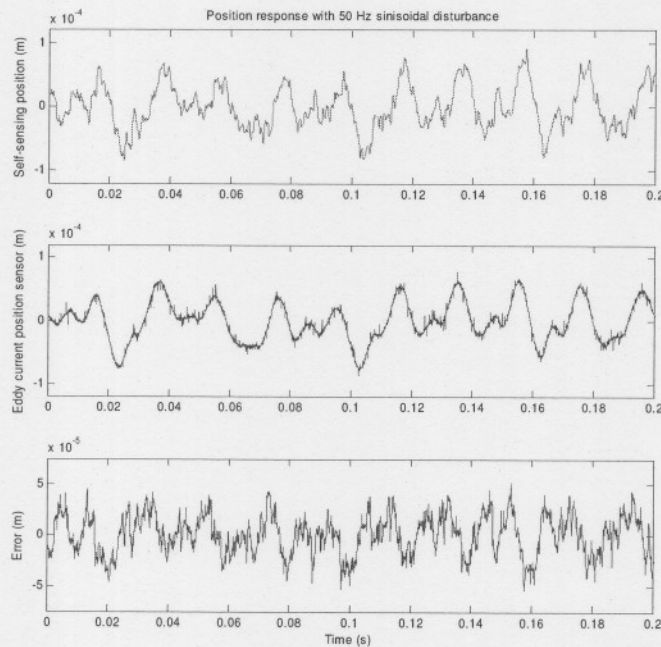


Figure 7.15: Sensor output with a 50 Hz sinusoidal disturbance

Figure 7.15 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 60 Hz .

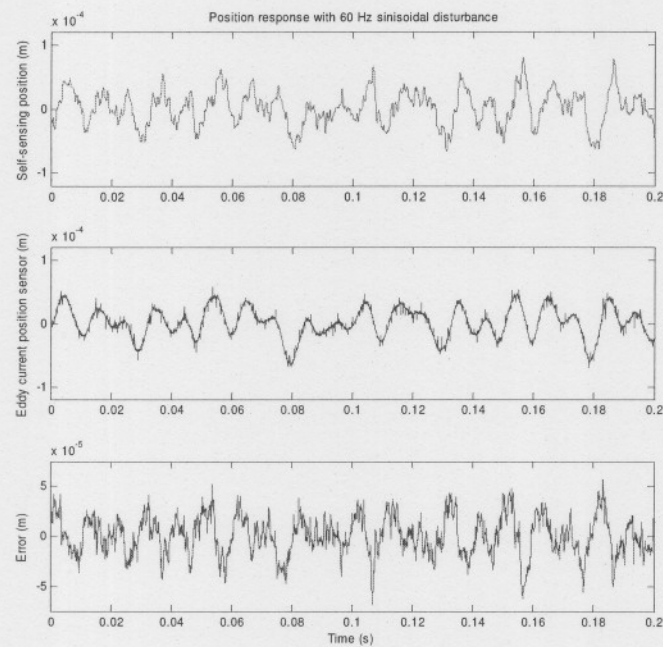


Figure 7.16: Sensor output with a 60 Hz sinusoidal disturbance

Figure 7.17 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 70 Hz . From these results it is determined that the sensor has a 1.5 ms delay at 70 Hz .

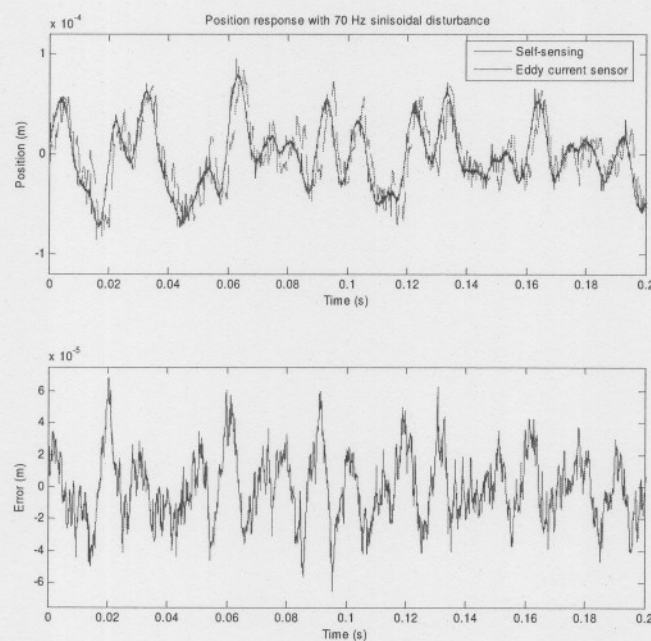


Figure 7.17: Sensor output with a 70 Hz sinusoidal disturbance

Figure 7.18 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 80 Hz .

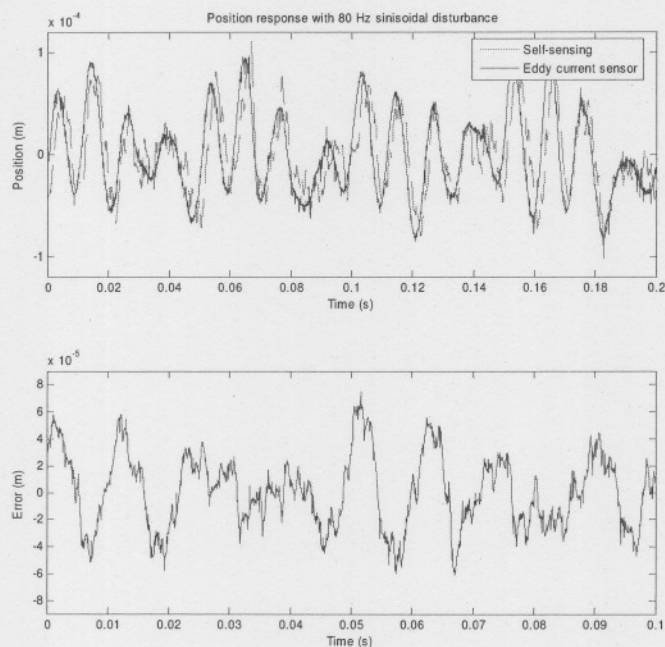


Figure 7.18: Sensor output with a 80 Hz sinusoidal disturbance

Figure 7.19 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 90 Hz .

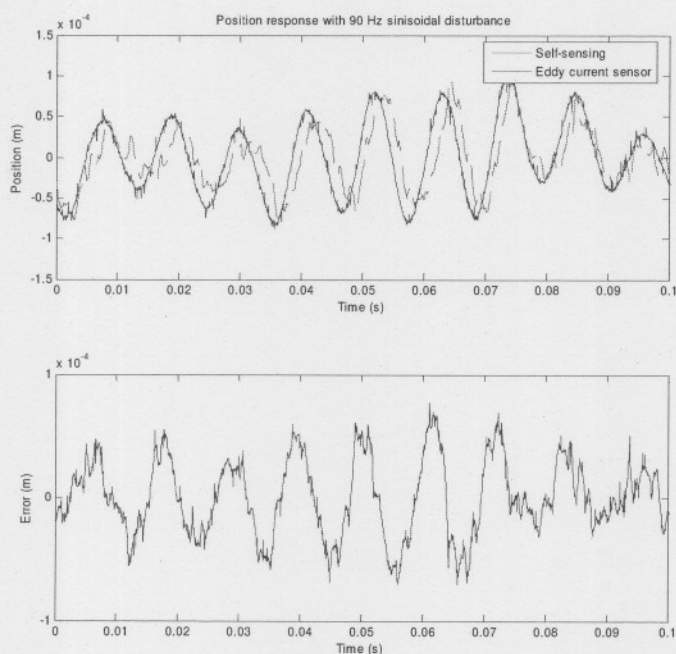


Figure 7.19: Sensor output with a 90 Hz sinusoidal disturbance

Figure 7.20 shows a sinusoidal disturbance of $30\text{ }\mu\text{m}$ at 100 Hz . The delay is 1.5 ms at 100 Hz .

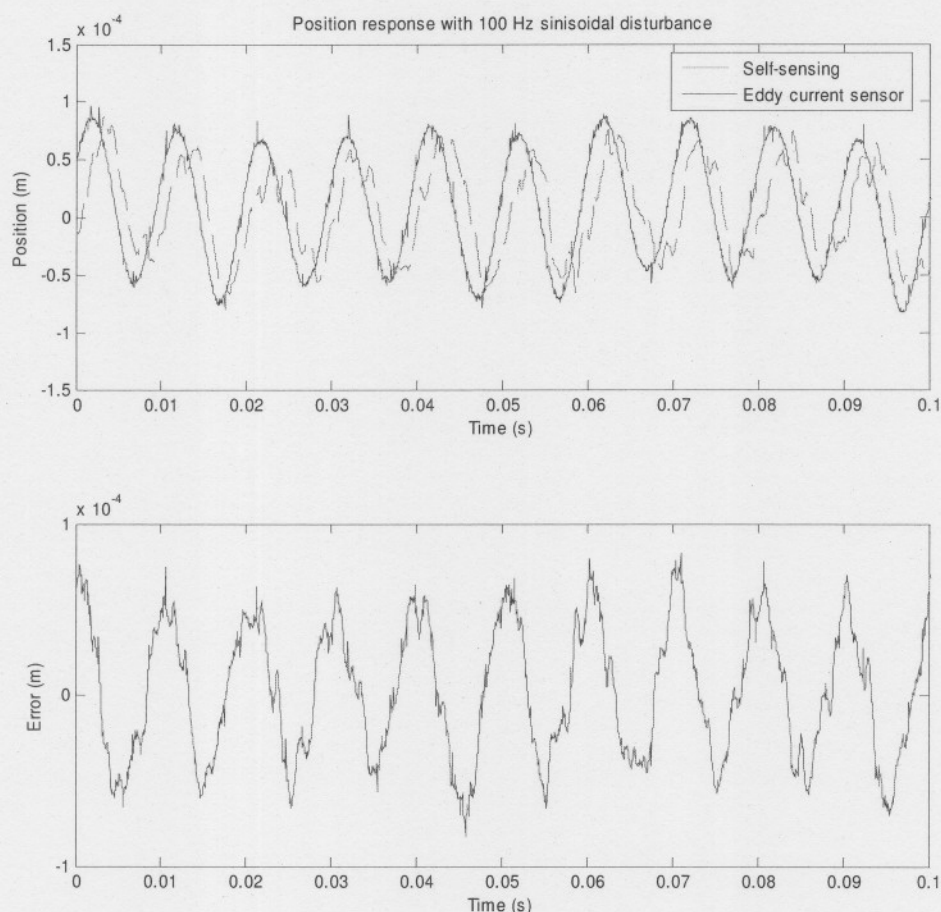


Figure 7.20: Sensor output with a 100 Hz sinusoidal disturbance

From the results of Figure 7.17 and Figure 7.20 it is determined that the sensor has a delay of 1.5 ms . This delay is not a phase shift, because the delay does not change over frequency.

7.6.3 Ramp response

Figure 7.21 shows when a sawtooth is given to the system.

From the dynamic evaluation the system was stable over a wide range of disturbances. This means that the sensor's bandwidth is high enough to accommodate high frequency disturbances. In the next section the effect of the noise and the nonlinearities will be discussed.

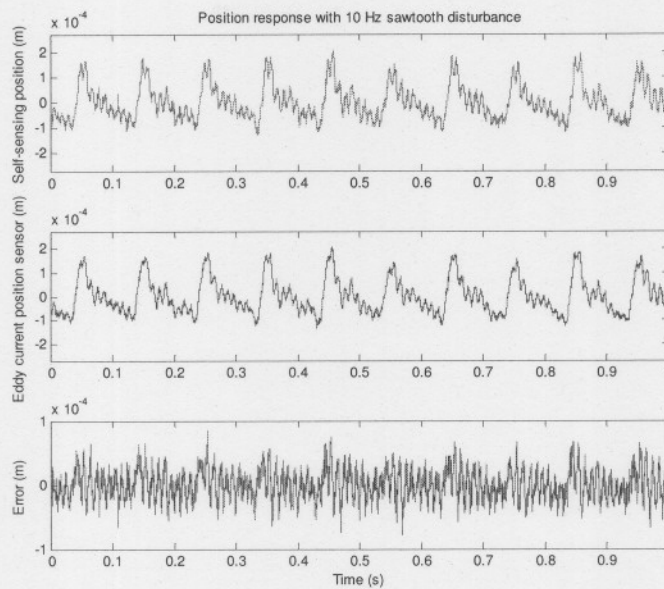


Figure 7.21: Sensor output with a 10 Hz sawtooth disturbance

7.6.4 Bandwidth

The bandwidth of the sensor was estimated by taking the fast Fourier transforms (FFTs) of the self-sensing sensor output and the eddy current sensor output. The FFTs of the position signals were taken from 5 Hz to 100 Hz in increments of 5 Hz. The FFT outputs are then converted to magnitude in dB with the eddy current sensor output as reference. From the result in Figure 7.22 it is determined that the bandwidth of the sensor is relatively low, in the order of 60 Hz. This method is however not a true measure of bandwidth. It nearly provides an estimation response relative to the eddy current probe.

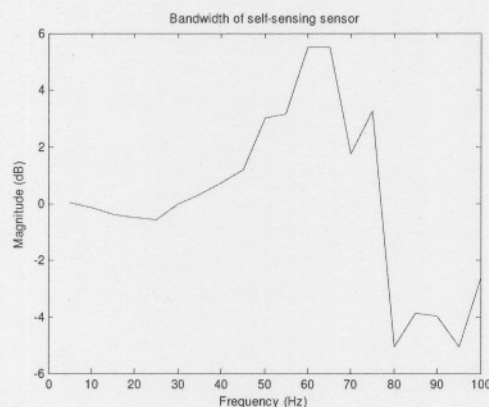


Figure 7.22: Bandwidth of self-sensing sensor

7.7 Nonlinearities

From Figure 7.1 it was determined that the current signal is not a pure triangular wave. The current wave could be viewed as a combination of a square wave and a triangular wave. This square wave is due to hysteresis loss in the rotor material. The square wave amplitude is inversely proportional to the air gap while the triangular wave is directly proportional to air gap. This inverse proportionality leads to a decrease in the amplitude of the square wave while the amplitude of the triangular wave increases when the air gap is increased. This interaction between the square wave and triangular wave contributes to an error in the self-sensing method.

Figure 7.23 shows the output of the forward path filter with a simulated current input and a current input with the nonlinearities discussed above. The air gap is varied with a sine wave. The square wave will be simulated to be inversely proportional to the air gap. This simulation only illustrates the effect of the inverse proportionality and is not a real representation of the current through the AMB. The dashed line shows the output of the simulated FPF without losses and the solid line shows the output with the simulated losses. It is determined that the output is largely affected by the losses.

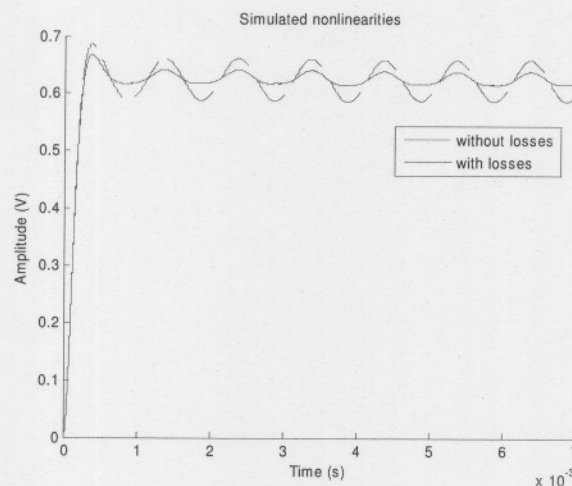


Figure 7.23: FPF output with losses

Figure 7.24 shows a Simulink® model of the self-sensing method. This model simulates the effect on the sensor if a 100 Hz ripple component is superimposed on the current. The current was practically measured and a 4% ripple component was found. This 4% ripple is simulated by the model.

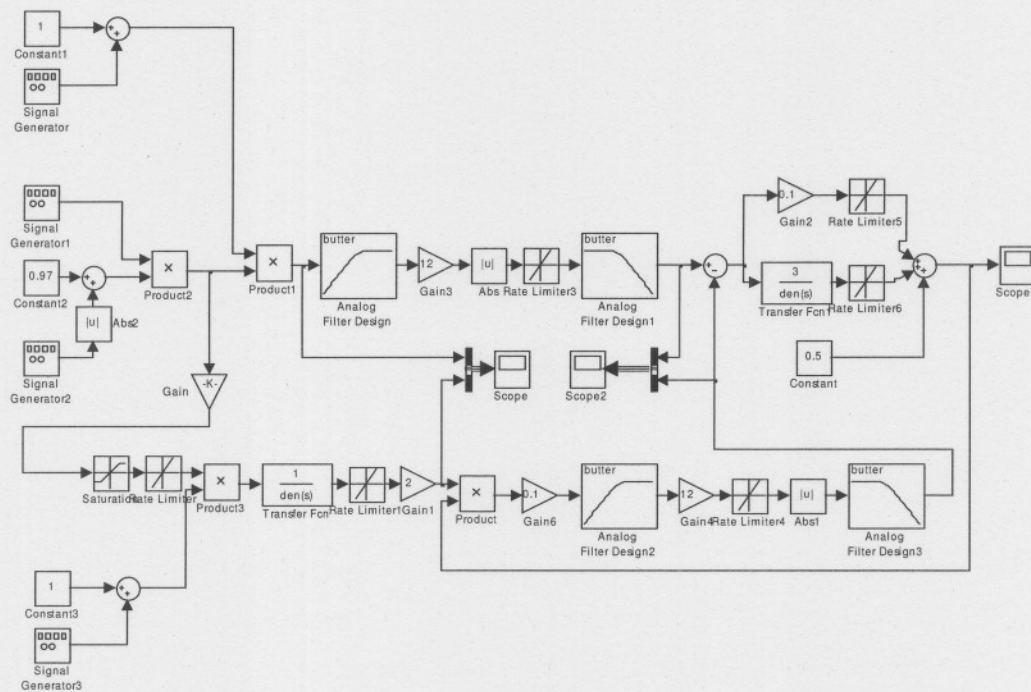


Figure 7.24: Simulink® model

Figure 7.25 shows the effect of the 4% ripple component on the sensor output. These results show that a small ripple component has a large effect on the sensor output.

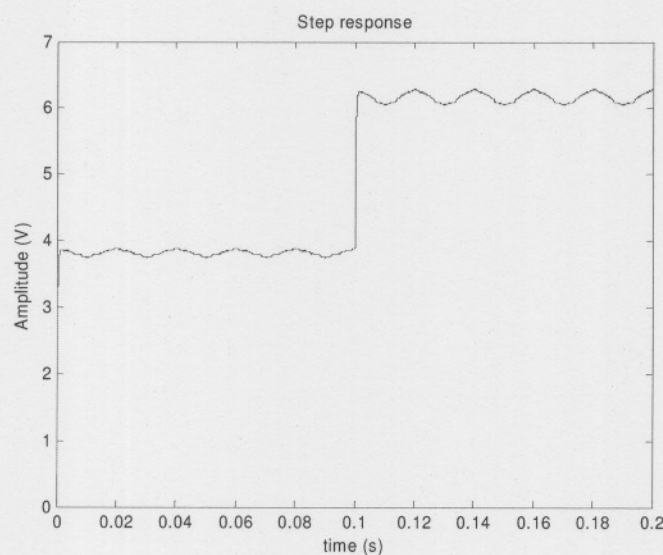


Figure 7.25: Simulink® model output

Figure 7.26 shows the ripple component on the power amplifier voltage. These results are practically measured. The H-bridge output voltage varies between 5 and 7 volts from

its operating voltage. If this ripple component (7 volt) is included into the simulation, then the output of the sensor is as shown in Figure 7.27. The estimated position is used to close the control loop in the simulation. The Eddy current sensor also follows the noise, as in section 7.6, since the self-sensing position is used to control the system. This is due to the effect of the PID controller. The AMB reacts as if the position changes with the noise.

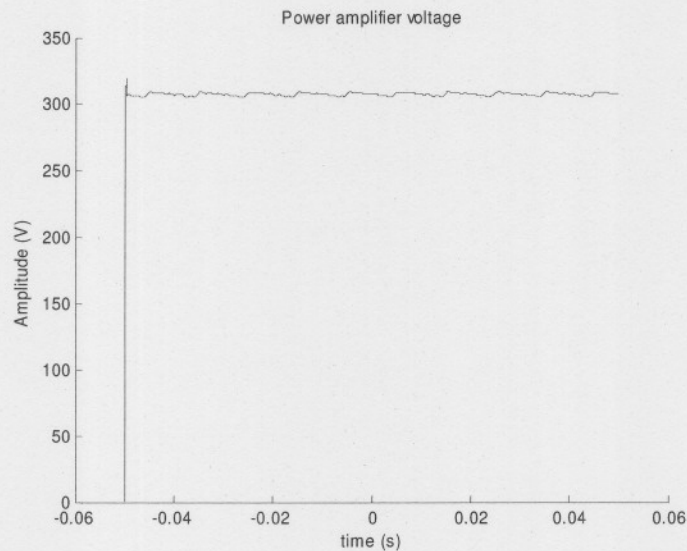


Figure 7.26: Power amplifier voltage

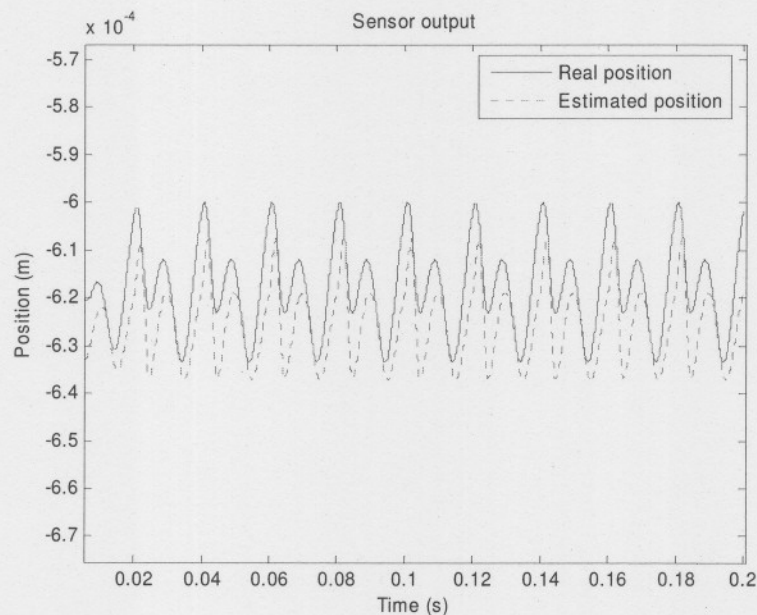


Figure 7.27: Self-sensing output with ripple

Figure 7.28 shows the simulation results of the sensor with an ideal amplifier used where the voltage is constant and no unbalance is present. The amplitudes of the ripple components of the ideal and non-ideal amplifiers, as shown in Figure 7.27 and Figure 7.28, differ by a factor of fifteen.

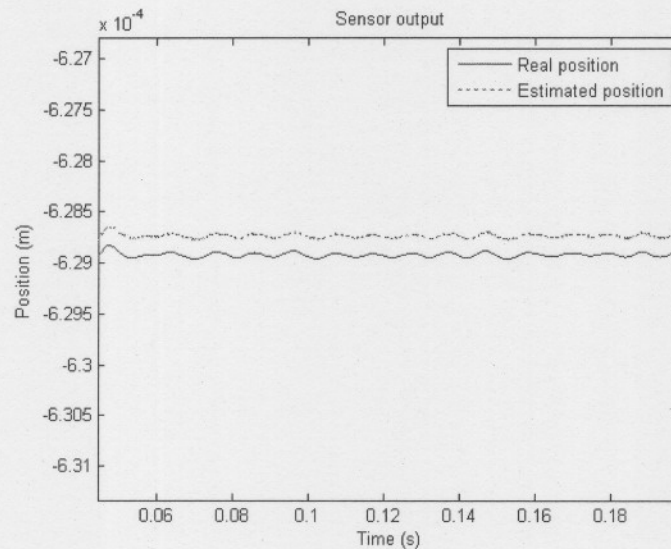


Figure 7.28: Self-sensing output without ripple

7.8 Conclusion

It is clear from the static tests and analysis that noise is present on the output of the self-sensing sensor. The noise increased when all the power amplifiers were switched on. This is due to the cross coupling of the magnetic fields. The noise also increases as the power amplifier delivers more current.

The self-sensing sensor performs well with the step response evaluation. The main concern is the noise. The power amplifiers are responsible for most of the noise, thus the only method to reduce the noise is to eliminate the ripple component on the power amplifier output.

From the sinusoidal response it was found that the sensor has a constant delay of 1.5 ms. This delay must be decreased to improve the bandwidth of the sensor. The sensor was able to defeat disturbances up to 100 Hz.

The linearity of the sensor is impaired by eddy current losses in the AMB. The eddy current losses could be minimized by using thinner laminations in the AMB magnetic circuit.

Conclusion and Recommendations

The aim of this project was to analyze and develop different types of position sensors for AMBs. The knowledge gained by developing the position sensors will support future research and development on AMBs. An overview on the developed sensors will be presented in the following order: 1) optical sensor 2) inductive sensor 3) self-sensing sensor.

8.1 Optical sensor

The barrier topology optical sensor was developed. The effect of the measurement distance was fully investigated and the linearity response was presented. From the results it was found that the infrared photo transistor and LED positioning plays an important part in the linearity of the optical sensor. The parallel layout worked the best to maximise the measurement distance and linearity. The characteristics of the infrared phototransistor have a large effect on the output, thus the type of phototransistor used is important. The larger the diameter of the infrared phototransistors and LEDs, the larger the linear range. Large offset effects were found when external infrared sources were inserted into the model. This offset problem was reduced by placing a phototransistor near the sensing phototransistor, but not in the infrared path of the LEDs. The external infrared source is then subtracted from the sensed signal. The sensor performed well when used with the magnetic bearing demonstrator model.

8.2 Inductive sensor

The inductive sensor relies on the output of the two series sensor coils. This output of the series coils are a typical amplitude modulated (AM) output. The AM output is then demodulated to position by an envelope filter (band-pass filter, full wave rectifier and an ac to dc converter). The delay of the envelope filter has large effects on the output at high frequencies. Another disadvantage of the filter is the ripple present on the output. The ripple is directly proportional to the bandwidth. The sensor has excellent linearity as shown in Figure 4.14.

The bandwidth was determined to be up to 1.5 kHz, depending on the allowable ripple component. The sensor was implemented on a homopolar radial active magnetic bearing. The sensor performed well under these test conditions.

8.3 Self-sensing sensor

The self-sensing method makes use of the PWM voltage of the power amplifier as well as the current through the coil. The sensor is simulated using a Matlab® model. The results from the simulation are found to be promising.

The sensor was implemented with analog electronics. All the characteristics used to implement the model were taken from the simulation. The main building blocks performed well during individual testing. All these building blocks were integrated to form the self-sensing sensor.

The self-sensing sensor was evaluated by integrating it with a heteropolar radial active magnetic bearing. This AMB makes use of four power amplifiers for each stator. The power amplifiers operate at 100 kHz switching frequency and ± 310 V output with a maximum current of 10 A.

The bottom coil's power amplifier was used for estimation and the top coil's power amplifier was used for control. The bottom amplifier was given a small reference voltage to limit the current through the coil. The self-sensing sensor was capable to suspend the rotor when the control loop was closed with the estimated position. From the experimental results it was determined that the self-sensing sensor performed relatively well under step response tests. The self-sensing sensor also performed well with high frequency sinusoidal disturbances, up to 100 Hz. It was determined that the sensor has a delay of 1.5 ms. The output of the sensor was found to be noisy under static and dynamic test conditions. This noise component was determined to be 100 Hz due to the power amplifier charge cycle. Nonlinearities are also of concern. The most important nonlinearity of concern is the laminated rotor. The hysteresis losses are relatively large due to the thickness of the laminated material used. These hysteresis losses deform the current component, adding a square wave component to the triangular current. This square wave component is inversely proportional to the air gap. This added component on the current induced an error on the sensor output.

8.4 Recommendations

- The optical sensor range may be increased if the current through the infrared LEDs is increased. The LEDs have a maximum rated current. If this increased current exceeds the LED rated current, the current through the LED must be pulsed. This pulsed signal will also decrease the susceptibility to external infrared sources. The reason for this is that the noise can be filtered out.
- In the case of the inductive sensor, the shape of the core of the sensor coils may be investigated. The cores currently used are halves of ferrite cores. These halves are not machined to fit uniformly over the rotor. If the core is machined the fringing of the flux will be reduced and the sensor will be more linear. The effect of the permeability of the core may also be investigated. The injected oscillation frequency must be investigated to determine the skin effect on the sensed material. Different types of ferromagnetic materials may be investigated to determine the effect on the linearity of the sensor.
- In the case of the self-sensing sensor, the hysteresis losses must be minimized. This is done by using thinner laminations in the AMB magnetic circuit. This will reduce the losses and increase the linearity of the self-sensing sensor.

The 100 Hz ripple component of the power amplifier must also be reduced. This is done by redesigning the power amplifier with allowable specifications. If the ripple of the power amplifier is reduced the noise in the self-sensing sensor output will be drastically reduced.

The estimation AMB model of the self-sensing sensor must be changed to simulate the AMB more closely. This more accurate model of the AMB will reduce the inconsistencies between the estimated current and real current, and will increase the accuracy of the sensor.

Shielding is an essential part of the sensor's design. Proper shielding will reduce susceptibility to high frequency noise.

The current sensors available on the market have relatively low bandwidth and are expensive. The current sensor developed for the self-sensing sensor must be investigated to understand the effects of different core materials and the effect of the air gap. The effect of external magnetic fields on the current sensor must be investigated.

By implementing a differential sensor where the air gaps on both sides of the rotor are estimated and subtracted a more sensitive and noise immune system could be obtained. Implementation by means of a digital signal processing (DSP) device is also an important future activity.

APPENDIX

Data on CD

A Datasheets

B Matlab® simulations

B.1 M-files

B.1.1 Current nonlinearities

B.1.2 FPF evaluation

B.1.3 High-pass filter transfer function

B.1.4 Low-pass filter transfer function

B.1.5 Optical sensor results

B.1.6 Self-sensing bandwidth estimation

B.1.7 Self-sensing model with changing air gap with no feedback

B.1.8 Self-sensing simulation

B.1.9 Self-sensing without estimator feedback

B.2 Simulink® model

B.2.1 Linear time invariant system

B.2.2 Self-sensing Simulink® model

C Orcad® simulations

C.1 4 Diode rectifier

C.2 Ac to dc converter

C.2.1 ac to dc converter

C.2.2 Low-pas implementation

C.3 Band-pass filter

C.4 Inductive sensor

C.5 Integrator

C.6 Optical sensor

C.7 Self-sensing schematics

C.8 UAF42

C.8.1 High pass filter

C.8.2 Low-pas filter

D Documentation

D.1 Dissertation (Microsoft® Word document)

D.2 References

E Measurements

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