Back-EMF Based Sensorless BLDC Control of a High Speed Permanent Magnet Machine

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This master dissertation is written to complete the Master of Science in Electromechanical Engineering, at the University of Ghent. I chose this dissertation subject ‘control of high-speed electrical machines’ because I believe that high-speed electrical machines will play an important role in making many systems more energy efficient and will open ways to completely new applications.

This dissertation aims to provide the reader both theoretical and practical insight in sensorless control of high-speed permanent magnet synchronous machines. It is hoped that this thesis serves as a basis whereupon future researchers can build to find better digital sensorless control algorithms.

First and foremost, I would like to thank my supervisors and counsellor, Prof. dr. ir. Peter Sergeant, Prof. dr. ir. Frederik De Belie and ing. Bert Hannon. They gave me valuable insight, clever suggestions and support when needed. This thesis would not be possible without Bert’s endless patience and his help with the practical aspects of the set-up.

I would like to thank Simon Wauters who worked with me on the same test set-up and helped solve many problems.

Lastly, I would like to thank Liselotte for the help with the linguistic aspects of this dissertation and the loving support throughout the year.
Abstract

High-speed electrical motors can play an important role in making motor driven systems more efficient. There are two most commonly perceived advantages of using high-speed electrical motor. Firstly, the reduction of system weight and size for a given magnitude of power conversion. This is particularly desirable in mobile applications, where any savings in weight directly result in reduced fuel burn and emissions. Secondly, adopting high-speed machines in certain applications greatly improves efficiency and reliability as a result of the elimination of intermediate gearing.

The electrical machine of choice for low-power, high-speed applications is the permanent magnet synchronous machine (PMSM). The PMSM has the advantage of having very high efficiency, power density and good dynamic behaviour. The main disadvantage is that the use of permanent magnets can significantly increase the capital cost. In order to control a PMSM optimally, accurate knowledge of the rotor position is necessary. The electrical peculiarities of such high speed machines can cause a series of problems with the stability of the control, if the position and speed of the rotor is not known with high precision. To improve reliability and reduce costs, a recent trend has emerged to drive machines sensorless. Sensorless or ‘indirect position sensing’ is a denominator for a collection of techniques that estimate the rotor position indirectly, without the use of mechanical sensors but with voltage and current measurements.

The control in this work is completely digital and implemented on a field-programmable gate array (FPGA). FPGAs have several advantages over traditional digital signal processors, especially in high speed applications where time constraints on the signal processing are high.

This work presents the research done towards fully-digital sensorless BLDC control of a high-speed PMSM (30 000 rpm, 3kW) on a test set-up developed at Ghent University. The sensorless control is based on the line-to-line back-EMF as first introduced by Kim et al. [2011b] in 2011. The research comprised of two main parts: basic development of the sensorless algorithm on a low speed machine and extended development on the high speed machine developed at Ghent University. Two variants of the digital sensorless control, one using merely the back-EMF and the other also using current measurement, were experimentally validated on the set-up from 500 rpm (1.7% of nominal speed) up to 10000 rpm. The algorithms show promise to work over the full speed range, but research can still be done to make the algorithm completely fail-safe.
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Academic year 2016-2017

Abstract—This paper presents the research done towards fully-digital sensorless control of a high-speed PMSM (30 000 rpm, 3kW) on a test set-up developed at Ghent University. The sensorless control is based on the detection of the zero-crossings of the line-to-line motional back-EMF and is implemented on an FPGA. A first successful implementation was done on a low-speed (3000 rpm, 5kW) PMSM with a high number of poles as a proof of concept. The research was then continued on the high-speed machine, adapting the sensorless control to the electrical peculiarities of high-speed machines. The main challenge was dealing with the long free-wheeling diode conduction period, an inherent property of high-speed machines. Two variants of the digital sensorless control were experimentally validated from 500 rpm (1.7% of nominal speed) up to 10 000 rpm and show promise to work well over the full speed range, although requiring some tuning.

Keywords—Sensorless, self-sensing, high-speed, PMSM, FPGA, line-to-line back-EMF, free-wheeling diode

I. INTRODUCTION

High-speed electrical motors can play an important role in making motor driven systems more efficient. There are two most commonly perceived advantages of using high-speed electrical motor. Firstly, the reduction of system weight and size for a given magnitude of power conversion. This is particularly desirable in mobile applications, where any savings in weight directly result in reduced fuel burn and emissions. Secondly, adopting high-speed machines in certain applications greatly improves efficiency and reliability as a result of the elimination of intermediate gearing [1].

The electrical machine of choice for low-power, high-speed applications is the permanent magnet synchronous machine (PMSM) [2]. The PMSM has the advantage of having very high efficiency, power density and good dynamic behaviour. The main disadvantage is that the use of permanent magnets can significantly increase the capital cost [3] [4].

In order to control a PMSM optimally, accurate knowledge of the rotor position is necessary. The electrical peculiarities of such high speed machines can cause a series of problems with the stability of the control, if the position and speed of the rotor is not known with high precision [5].

To improve reliability and reduce costs, a recent trend has emerged to drive machines sensorless. Sensorless or ‘indirect position sensing’ is a denominator for a collection of techniques that estimate the rotor position indirectly, without the use of mechanical sensors but with voltage and current measurements.

The control in this work is completely digital and implemented on a field-programmable gate array (FPGA). FPGAs have several advantages over traditional digital signal processors, especially in high speed applications where time constraints on the signal processing are high. It is suggested that FPGAs will dominate sensor and filtering applications [6]. FPGAs allow rapid prototyping of applications [7] and have become increasingly cheap [8], allowing their entry in more and more applications. Although some sensorless controls were implemented on FPGAs [9] [10], little is written on the possibilities of using sensorless control of high-speed machines implemented on FPGAs.

This work presents the research done towards fully-digital sensorless BLDC control of a high-speed PMSM (30 000 rpm, 3kW) on a test set-up developed at Ghent University. The sensorless control is based on the line-to-line back-EMF as first introduced by [11] in 2011. The research comprised of two main parts: basic development of the sensorless algorithm on a low speed machine and extended development on the high speed machine developed at Ghent University. Two variants of the digital sensorless control, one using merely the back-EMF and the other also using current measurement, were experimentally validated. They were experimentally validated on the set-up from 500 rpm (1.7% of nominal speed) up to 10 000 rpm and show promise to work well over the full speed range, although requiring some tuning.

II. BLDC CONTROL

The very high fundamental frequency that high speed electrical machines need to be supplied with pose significant challenges for the power electronics supplying the machine. The authors of [12] conclude that the use of a voltage source inverter (VSI) with block commutation offers low switching losses, simple control and easy implementation of sensorless control. In literature, the chosen inverter topology and commutation strategy is also referred to as ‘variable dc link inverter’ or Pulse Amplitude Modulation (PAM) inverter [13]. The inverter part is controlled in six-step or block commutation, which means that each switch is conducting for 120 electrical degrees and, therefore, switched only with the fundamental frequency of the machine. The dc-dc converter modulates the amplitude of the current blocks. On the other hand, the more commonly used Pulse Width Modulation (PWM) changes the amplitude of the current blocks by switching the inverter at a much higher rate. [14] compared the classical PWM scheme with the variable DC link inverter for high speed control and came to two arguments that favour PAM over PWM. First, although PWM provides better control dynamics [13], the advantage of only having to switch
at the fundamental frequency makes the combination of a PAM inverter with a PMSM very attractive for high speed operation. At high speed, only a few PWM pulses can be used for the speed control during the on-time of the interval. Second, PWM may cause commutation delays or an irregular switching frequency that have a significant influence on the phase current and drive performance at high speed. Since the commutating instants depend on the rotor position, it does not usually coincide with the end of a PWM switching period. The commutation is normally performed synchronised with the end of the present PWM period to start the next inverter sequence 1(a). Even though this delay can be neglected in a normal speed range, it becomes problematic at high speed since the 120° intervals become relatively small. To avoid such an undesirable delay, the next inverter sequence has to be applied as soon as the commutation signal interrupt occurs. In some PWM schemes, this may yield an irregular switching frequency much larger than the switching frequency \( f \), under high-duty conditions as shown in Figure 1(b).

When a PMSM is driven by six-step block commutation, its behaviour is similar to that of a DC machine. It is then commonly referred to as a brushless DC (BLDC) machine. The six possible combinations in terms of energised phases are usually provided by Hall-effect sensors mounted in the stator. The Hall-effect sensors output a binary signal when either a north or south pole of the rotor magnets passes. By using three sensors, shifted over 120°, one obtains six possible combinations, precisely enough to uniquely define every commutation instant. Six-step commutation is best combined with PMSMs with a trapezoidal distribution of air-gap induction, such that the flat part of the trapezoidal back-EMF coincides with the current blocks, resulting in constant torque, as illustrated in Fig. 2.

![Fig. 1: Relation between PWM switching period and commutating instant. (a) Commutation delay (b) Irregular frequency](image)

III. NON-IDEAL COMMUTATION BEHAVIOUR

Note that at any given angle in Fig. 2, one of the phases has zero current. The assumption that current immediately stops flowing in the non-energised phase after opening one of its switches is an idealisation that does not hold in reality. Due to the inductive behaviour of the stator windings, current has the tendency to keep flowing. To provide a path for the remaining current in each phase after commutation, free-wheeling diodes over the power switches provide an alternative conduction path. As a consequence, during the time that a free-wheeling diode is conducting, all three phases are conducting current instead of the theoretical two. The phase current waveform thus deviates from the ideal block wave and is more trapezoidal, as shown in Figure 4 [15]. The diode conduction period is influenced by the winding resistance, the back-emf, the winding inductance, the load, and in particular by the ratio \( \frac{\omega(L-M)}{R} \), where \( \omega \) is the electrical pulsation, and \( R, L \) and \( M \) are the winding resistance and self- and mutual-inductances per phase, respectively [15]. We can already establish that for high speed machines the free-wheeling period will be quite long, since the electric pulsation is relatively high, to the point that the phase currents flow more or less continuously.

IV. SENSORLESS CONTROL ALGORITHMS

A. Overview

A number of different methods exist to estimate the position of the rotor. The methods can be devided into three main...
High-Speed Motor

<table>
<thead>
<tr>
<th>Quantity</th>
<th>Value</th>
<th>High-Speed Motor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nom. Voltage</td>
<td>60 V</td>
<td>Nom. Voltage</td>
</tr>
<tr>
<td>Nom. Speed</td>
<td>3000 rpm</td>
<td>Nom. Speed</td>
</tr>
<tr>
<td>Nom. Power</td>
<td>4.5 kW</td>
<td>Nom. Power</td>
</tr>
<tr>
<td>Nom. Current</td>
<td>80 A</td>
<td>Nom. Current</td>
</tr>
<tr>
<td>Pole Pairs</td>
<td>7</td>
<td>Pole Pairs</td>
</tr>
<tr>
<td>Nom. Freq.</td>
<td>350 Hz</td>
<td>Nom. Freq.</td>
</tr>
</tbody>
</table>

Table I: Nameplates of Torcman TM685-40 motor and custom high speed motor.

groups: Methods based on the motional back-EMF, observer-based methods, methods based on inductance variation and methods based on flux linkage variation. A description of the working principles of the state-of-the-art in sensorless methods can be found in [16]. Most high speed PMSMs use surface-mount permanent magnets because of the simple geometric form and construction of the rotor. This excludes the use of sensorless methods that exploit the difference in saliency, such as Inductance Variation, since the difference in inductance is negligible. Furthermore, because of the high speeds at which the machine at hand operates, complex algorithms such as Flux Linkage Variation or Observer-Based Methods might take too long to calculate on-line, especially for high speed machines where time frames are even smaller. Observer-based methods require accurate machine models. Constructing models and finding the right parameters for these models can be very time consuming. The methods using the Motional Back-EMF seem the most simple and robust, and without an upper bound concerning speed.

B. Motional back-EMF

Spatial movement of the flux vector from the permanent magnets on the rotor induces a motional back-EMF in the stator windings. Since the instantaneous magnitude of this EMF is proportional to the relative position of the permanent magnets, the EMF contains information about the position. In an ideally commutating BLDC, the phase voltage over the unexcited phase is equal to the back-EMF. The zero-crossings of this signal are very useful for determining the commutation, because their relative position to the commutation instants is fixed and they occur precisely at rotor positions where the one of the phase winding is not excited [16]. The zero-crossings lead the commutation instants by exactly 30°, such that a phase shift is required to obtain the instants.

C. Line-to-line back-EMF

In 2011, [11] presented a method that improves signal-to-noise ratio drastically while also eliminating the need for the classical 30° phase shift. Unlike the zero-crossing of the open phase, the zero-crossing of the line-to-line back-EMF difference corresponds exactly to the commutation instant. Indeed, the line-to-line voltages always lag with respect to the phase voltages, as one can observe in Figure 2. Furthermore, this method eliminates the need for an accessible star-point of the machine (or a virtual one) that is required to measure the phase voltage. Note that the the coincidence of zero-crossings and commutation instants means that the zero-crossings of the line-to-line back-EMFs can be used to make simulated Hall-effect sensors. In Figure 2 we see that when line-to-line voltage $EB - EC$ is positive, Hall-effect sensor $H1$ is low. Thus, the output signals of the method can be directly applied to the conventional commutation table, as if they were obtained from the real Hall-effect sensors.

D. Influence of non-ideal commutation

To detect zero-crossings, the phase current waveform has to be discontinuous, with no current flowing for a sufficiently long interval near zero-crossing of the back-EMF [15]. However, in Section III we explained that during commutation, the freewheeling diodes conduct for some time due to the inductive behaviour of the windings. If the conduction period becomes longer than 30° the zero-crossings can no longer be detected. This demands an appropriately designed motor where not only efficiency and flux density are design parameters, but also the inductances of the windings and by consequence the conduction angle of the diodes. In [15] a complete design methodology for a high speed, 120 000 rpm motor is given. The authors show that whilst motors that have widely varying design parameters exhibit more or less the same performance, the diode conduction angle can vary by a factor of 2 or more. In Figure 5 the back-EMFs of two motors with similar efficiency are given.

V. DESCRIPTION OF THE TEST SET-UP

The sensorless algorithm based on the line-to-line back-EMF was implemented on an experimental test set-up developed at the University of Ghent. Test were conducted on a low-speed and high-speed machine, with the nameplates from Table I. Both motors have externally mounted Hall-sensors. Line-to-line voltage measurements are made using simple voltage dividers between the motor terminals. The current of two phases and current in the variable DC-link are measured with modules from LEM current transducers. These signals are fed into a National Instruments (NI) compactRIO controller. The controller features a real-time operating system, a reconfigurable FPGA and an ethernet port. The controller is equipped with NI 9223 C Series analog input measurement modules, that can sample measurements at a rate of one million samples per second. The FPGA is programmed using NI LabVIEW on a host PC connected over ethernet.
VI. DEVELOPMENT ON A LOW-SPEED MACHINE

As a proof of concept, a digital sensorless algorithm was developed on the Torcman motor, a small BLDC motor with a high number of pole pairs. Due to this high number of poles, the nominal el. frequency is only 3 times smaller than the nominal frequency of the high-speed machine.

A. Requirements for a ZCD Algorithm

After some initial attempts to detect zero-crossings and to reject unwanted noise and commutation spikes, the key requirements for a zero-crossing detection (ZCD) algorithm were identified:

- Consistent and controllable timing
- Measuring only in specific intervals to reject peaks and noise
- A 'band' around zero that triggers the ZCD

B. Sensorless commutation using detection intervals

It was found that the commutation spikes can be easily rejected if measurements are only made in an interval where the line-to-line voltage is close to zero for an extended time. This allows to optimise the ZCD within this interval, without having to consider what effect the optimisations would have on detecting ‘false’ crossings outside this interval. The interval is constructed using a debounce filter, that processes the raw sampled data and only pass through a change in state whenever the sampled data has remained constant for a defined period of time. If the time constant of the debounce filter is set correctly, it will reject the short instants where the line-to-line voltage is within the interval due to the commutation spikes, but will accept the periods where the line-to-line voltage is within the interval for prolonged periods. Figure 6 illustrates this, with (a) being the line-to-line back-EMF with commutation spikes, (b) is a boolean value that is ‘high’ where (a) is within the indicated interval. A debounce filter is then applied with the time constant shown in red, resulting in the debounced signal (c). The detection intervals allow to easily detect single zero-crossings. However, whether a crossing goes from negative to positive or positive to negative has to be determined to uniquely define the sequence of commutation instants. This is done by sampling the back-EMF once at the beginning of the detection interval. This algorithm successfully generated virtual Hall-effect sensor signals that were used to drive the Torcman motor sensorless over the full speed range.

VII. EXTENDED DEVELOPMENT ON A HIGH-SPEED MACHINE

A. Long diode conduction period

Figure 7 shows the waveform of the line-to-line back-EMF of the high-speed motor developed at Ghent University. We see that the diode conduction period (DCP) is indeed problematic for a ZCD algorithm. Already under low load conditions the DCP can be as long as the non-conducting interval, making the use of solely a debounce filter insufficient. At partial load conditions the DCP quickly becomes longer than 30° el., completely inhibiting the use of back-EMF based methods.

B. Shortening the diode conduction period

In order to extend the range where the machine can be operated sensorless, it is desired to shorten the DCP. Two possibilities were tested in simulation: first, synchronising the dc-link voltage with the commutation instants. Second, advancing or delaying the commutation instants. The first possibility is based on the idea that during the DCP, the non-energised phase is temporarily connected to the positive DC-bus (see Figure ??). The voltage over the phase (red arrow) then counters the current (blue arrow) and forces it to exponentially diminish to zero. However, whether a crossing goes from negative to positive or positive to negative has to be determined to uniquely define the sequence of commutation instants. This is done by sampling the back-EMF once at the beginning of the detection interval. This algorithm successfully generated virtual Hall-effect sensor signals that were used to drive the Torcman motor sensorless over the full speed range.

Fig. 5: Simulated back-EMFs of a motor with low winding inductance (left) and high winding induction (right).

![Fig. 6: Illustration of the debounce filter applied to a threshold on the line-to-line voltages.](image-url)
Fig. 7: Line-to-line voltage and phase current of the high-speed machine under different load conditions.

<table>
<thead>
<tr>
<th>Commutation Delay (° el.)</th>
<th>Diode Conduction Angle (° el.)</th>
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<tbody>
<tr>
<td>-20</td>
<td>18.0</td>
</tr>
<tr>
<td>-12</td>
<td>17.28</td>
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<tr>
<td>-6</td>
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<td>10</td>
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</tr>
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<td>20</td>
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Table II: Commutation delay and corresponding diode conduction angle.

uation was delayed and advanced while measuring the DCP. The results, displayed in Table II show that although delaying the commutation instants significantly lengthens the DCP, advancing the commutation instants shows no real improvement.

C. Detection interval adaptation

To deal with long DCPs, however not longer than 30° el., two different variations on the detection interval method based on the debounce filter are implemented.

C.1 Delayed detection interval

The first method uses a detection interval that in first instance includes the DCP. A new interval is then created of which the beginning is delayed and the ending coincides with the ending of sensing interval, as illustrated in Figure 8. The delay should correspond to a fixed number of electrical degrees, and thus depend on the speed of the machine. It is then of course necessary to prevent that the conduction period becomes longer than the fixed delay, otherwise the algorithm will fail to correctly detect the zero-crossing. This can be done using current measurements: if the current is zero for a shorter time than 30° minus the DCP (in degrees), the speed or load should be lowered.

C.2 Using current measurements

If accurate and fast current measurements are available, as is the case on the test set-up, a simpler algorithm can be implemented that requires less tuning. If we use the interval where the current is near zero in the phase that is non-energised as detection interval, the diode conduction period is completely ignored. We can again use a debounce filter to make sure this interval is not intermitted by measurement errors. The advantage of this method is it solidly rejects the diode conduction period with very little tuning. If the diode conduction period changes due to load or speed changes, the detection interval will adapt smoothly with it. However, if the algorithm fails to detect a zero-crossing, there is little possibility for recovery, since the current waveform is a consequence of the commutation instants. If commutation doesn’t happen because of a missed zero-crossing, the current waveform will considerably deviate from the one during normal operation. As a consequence, the next detection interval will be wrong or even be absent, such that no more zero crossings can be detected. The motor will then stall. This could be solved by forcing a timed commutation at for example 65° el. after the last commutation if no commutation has happened after 62° el.

D. Robust sign detection

During the experiments on the high speed machine it appeared that using a single measurement to determine the sign of the back-EMF was no longer accurate since the damped oscillations after the diode conduction period (caused by the dynamic behaviour of the free-wheeling diode) caused the algorithm to measure a wrong sign. This was solved using another debounce filter that keeps track of the sign of the back-EMF.

VIII. RESULTS

The complete implementation of the sensorless algorithm for the high-speed machine can be summarised as follows: The algorithm waits until it detects it has entered the detection interval, either by the method from paragraph VII-C.1 or paragraph VII-C.2. A zero-crossing signal is then indicates that the back-EMF is almost zero. Depending on the sign of the back-EMF, a virtual Hall-effect sensor is then set either ‘low’ or ‘high’.
The algorithm is then locked to prevent detection of rapid successive zero-crossings until the back-EMF is again outside the detection interval. Both the algorithm using the delay during the diode conduction and the algorithm using current measurements were tested on the test set-up. Due to technical problems on the set-up it was not possible to test the algorithms above 12000 rpm. Measurements were made at 500, 2000, 5000 and 10000 rpm showing that both variations on the algorithm worked correctly. Both methods for determining the detection interval needed some tuning for different speeds. The tuning follows a general trend: for higher speeds the time constants of the debounce filters have to be shorter. For low speeds the ZCD should use a low threshold value, such as 0.02 such that commutation instants are not advanced too much. For higher speeds, the threshold value should be a bit higher, for example 0.05 in order not to miss a zero-crossing. The amount of samples (time constant of debounce filter) that is used to determine the sign of the back-EMF has to be lowered for higher speeds, as well as the number of samples that is used for determining the detection interval. The real Hall-effect sensors are slightly delayed to make the difference between sensed and sensorless commutation more pronounced. In Figure 9 we see that switching between the delayed Hall-effect sensor signals and the accurate sensorless signals slightly shortens the diode conduction period, as we could expect from the simulation results.

IX. CONCLUSION

An digital algorithm based on the line-to-line back-EMF method from [11] was developed. The sensorless algorithm uses a debounce filter to establish 'detection intervals' where the back-EMF is measured for zero-crossings. The detection intervals are constructed in such way that they ignore the long diode conduction period that is inherent to high-speed machines. On variation creates a detection interval by using a delay during the DCP. Another variation measures when the phase current is zero and uses this information to construct the interval. Both variations were experimentally verified on a test set-up from 500 rpm until 10000 rpm, 1.7% and 33% of nominal speed respectively. Before testing over the full speed range, a safety should be build-in that forcibly commutates if a commutation is missed by the algorithm, to prevent sudden stalls of the motor. The algorithm itself shows promise to work over the full speed range.

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<tr>
<td>$d$</td>
<td>Along direct axis</td>
</tr>
<tr>
<td>$q$</td>
<td>Along quadrature axis</td>
</tr>
<tr>
<td>$s$</td>
<td>Referred to the stator</td>
</tr>
<tr>
<td>$f$</td>
<td>Rotor reference frame</td>
</tr>
<tr>
<td>$s$</td>
<td>Stator reference frame</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Electrical pulsation</td>
</tr>
<tr>
<td>$\Psi_p$</td>
<td>Permanent magnet flux vector</td>
</tr>
<tr>
<td>$DC^+$</td>
<td>Positive DC-bus voltage</td>
</tr>
<tr>
<td>$DC^-$</td>
<td>Negative DC-bus voltage</td>
</tr>
<tr>
<td>$E_p$</td>
<td>Back-EMF</td>
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<tr>
<td>$i$</td>
<td>Phase current</td>
</tr>
<tr>
<td>$L$</td>
<td>Inductance</td>
</tr>
<tr>
<td>$M$</td>
<td>Mutual inductance</td>
</tr>
<tr>
<td>$R$</td>
<td>Resistance</td>
</tr>
<tr>
<td>$t_{bl}$</td>
<td>Blanking time</td>
</tr>
<tr>
<td>$u$</td>
<td>Phase voltage</td>
</tr>
<tr>
<td>$V_{CE}$</td>
<td>Collector-Emitter Voltage</td>
</tr>
<tr>
<td>Back-EMF</td>
<td>Back-Electromotive Force</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>BLAC</td>
<td>Brushless AC</td>
</tr>
<tr>
<td>BLDC</td>
<td>Brushless DC</td>
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<tr>
<td>CSI</td>
<td>Current Source Inverter</td>
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<tr>
<td>DCP</td>
<td>Diode Conduction Period</td>
</tr>
<tr>
<td>DSCP</td>
<td>Dynamic Short Circuit Protection</td>
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<tr>
<td>el.</td>
<td>Electrical</td>
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<td>FESS</td>
<td>Flywheel Energy Storage</td>
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<tr>
<td>FPGA</td>
<td>Field-programmable Gate Array</td>
</tr>
<tr>
<td>HDL</td>
<td>Hardware Description Language</td>
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<tr>
<td>I/O</td>
<td>Input/Output</td>
</tr>
<tr>
<td>IGBT</td>
<td>Insulated Gate Bipolar Transistor</td>
</tr>
<tr>
<td>IM</td>
<td>Induction Machine</td>
</tr>
<tr>
<td>IMPMSM</td>
<td>Internal Magnet Permanent Magnet Synchronous Machine</td>
</tr>
<tr>
<td>NI</td>
<td>National Instruments™</td>
</tr>
<tr>
<td>PAM</td>
<td>Pulse Amplitude Modulation</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed Circuit Board</td>
</tr>
<tr>
<td>PMSM</td>
<td>Permanent Magnet Synchronous Machine</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse Width Modulation</td>
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<tr>
<td>SMPMSM</td>
<td>Surface Magnet Permanent Magnet Synchronous Machine</td>
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<td>SRM</td>
<td>Switched Reluctance Machine</td>
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<tr>
<td>VSI</td>
<td>Voltage Source Inverter</td>
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<td>ZCD</td>
<td>Zero-Crossing Detection</td>
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CHAPTER 1

Introduction
At the Paris climate conference (COP21) in December 2015, 195 countries adopted the first-ever universal, legally binding global climate change deal. The European Union fulfills its part of the deal within the ‘2030 climate and energy framework’, that sets three key targets: At least 40% cuts in greenhouse gas emissions, at least 27% share for renewable energy and at least 27% improvement in energy efficiency. Often overlooked is that the lion’s share of the commitment will have to be delivered through efficiency of final energy consumption across the economy [Institute for European Environmental Policy, 2016]. As motor driven systems account for approximately 65% of the electricity consumed by EU industry, implementing high efficiency motor systems could save Europe over 200 billion kWh of electricity per year [De Keulenaer et al., 2004].

High-speed electrical motors can play an important role in making motor driven systems more efficient. There are two most commonly perceived advantages of using high-speed electrical motor. Firstly, the reduction of system weight and size for a given magnitude of power conversion. This is particularly desirable in mobile applications, where any savings in weight directly result in reduced fuel burn and emissions. Secondly, adopting high-speed machines in certain applications greatly improves efficiency and reliability as a result of the elimination of intermediate gearing [Gerada et al., 2014]. High-speed electrical machines can thus replace or complement existing high-speed mechanical systems, such as high-speed spindle applications, gas compressors or turbochargers, or even open ways to whole new application areas, such as flywheel energy storage. Nevertheless, operating the machines at high speed requires a particular design of the entire system.

The electrical machine of choice for low-power, high-speed applications is the permanent magnet synchronous machine (PMSM) [Borisavljevic, 2013]. The PMSM has the advantage of having very high efficiency, power density and good dynamic behaviour. The main disadvantage is that the use of permanent magnets can significantly increase the capital cost.

In order to control a PMSM optimally, meaning that we want it to have good dynamic performance and optimal power conversion, accurate knowledge of the rotor position is necessary. The electrical peculiarities of such high speed machines can cause a series of problems with the stability of the control, if the position and speed of the rotor is not known with high precision [Niedermayr et al., 2016]. Measuring the rotor position is usually done using sensors on the rotor shaft, such as a resolver and encoder, or sensors on the stator, such as Hall-effect sensors or measurement coils. However, installation of these sensors increases the cost of the system and during the lifetime of the machine, these sensors can be subject to heavy vibrations and thermal loads [De Belie, 2010], especially in high-speed machines. Furthermore, position sensors are not available at very high speed due to reliability and precision issues.

To improve reliability and reduce costs, a recent trend has emerged to drive machines sensorless. Sensorless or ‘indirect position sensing’ is a denominator for a collection of techniques that estimate the rotor position indirectly, without the use of mechanical sensors but with voltage and current measurements.

This thesis presents the research done towards fully-digital sensorless control of a high-speed PMSM (30 000 rpm, 3kW) on a test set-up developed at Ghent University. The sensorless control is based on the motional back-EMF, the voltage induced in the stator windings by the movement of the rotor’s permanent magnets. The control is completely digital and implemented on a field-programmable gate array (FPGA). FPGAs have several advantages over...
CHAPTER 1. INTRODUCTION

traditional digital signal processors, especially in high speed applications where time constraints on the signal processing are high. Due to their reconfigurable layout and the inherent parallelism that comes from it, suggest that FPGAs will dominate sensor and filtering applications [Banovic et al., 2005]. Another advantage of their reconfigurable lay-out is that it allows rapid-prototyping of applications [Ooi et al., 1988]. Furthermore FPGAs have become increasingly cheap [FPG, 2007], allowing their entry in more and more applications.

Fully-digital sensorless control of PMSMs can be found in literature, for example the Digital Sliding-Mode Sensorless Control from Rivera Dominguez et al. [2014] or the back-EMF Threshold Self-Sensing Method described by Darba et al. [2015]. However, little is written about the fully-digital implementation of sensorless control for high-speed PMSMs, although the use of FPGA technology in motor control looks very promising.

In Chapter 2 “Introduction to High Speed Control” we introduce the reader to general concepts concerning high speed machines and their application areas. We argue why the PMSM, combined with a variable dc-link converter has the preference for most high speed, low power applications. It is then explained why FPGAs complement this arrangement.

Chapter 3 “Basic Principles of PMSMs” covers the working principles of the PMSM, along with its mathematical modelling. An optimal control strategy, ‘Maximum Torque per Ampere’ (MTPA), is given and explained for normal operation and field weakening operation. We then differentiate between the two main types of PMSMs, that is, brushless AC machines and brushless DC machines. Since the brushless DC machine has the preference for high speed, we highlight its behaviour during commutation, since this will improve our understanding later.

In Chapter 4 “Sensorless Control Algorithms”, different position estimation methods are presented, diving them into four main classes: methods based on the back-EMF, observer-based methods, methods based on inductance variation and methods based on flux linkage variation. We study the method using the back-EMF in greater detail and argue why this is the method of preference for high speed, followed by possible practical difficulties that come with implementing this method. The main practical difficulty at high speed is the non-ideal behaviour during commutation, which is a fundamental constraint for sensorless algorithms based on the back-EMF and is, as we will find out later, is difficult to alter. Lastly, we introduce an alternative method based on the back-EMF that alleviates most practical difficulties. This method determines the commutation instants for a BLDC on the zero-crossings of the line-to-line back-EMF.

The test set-up on which the experimental research is conducted, is described in Chapter 5. An overview of the hardware is given, combined with a more detailed explanation of the power electronics and control electronics. We give an overview of the software and some important considerations regarding programming software on FPGAs, as correct timing and high throughput are of vital importance for the high speed control.

In 6 “Methodology” a brief overview of the methodology of the research is given. The research comprised of two main parts: basic development of a sensorless algorithm on a low speed machine and extended development on the high speed machine developed at Ghent University.

The experiments and results on the low speed machine are given in Chapter 7. We present a novel commutation algorithm that uses detection intervals based on a debounce filter to reject false zero-crossings of the line-to-line back-EMF. This algorithm allowed successful sensorless commutation of the low speed machine with almost no tuning required.
We then move our research to the high speed machine in Chapter 8 “Extended Development on a High Speed Machine”. The sensorless algorithm developed in the previous chapter no longer suffices due to the non-ideal behaviour during commutation, the conduction of the freewheeling diodes. In simulation we experiment with two different options to shorten this period of conduction, but none of them can shorten this period significantly. We therefore decide to build our algorithm such that it ignores this period of conduction. Two different solutions are found, one using a delay to ignore this period and one using current measurements. Although the first solution requires a little more tuning for different speeds, it is more robust than the second solution. The complete digital algorithm is then tested on the high speed test set-up. Both algorithms perform well from very low speeds up to 10 000 rpm. The algorithm using the delay shows promise to work over the full speed range, but could only be tested until 15 000 rpm due to technical problems.
CHAPTER 2

Introduction to High-Speed Control
2.1 Today’s Trend Towards Higher Efficiencies

For the past decades, due to the rapid development of emerging economies and the growth of household consumption level, the gap between energy supply and demand has become more and more prominent. A decoupling between economic growth and environmental impact is highly needed. Improving energy-efficiency is a straightforward way to obtain such decoupling. As shown by De Keulenaer et al. [2004], switching to energy-efficient motor systems would save the European Union:

- up to 202 billion kWh in electricity consumption annually
- €4.7 billion in environmental costs by 2025
- 79 million tonnes of CO2 emissions (at that time by 2015), one quarter of the EU’s Kyoto target
- 45 GW reduction in the need for new power plant capacity by 2025

The key challenges to increase efficiency in systems driven by electrical machines are situated in three areas [Mecrow and Jack, 2008]:

- (area 1) to integrate the drive and the driven load to maximise system efficiency
- (area 2) to extend the application of variable-speed electric drives into new areas through reduction of power electronic and control costs
- (area 3) to increase the efficiency of the electrical drive itself

2.1.1 High-Speed Electrical Machines Meeting Key Areas

High-speed electrical machines are electromechanical transducers with typical operational speeds in excess of 10 000 rpm and rpm/$\sqrt{\text{kW}}$ in excess of $1 \times 10^5$ [Gerada et al., 2014]. It appears that high-speed electrical machinery tackles the three key areas outlined above exceptionally well and shows promise to increase system efficiency.

A commonly perceived benefit (area 1) in certain applications is the improved efficiency and reliability as a result of the elimination of intermediate gearing (direct drives). Besides that, high speed electrical machines open a way to several new applications (area 2), where they either replace existing high speed mechanical systems or complement the existing high-speed mechanical system. Lastly, recent developments in materials and design allow these high speed drives to become increasingly efficient (area 3). An overview of these developments can be found in [Gerada et al., 2014]. In what follows, a few emerging applications of high speed electrical machines will be outlined, supporting the claim that these machines can be integrated to maximise system efficiency and that they push the application of variable-speed electrical drives into new exciting domains.
2.1.2 Overview of Emerging Applications

2.1.2.1 High-Speed Electrical Machines for More Electric Engines

An advantage of high-speed machines is the reduction of system weight for a given magnitude of power conversion. This is particularly desirable in mobile applications, where any savings in weight results directly in reduced fuel burn and emissions. Consequently, the concept of having high-performance traction machines integrated within hybrid drive-trains to improve fuel efficiency and reduce emissions is now commonplace in vehicles. There are several possible applications where high-speed electrical machines are built around a future engine, for example mounting the electrical machine on the same shaft as the turbine and the compressor wheels in a turbocharger.

2.1.2.2 Flywheel Energy Storage Systems

![Figure 2.1: Example of a flywheel energy storage system assembly.](image)

Flywheel energy storage systems (FESS) operate by mechanically storing energy in a rotating flywheel. Electrical energy is stored by using a motor that spins the flywheel, thus converting the electric energy into mechanical energy. To recover the energy, the same motor is used to slow the flywheel down, converting the mechanical energy back into electrical energy. Traditional flywheel designs have large diameters, rotate slowly, and have low power and energy densities. More modern flywheels are designed to rotate at higher speeds. Such flywheels achieve higher power densities than the NiMH batteries typically used in hybrid vehicles, albeit having lower energy densities [Gerada et al., 2014]. This gives FESS a number of advantages over battery technologies in applications where high peak power output for a short amount of time is required.
2.1.2.3 Industrial Air Compressors and Air Blowers

In many industrial applications, there is an ever-increasing demand for higher quality and oil-free compressed air. High-speed electrical machines that operate at power levels of 100-500 kW and speeds of 80-15 000 rpm, using magnetic or air bearings, are being used in the latest generation ‘oil-free’ direct-drive industrial compressors, in the range of 4-9 bar [Gerada et al., 2014].

2.1.2.4 Other applications

The drive for the development of high-speed electrical machines can also be found in the following applications:

- high speed spindle machining
- gas compressor applications
- microturbines

2.2 Different High-Speed Electrical Machines and Their Drives

Following the above, a clear need for high-speed electrical machinery has been established. The most suitable electrical machines for high-speed will now be considered and a limited trade-off study will be made. A specific inverter topology for high speed will be highlighted.

2.2.1 Electrical machines suitable for high speed operation: A Trade-Off Study

Of the many types and variants of electrical machines that exist, the induction machine (IM), synchronous machine with permanent magnets (PMSM) and the switched reluctance machine (SRM) might be the most suitable for high-speed operation. Excluded from this trade-off study are DC machine and the universal motor because of the presence of commutator brushes in these topologies. The friction losses from these brushes limit the maximum speed to 25000 rpm. Other more exotic topologies such as piezo drives and homopolar and heteropolar machines are also not considered. The IM, SRM and PMSM all have their respective drawbacks and advantages. In what follows, the three types of machines are compared, based on a selection of specifications. Of course, depending on the application, other requirements might present themselves. The specifications are:

- cost
- losses and efficiency
- rated speed, rated power and power density
- load profile
CHAPTER 2. INTRODUCTION TO HIGH-SPEED CONTROL

Cost. Soong et al. [1999] argue in their machine selection for a 20 kW, 57 krpm centrifugal compressor that PMSMs require expensive materials (inconel, neodymium-iron-boron magnets) and techniques (sintering of the magnets, fiber wrap to retain the magnets) making the relative cost compared to IMs twice as high. However, one has to note that the authors had a production department at hand familiar with the IM. Their design of a high speed IM required the copper-based metal matrix composite Glidcop, which is difficult to fabricate and has to be brazed instead of welded. Additionally, the authors estimated that the cost for the SRM would be in the middle of the IM and PMSM, but the SRM did not meet their efficiency requirements. A further note in favor of the PMSM is that their proposed design would reach over 1.5 times the required speed.

Losses and efficiency. Losses in electrical machines can be split in stator copper losses, rotor copper losses and iron losses, with the latter consisting of eddy current losses and hysteresis losses. Since hysteresis losses increase approximately linearly with the frequency and eddy current losses with the square of the frequency [Melkebeek, 2014], we expect them to be a major source of losses for high-speed electrical machines. To counter eddy currents, slitting is applied to IMs. Making axial slits in the rotor has the effect of guiding the fundamental flux component into the rotor while presenting a higher impedance path to the eddy currents traveling on the rotor surface. However, slitting also increases air-gap friction loss, which at high speed can even outweigh the reduced eddy current loss [Gerada et al., 2014]. Besharati et al. [2015] designed an SRM for high speed automotive traction. The theoretical efficiency of their machine is 91%, although the authors admit that air friction losses were not considered even though they can be significant due to the teeth of the rotor. Generally, the efficiency of SRMs is not that high and depends on the air-gap of the machine. Vibrations at high speed operation might not allow a sufficiently small air-gap required for high efficiencies. The PMSM generally has the highest efficiencies due to the absence of rotor current losses [Soong et al., 1999].

Rated speed, power and power density. Although a little bit dated, Borisavljevic [2013] outlines some design windows for high-speed electrical machines. The conclusion at the time of writing is that PMSMs are most suitable for low-power high speed applications. For lower speeds (≤ 9000 rpm) and higher power (2 MW up to 30 MW), the synchronous machine fed from current source inverters is the favourable solution, while for high speeds (≤ 100 000 rpm) and lower power (≤ 2 MW), the IM is typically favoured. The conclusion was based on four considerations: 1) the mechanical stresses in the rotor; 2) critical speeds; 3) rotor cooling; 4) specific power output per rotor volume.

An interesting parameter related to speed and power is the concept of ‘Dynamical Speed’. This useful ‘guide number’ $G$, as introduced by van Millingen and van Millingen [1991], can be used to assess the likely severity of dynamical problems at high speed. These problems include high values of bearing DN (diameter times speed), critical speeds, stresses and sensitivity to good balancing. The guide number is given as

$$G = n \sqrt{P} \text{ in [rpm} \sqrt{\text{kW}}]$$

(2.1)

In [van Millingen and van Millingen, 1991] a classification of the severity of expected problems in relationship to the values of $G$ is given: The Dynamical Speed can give a good first estimation whether or not a required combination of speed and power will be mechanically feasible.

Load profile. Another interesting comparison is made by Schultz and Huard [2013], namely comparing the motors in terms of efficiency for different operating points (the ‘drive cycle’ for
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\[
\begin{array}{|c|c|}
\hline
\text{range of } & \text{Expected problems} \\
\frac{G}{10^{3} \text{ rev/min} \cdot \text{kW}} & \text{} \\
< 10^3 & \text{negligible} \\
1 - 5 \cdot 10^5 & \text{low} \\
5 - 10 \cdot 10^5 & \text{moderate} \\
> 10^6 & \text{severe} \\
\hline
\end{array}
\]

Table 2.1: Classification of expected problems in relation to G. [van Millingen and van Millingen, 1991]

an electric car). Although not high-speed, we can expect that these general considerations still apply. For low speed and high torque, the losses of the IM are about three times higher than the losses of the PMSM, mainly due to the stator copper losses. At medium speed and medium torque, the PMSM has extremely low losses and from this point the efficiency of the IM starts to improve quite substantially. However, losses for the IM are still twice as high. For high speed and low torque the IM outperforms the PMSM. This can be intuitively explained: the PMSM has a constant magnetic flux due to its permanent magnets. For low torque, this magnetic flux is excessive and causes iron losses. On the other hand, the IM only generates the required amount of flux, resulting in lower iron losses. Overall, the authors conclude that for ‘city driving’ (mainly low speed, high torque), and ‘rural driving’ (medium speed, medium torque) the PMSM outperforms the IM significantly, while for ‘highway driving’ the PMSM only uses 1% less energy. Overall, we can extrapolate that PMSMs are most suitable for applications with significant speed changes, both in frequency of the changes and in magnitude.

2.2.2 Inverter topology suitable for high speed operation

The very high fundamental frequency that high speed electrical machines need to be supplied with pose significant challenges for the power electronics supplying the machine. A full review of existing inverter topologies would be beyond the scope of this thesis, but is extensively discussed by Zwyssig et al. [2006], comparing different topologies in combination with a PMSM. The authors conclude that the use of a voltage source inverter (VSI) with block commutation offers low switching losses, simple control and easy implementation of sensorless control (see later on).

In literature, the chosen inverter topology and commutation strategy is also referred to as ‘variable DC-link inverter’ or Pulse Amplitude Modulation (PAM) inverter [Zwyssig et al., 2007]. It consists of a standard voltage source inverter topology and an additional dc-dc converter as shown in Figure 2.2. The inverter part is controlled in six-step or block commutation, which means that each switch is conducting for 120 electrical degrees and, therefore, switched only with the fundamental frequency of the machine. The dc-dc converter modulates the amplitude of the current blocks. On the other hand, the more commonly used Pulse Width Modulation (PWM) changes the amplitude of the current blocks by switching the inverter at a much higher rate. Figure 2.3 illustrates the difference between the two strategies for a simple block wave:

Kim and Youn [2002] compared the classical PWM scheme with the variable DC link inverter for high-speed control and came to two arguments that favour PAM over PWM.

Firstly, although PWM provides better control dynamics [Zwyssig et al., 2007], the advantage of only having to switch at the fundamental frequency makes the combination of a PAM inverter
with a PMSM very attractive for high speed operation. At high speed, only a few PWM pulses can be used for the speed control during the on-time of the interval. As an example: since the 60° interval of a 2-pole motor becomes 200 µs at 50000 rpm, if a switching frequency of 16 kHz is employed the number of PWM pulses during 60° is only 3.2. Unless the resolution of the pulse width is considerably higher, this may result in a speed ripple at steady state.

Secondly, PWM may cause commutation delays or an irregular switching frequency that have a significant influence on the phase current and drive performance at high speed. Figure 2.4(a) shows a case of commutation delay. Since the commutating instants depend on the rotor position (see next chapter), it does not usually coincide with the end of a PWM switching period. In this case, the commutation is normally performed synchronised with the end of the present PWM period to start the next inverter sequence. Even though this delay can be neglected in a normal speed range, it becomes problematic at high speed since the 120° intervals become relatively small. To avoid such an undesirable delay, the next inverter sequence has to be applied as soon as the commutation signal interrupt occurs. In some PWM schemes, this may yield an irregular
switching frequency much larger than the switching frequency \( f \), under high-duty conditions as shown in Figure 2.4(b).

![Diagram](image_url)

**Figure 2.4: Relation between PWM switching period and commutating instant.** (a) Commutation delay (b) Irregular frequency

### 2.3 Recent Advancements in Signal Processing: FPGAs

To run the high-speed electrical machine efficiently and provide a way to control speed and torque, the inverter needs specifically timed controlling signals to tell when it needs to switch its power switches (IGBTs, Mosfets, Thyristors,...) on or off. These signals are usually generated by digital signal processors (DSP). DSPs have the advantages of simple circuitry, software control and flexibility in adaptation to various applications; it suffers the disadvantages of sluggishness and limited computation resources due to sequential computation, complicated design process and long development time cycle [Ooi et al., 1988]. The limited computational resources are especially disadvantageous for high speed operation, where signals have to be generated in periods of microseconds or even hundreds of nanoseconds. Field Programmable Gate Arrays (FPGAs) alleviate these problems by providing an economic solution and a fast circuit response due to its simultaneous instead of sequential execution. The ‘true parallelism’ of FPGAs allow different control and measurement loops to be run simultaneously. Furthermore, FPGAs are by nature reprogrammable and thus very convenient for laboratory implementation of a project.
To understand how FPGAs provide this advantage, some basic understanding of their inner workings is required. An FPGA is a reprogrammable chip composed of three basic components: logic blocks, programmable interconnects and Input/Output (I/O) blocks. The logic blocks are a collection of lookup-tables, multipliers and multiplexers that generate a desired logic output from an input. Signals are routed through the programmable interconnects from one logic block to the next. Communication to the surrounding circuitry is done by the I/O-blocks. An FPGA is programmed by physically changing the interconnects, using a hardware description language (HDL), which is in fact a schematic design of the lay-out. This means that any program to be implemented on an FPGA needs to be translated first into a HDL, a process that can be time consuming. The advantages of this reconfigurable layout and the inherent parallelism that comes from it, suggest that FPGAs will dominate sensor and filtering applications [Banovic et al., 2005]. Furthermore FPGAs have become increasingly cheap [FPG, 2007], a trend that has no reason to stop in the near future. This offers possibilities for cheap and extremely fast electrical machine drives controlled by FPGAs, no longer limited by processing throughput but solely by the maximum switching frequency of the power switches.

2.4 Conclusion

In this introductory chapter, a need for more efficient electrical machines has been established, and we have explained how high speed electrical machines fulfil this need by meeting three key areas: improving the efficiency of the complete mechanical system, improvement in efficiency of the machines themselves and allowing to introduce electrical motors in completely new application areas. Some of these new application areas were briefly explained. Subsequently, a trade-off
study was made between electrical machines suitable for high speed operation. Although one could argue that for certain applications the IM could be a better choice, the PMSM is overall more efficient, lighter and can handle highly dynamic loads better. Moreover, the combination of a PMSM with a VSI with block commutation is easy to implement and has clear advantages for high speed operation. The very promising FPGA technology can support this arrangement, providing timely signals to control the machine even at extremely high speeds, due to its inherent parallel execution.

The focus of this thesis, limited in time, is therefore high-speed operation of PMSMs with a PAM inverter and controlled by FPGAs. The combination of high speed PMSMs with PAM inverters is not new in literature, and a speed record of 1 000 000 rpm was attained by Zwyssig et al. [2007] with such an arrangement. Fully-digital control of PMSMs can also be found in literature, for example the Digital Sliding-Mode Sensorless Control from Rivera Dominguez et al. [2014] or the back-EMF Threshold Self-Sensing Method described by Darba et al. [2015]. However, little is written about the fully-digital implementation of control for high-speed PMSMs, where PAM inverter, PMSM and FPGAs are combined, although the use of the emerging FPGA technology looks very promising for motor control.
CHAPTER 3

Basic Principles of PMSMs
3.1 Classification

In the previous chapter we narrowed down the subject of this thesis to the family of PMSMs because they are particularly attractive for high speed applications. The advantages of PMSMs are high power-to-weight ratio, good dynamic performance, simple and robust rotor construction and absence of brushes. The general physical characteristics of these machines are that the stator is composed of a three-phase winding, in a way that it generates a rotating field. The rotor contains permanent magnets to produce a magnetic field [Glumineau and de Leon Morales, 2015]. This immediately leads to another advantage: since the stator contains the current-bearing windings, and not the rotor, the heat dissipation is better than in DC commutator machines [Al-Hadithi, 1992]. The magnets in the rotor can be placed in several ways. Following magnet position, the PMSM can be classified into four major types [Glumineau and de Leon Morales, 2015]:

**mounted magnets type**  This configuration is easy to obtain, gluing the magnets to a solid rotor and often covering them with a non-ferrous sleeve, to increase mechanical strength. The inductances do not depend on the rotor position and the inductance of the direct axis \( L_{sd} \) (aligned with the magnetic flux, N-S axis of the magnet) and quadrature axis \( L_{sq} \) (orthogonal to the magnetic flux) are practically equal.

**Inset magnets type**  The space between the magnets is filled with iron (see Fig. 3.1), causing a small difference between direct axis and quadrature.

**Interior magnets type**  The magnets are buried in the iron of the rotor. The rotor’s magnetism is anisotropic, the inductances depend on the rotor position. The inductance along the direct axis is generally lower than the inductance along the quadrature axis [Melkebeek, 2016]. The small ‘iron bridges’ where the magnets are close to the surface can reduce robustness in this construction and furthermore manufacturing and control are more complicated for this type.

**Flux concentrating type**  The magnets and their axes are radial, creating a saliency effect. This lay-out is also referred to as spoke type.

Surface mounted magnets (SMPMSM) and interior magnets (IMPMSM) are most commonly used in the industry, because flux concentrating type machines require magnetic isolation of the rotor shaft and bearings.

A second classification can be made based on the profiles of the back-electromotive force (back-EMF), the waveform of the voltage induced by movement of the permanent magnets. The shape of the back-EMF mainly depends on the layout of the windings in the stator, where a sinusoidal winding distribution will generate a more sinusoidal back-EMF and a concentrated winding distribution a more trapezoidal back-EMF. The lay-out of the rotor magnets also influence the back-EMF shape. When the back-EMF is trapezoidal, the machine is called a brushless DC machine because its working principle is very similar to the DC machine. With a sinusoidal waveform the machine is called a Brushless AC Machine [Melkebeek, 2016] or sometimes referred to as PMSMs, because their operation is comparable to synchronous machines.
CHAPTER 3. BASIC PRINCIPLES OF PMSMS

3.2 Working Principle of PMSMs

3.2.1 Brushed DC Machine

To understand PMSMs, a look into their brushed counterparts will make the reasoning behind the construction and function much easier to understand. In an ideal DC commutator machine, torque is proportional to the product of flux and armature current 3.1.

\[ T = K I \Phi \]  

(3.1)

The speed of the machine is proportional to the voltage supplied to the machine and inversely proportional to the field flux:

\[ E_a = K \Omega \Phi \]  

(3.2)

Commutation by use of brushes ensures that the armature current layer is always co-incident with the field flux axis. Thus, the generated torque is maximal for a given current and flux. Figure 3.2 explains how commutation ensures that torques always align.

Naturally, the use of brushes in a DC machine is a major drawback. As stated before, the brushes limit the use of the machine at high speed, but also at normal speeds they are to be avoided, since they are always under severe mechanical and electrical stresses. This requires that the machine is regularly maintained and thus the longevity of the machine is reduced. Thus, a need for a non-mechanical commutation naturally arises. We explain in the next section how this is obtained.
CHAPTER 3. BASIC PRINCIPLES OF PMSMS

3.2.2 Brushless Machines

The PMSM can be seen as a DC machine turned inside out. The permanent magnets are installed on the rotor and not on the stator, while the wounded poles are found on the stator, not on the rotor. Commutation is no longer physical, but caused by switches in the windings of the stator. Consider Figure 3.3: to spin the rotor in a clockwise direction, phase C should be energised in the opposite direction, subsequently phase B in forward direction, and so on.

One can easily see that this is not a very efficient way to produce torque, since only one phase is energised at a time. To produce more torque two or three phases should be energised at the same time. In practice, PMSMs are fed with a sinusoidal current or current blocks. Both types of waveforms can be used interchangeably for a PMSM, but a PMSM with a more trapezoidal back-EMF will generate smoother torque with current blocks, while a sinusoidal machine combined sinusoidal current also generates a flat torque characteristic (cf. infra). Before further differentiating between these two types of machines, we will now derive a general mathematical
CHAPTER 3. BASIC PRINCIPLES OF PMSMS

Figure 3.3: Three phase commutation of BLDC Darba [2016]

model for the PMSM, regardless of the type of current supply.

3.3 Mathematical Modelling of PMSMs

To gain better understanding of the operation of a BLDC Machine, we derive the classical dynamic model starting from the most general Space Vector notation for voltages and currents:

$$ u_s^s = R_s \cdot i_s^s + \frac{d}{dt} \Psi_s^s $$ (3.3)

with $R_s$ the stator resistance, assumed symmetric for the three stator windings, $u_s^s$ the stator voltage, $i_s^s$ the stator current and $\Psi_s^s$ the stator magnet flux vector. This equation is first transformed using the Clarke transformation into a two-axis $\alpha\beta$-reference plane [?]. Using the Park transformation to transform this fixed reference frame into a rotating reference frame that is synchronous with the field yields:

$$ u_f^s = R_s \cdot i_f^s + \frac{d}{dt} \Psi_f^s + j\omega_s \Psi_f^s $$ (3.4)

With, for the flux:

$$ \Psi_s^f = L_s \cdot I_f^s + \Psi_f^p $$ (3.5)

Where the change from superscript $s$ into $f$ indicated the change of reference frame. We attach the real axis (direct axis or $d$-axis) of the synchronous coordinate system to the rotor flux and the imaginary axis (quadrature axis or $q$-axis) perpendicular to the rotor flux, according to the right hand rule. Thus the quadrature component of the rotor flux becomes equal to zero [Quang and Dittrich, 2015]. We may write:

$$ \Psi_{sd} = L_{sd} i_{sd} + \Psi_p $$

$$ \Psi_{sq} = L_{sq} i_{sq} $$ (3.6)

Generally, $L_{sd}$ and $L_{sq}$ are not equal and dependent on the construction of the pole gaps on the rotor surface. However, for PMSM with surface permanent magnets the inductances are nearly identical. For the internal magnet type rotor, the $q$-axis inductance is generally larger than the $d$-axis inductance.
It is interesting to note that, in this choice of reference frame, a current with a negative d-axis component will counteract the permanent magnet flux. This demagnetizing effect is called field weakening [Gerling, 2015]. Field weakening allows to increase the speed of the motor above the nominal speed, without exceeding the nominal voltage. We refer to the speed equation of the brushed DC motor 3.2 for an explanation. Substituting (3.5) and (3.6) into (3.4) yields:

\[
\begin{align*}
    u_{sd} &= R_s i_{sd} + L_{sd} \frac{d}{dt} i_{sd} - \omega_s L_{sq} i_{sq} \\
    u_{sq} &= R_s i_{sq} + L_{sq} \frac{d}{dt} i_{sq} + \omega_s L_{sd} i_{sd} + \omega_s \Psi_p
\end{align*}
\] (3.7)

The general electromechanical torque equation for a rotating field electrical machine is [Quang and Dittrich, 2015]:

\[
T_{em} = \frac{3}{2} N_p \Im \{ \Psi_s^* i_s \}
\] (3.8)

Where \(N_p\) is the number of pole-pairs and the asterisk represents the complex conjugate. This gives, in the \(dq\)-reference frame:

\[
T_{em} = \frac{3}{2} N_p (\Psi_{sd} i_{sq} - \Psi_{sq} i_{sd})
\] (3.9)

After inserting the equation for the flux (3.5), we obtain a useful expression for the electromagnetic torque:

\[
T_{em} = \frac{3}{2} N_p (\Psi_p i_{sq} + i_{sd} i_{sq} (L_{sd} - L_{sq}))
\] (3.10)

We find that the torque production for PMSMs can be expressed as a sum of a main torque component and a reluctance torque component. For a SMPMSM the inductances \(L_{sq}\) and \(L_{sd}\) are about the same, so the reluctance component contributes very little to the torque. For an internal magnets type PMSM however, the q-axis reactance is larger than the d-axis reactance [Melkebeek, 2016], meaning that a negative d-axis component of the current will contribute positively to the total torque.

The above considerations are summarised in Figure 3.4. The stator current is drawn alongside its quadrature and direct component. According to the torque equation, only the quadrature component will contribute significantly to the torque for trapezoidal PMSMs. For the sake of completeness, the voltage induced by the permanent magnet flux, \(E_p\) is also drawn. This voltage is the back-EMF. The angle and magnitude of the stator voltage \(U\) can be deduced from (3.7).

### 3.4 Optimal control strategy

#### 3.4.1 Aim of control

The torque equation of an ideal DC commutator machine (3.1) is quite favourable, since to obtain a required torque one simply provides a certain armature current. Moreover, because the symmetry axis of the armature current layer is co-incident with the field flux axis, the torque is maximal for a given armature current [Melkebeek, 2016]. Furthermore, orthogonality between the flux and current (in case of no saturation) ensures that no additional dynamics are introduced when varying the current. In rotating field machines torque results from the interaction of a rotating (sinusoidal - at least fundamentally) current layer \(a(x,t)\) and field distribution \(b(x,t)\).
The torque resulting from the field-current interaction is constant in time, but its magnitude depends on the angle $\theta$ between the symmetry axes of field and current layer [Melkebeek, 2016]:

$$T \sim \hat{A}\hat{B}\cos\theta$$  \hspace{1cm} (3.11)

It is clear that for a given current layer, the torque is maximal for co-incident field and current layer, i.e. $\theta = 0$. Consequently, a behaviour analogous to the DC commutator machine can be obtained with a rotating field machine if:

1. the torque producing current can be controlled independently
2. the angle $\theta$ can continuously be controlled at $\theta = 0$

This operating condition is called field orientation. When $\theta$ is set at other values than zero, the control is commonly called vector control. Note however that in this case current and field will interact with each other, introducing additional dynamics.

### 3.4.2 Maximum Torque per Ampere

Depending on the demands of the application, different control strategies within vector control can be used to optimise different performance objectives. Such objectives can be minimisation of active and total losses, power factor maximisation, maximum torque per ampere, maximum torque per voltage and maximum power transfer [Bozhko et al., 2016]. These optimisation methods are designed for steady-state operation. Maximum Torque per Ampere (MTPA) control minimises the stator current for a given machine torque. The basic MTPA control objective is
achieved by controlling stator current main and reactance components, expressed in terms of previously introduced \( dq \)-reference frame. Of course, for a permanent magnet machine, the rotor flux cannot be controlled.

We refer again to the torque equation of a PMSM as previously introduced:

\[
T_{em} = \frac{3}{2} N_p (\Psi_p i_{sq} + i_{sd} i_{sq} (L_{sd} - L_{sq}))
\]

Using this equation, we can draw characteristics of equal torque value in the \( dq \)-plane, for both SMPMSM and IPPMSM. The difference between the two characteristics results from the fact that for SMPMSM \( L_{sd} \approx L_{sq} \) and for IPPMSM \( L_{sd} < L_{sq} \).

From Fig. 3.5 it is obvious that torque control is much easier for machines with surface mounted magnets (\( L_{sd} \approx L_{sq} \), here a linear relation exists between the current \( i_{sd} \) and the torque) than for machines with buried magnets (\( L_{sd} < L_{sq} \), for these machines \( i_{sd} \) and \( i_{sq} \) have to be controlled simultaneously and in addition, the relation is non-linear) [Gerling, 2015]. In what follows, the more complex case model of IPPMSMs will be used, but at any point these considerations can be simplified for the SMPMSM.

The solution to the question “how to choose the current components \( i_{sd} \) and \( i_{sq} \) in order to obtain maximum torque” can be answered for three cases:

### 3.4.2.1 Nominal Speed Operation

For a certain torque, the total current is minimal if the current vector and the gradient of the torque have the same direction. In Fig. 3.6 this is represented by the dark blue curve. The upper torque limit is imposed by the nominal stator current, which should not be exceeded for thermal reasons.

Note however that for SMPMSMs this is simplified to orienting the current vector coincident with the back-EMF \( E_p \), since the torque characteristic is then as in Figure 3.5 (left).
3.4.2.2 Minor Field Weakening Operation

In this operation, the voltage limitation imposed by the maximum DC-bus voltage becomes relevant. By translating the voltage into current limits, we find elliptical functions:

\[
u^2_{\text{max}} \geq u^2 = u_d^2 + u_q^2 = (\omega L_q i_q)^2 + (\omega L_d i_d + \omega \Psi P)^2 \Rightarrow \frac{u^2_{\text{max}}}{\omega^2} \geq (L_q i_q)^2 + (L_d i_d \Psi P)^2 \tag{3.13}
\]

The maximum torque is now of course lower than in the previous case. The optimum curve for reaching the required torque now partly proceeds along the voltage limit curve, see Fig. 3.7. The current limit is also still relevant; following the voltage limit, we reach the current limit at a certain point. We see that the current vector now leads the back-EMF \( E_p \).

Figure 3.6: Optimal torque characteristic without relevant limits (based on [Gerling, 2015]).

Figure 3.7: Optimal torque characteristic for minor field weakening (based on [Gerling, 2015]).
3.4.2.3 Strong Field Weakening Operation

With strong field weakening, only the voltage limit is relevant; the limit of the maximum total current is not longer reachable. The current now leads the back-EMF even more. We now find the additional black characteristic in Fig. 3.8 that indicates the decreasing torque for further increasing current, while the voltage limit is held. As these operating points can also be reached with lower current, they have no practical relevance. Therefore, this operating area is excluded.

![Figure 3.8: Optimal torque characteristic for strong field weakening (based on [Gerling, 2015]).](image)

3.5 Brushless AC Machines

Now that we have a general model and control strategy for PMSMs, we can further differentiate between the two types of machines. As stated before, BLACs or ‘PMSMs with a sinusoidal distribution of air-gap induction’ prefer sinusoidal current. Operating the supplying inverter appropriately with a high switching frequency, the phase currents of the motor can be nearly sinusoidal because of the low-pass effect of the phase impedances. This creates an artificial three-phase system with variable voltage and variable frequency. It can be shown that such operation, combined with a sinusoidal back-EMF generates a constant torque, as shown in Figure 3.9.

However, in Section 3.4 it became clear that, in order to optimally control a PMSM, knowledge of the position of the field with respect to the stator phases is necessary. For a PMSM, the position of the magnetic field is fixed by the position of the permanent magnets, and thus by the position of the rotor shaft. Consequently, installing a sensor on the shaft, such as a resolver, encoder, tachometer, Hall sensors or measurement coils should be sufficient to determine the rotor position and indeed the position of the field. One can intuitively see that sinusoidal-fed machines require a fairly high resolution to synchronise the stator currents with the rotor flux. Indeed, it can be shown that the sinusoidal machine requires a resolution of 5.6° electrical [Melkebeek, 2016].

Immediately, one notices the drawbacks of such a supply with respect to high-speed operation. The PWM used to obtain the sinusoidal current waveforms suffers from the hindrances described

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in 2.2.2 at high speed, making it difficult to shape the waveforms. Furthermore, BLAC drives usually use an encoder to obtain the required position resolution. However, encoders are costly, have low reliability, are complicated to install and most importantly, limit the speed range [Kim et al., 2011a].

3.6 Brushless DC Machines

Brushless DC machines or ‘PMSMs with a trapezoidal distribution of air-gap induction’ on the other hand are fed with current blocks with a spatial displacement of 120° electrical, as shown in Figure 3.10. In the ideal case as presented here, the generated torque is constant. Note that at any given point, one of the phases has zero current. Again, for optimal control and better dynamics, the rotor position is required to synchronise the current commutation with the actual rotor position. Since commutation only happens once every 60° electrical, the resolution for the position should not be as high as for the BLAC machine. Nevertheless, even small deviations from the optimal commutation instant can introduce quite substantial torque ripple and make the machine draw excessive current for a given torque (no longer MTPA). If
we examine Figure 3.10 closely, we notice that there are 6 possible combinations in terms of energising phases. Therefore, this operation is often called *Six-Step Commutation*. Hall-effect sensor arrangements are most commonly used to determine the commutation instants. Hall-effect sensors are transducers that change their output voltage in response to a magnetic field. They are mounted in the stator, shifted over 120° electrical. The Hall-effect sensors each output a binary signal, for example 'high' when the north pole of the permanent magnet passes the sensor and 'low' when the south pole passes. Consequently, the duration of a 'high' or 'low' signal is precisely 180° electrical. Combined with the 120° displacement, this provides 6 possible 'high' or 'low' combinations of the sensor outputs, precisely enough to uniquely define every combination of conducting phases. It is important to stress that the Hall-effect sensors have to be mounted accurately for the sake of aligning the current vector precisely with the back-EMF vector to obtain MTPA. The combination of rotor position information and inverter now act as an electrical commutator, just like the mechanical commutator in a DC machine, hence the name *Brushless DC Motor*. The above can be summarised in the following table, where the hall signals determine the sequence of energised phases for a clockwise rotation of the machine. In Figure 3.12 sequence No. 2 is highlighted on the inverter and a wye-connected machine.
CHAPTER 3. BASIC PRINCIPLES OF PMSMS

3.6.1 Non-Ideal Behaviour During Commutation

The assumption that current immediately stops flowing in the non-energised phase (for example phase V in Fig. 3.12) after opening one of its switches is actually an idealisation that does not hold in reality. Due to the inductive behaviour of the stator windings, current has the tendency to keep flowing and the voltage across the switch would keep rising until the switch breaks down. To prevent this from happening, anti-parallel diodes (so-called ‘free-wheeling diodes’) over the switches can take over the current and provide an alternative conduction path. As a consequence, during the time that a free-wheeling diode is conducting, all three phases are conducting current instead of the theoretical two. There are two conduction paths, the main
one being from the supply through the switches that are on and two windings, and the other one being through one of these two windings and the third winding that was switched off, over the free-wheeling diode. This is illustrated in Figure 3.13 for switching from Sequence No. 1 to Sequence No. 2 as given in Table 3.1.

Figure 3.13: Conduction path during free-wheeling period for switching from Seq. No. 1 to Seq. No. 2.

The phase current waveform thus deviates from the ideal block wave and is more trapezoidal, as shown in Figure 3.14. The diode conduction period is influenced by the winding resistance, the back-emf, the winding inductance, the load, and in particular by the ratio \( \frac{\omega(L-M)}{R} \), where \( \omega \) is the electrical pulsation, and \( R, L \) and \( M \) are the winding resistance and self- and mutual-inductances per phase, respectively [Zhu et al., 1995]. We can already establish that for high speed machines the free-wheeling period will be quite long, since the electric pulsation is relatively high, to the point that the phase currents flow more or less continuously.

Figure 3.14: Non-ideal phase current waveform (based on [Zhu et al., 1995]).

3.7 Conclusion

In this chapter we have deepened our understanding of PMSMs. A classification by both rotor lay-out and winding distribution was made. Subsequently, the basic working principle of PMSMs was explained and related to the simple brushed DC machine. After constructing a general mathematical model of the PMSM, an optimal control strategy, ‘maximum torque per ampere’ was chosen and explained for different operating points. We then further differentiated between Brushless AC and Brushless DC machines, which are essentially both PMSMs with a different
supply. We highlighted how the control strategy differs for both types of supply, and why the BLDC type supply has the preference for high speed machines. We will therefore narrow the focus of the subsequent chapter to BLDC machines. Lastly, special attention was given to the non-ideal behaviour during commutation of BLDC machines. Although this has no significant influence on normal operation, the free-wheeling currents will complicate sensorless commutation. Why sensorless commutation is needed and how it could be implemented is explained in the next chapter.
CHAPTER 4

Sensorless Control Algorithms
4.1 Introduction

In Sections 3.5 and 3.6, the necessity for accurate position knowledge of the rotor flux became clear. Mounting sensors on the shaft, such as a resolver or encoder, high-resolution tachymeter had the preference for BLACs, while low-resolution Hall-effect sensors or measurement coils are sufficient to determine the rotor position for BLDCs. However, installation of these sensors is often costly and during the lifetime of the machine, these sensors can be subject to heavy vibrations and thermal loads [De Belie, 2010]. Failure of a sensor can lead to instability and catastrophic damage to the machine. An additional limitation to the use of mechanical sensors is operation of a machine at high speed, especially with high resolution sensors such as encoders. Furthermore, position sensors for very high speeds could be expensive.

Figure 4.1: Scheme of the control structure needed for indirect position-sensing [Acarnley and Watson, 2006]

To improve reliability and reduce costs, a trend emerged in the late eighties to drive machines sensorless. By “sensorless” we mean the collection of techniques that estimate the rotor-position by on-line analysis of the machine’s voltages and currents [Acarnley and Watson, 2006]. The term sensorless could thus be considered misleading in this regard: The techniques are position-sensorless, but do require voltage and current measurements. “Indirect position sensing” is a more justified term. Voltages and current measurements are in most drives readily available, often to monitor if the machine is operating safely. In the following sections, different existing position estimation techniques for BLDCs are explained and techniques applicable to BLACs are briefly mentioned. An appropriate estimation technique is chosen for high speed BLDC machines, within the confines of this thesis.

Ideally, a sensorless technique should be able to cover the whole speed range of the machine, from start-up to nominal speed and if possible even in field-weakening. Operation in field-weakening has important implications for position sensing: the current needs to lead the voltage by a certain angle (dependent on the speed) [Safi et al., 1995]. This means that rotor-position sensing for BLDCs requires a much higher resolution than the 60° electrical provided by Hall-effect sensors, since commutation instants are no longer fixed in time with respect to the back EMF but may lead with a certain angle.

A number of different methods exist to estimate the position of the rotor. In this chapter, a state-of-the-art is given with their respective advantages and disadvantages. We give special
attention to whether or not a given estimation method is suitable for high speed operation. From these methods we choose one that fulfils the requirements for high speed the most. The practical difficulties that are related to this method are mentioned, together with a useful variant of this method that solves some of these difficulties.

4.2 Position Estimation Methods

4.2.1 Methods Based on Motional Back-EMF

Spatial movement of the rotor permanent magnet flux vector induces a motional back-EMF in the stator windings. Since the instantaneous magnitude of this EMF is proportional to the the relative position of the permanent magnets, the EMF contains information about the position of the rotor. Unfortunately, the EMF is not directly measurable at the terminals of the machine: (3.4) from the previous chapter shows that the phase voltage also contains a resistive voltage drop and a drop caused by the self-induction of the windings. A further obstacle to back EMF sensing is that the EMF is proportional to the speed of the machine, so that it only becomes useful when a certain minimal speed is reached. This means in practice that back-EMF sensing can not be used from standstill and that another sensorless method is required for start-up. Alternatively, the machine could be started open-loop, by gradually increasing the voltage and frequency, allowing the rotor to synchronise to the field. Open-loop acceleration without knowledge of the initial position can cause oscillations or cause the machine to speed up in the reverse direction, which might be undesirable in some applications.

4.2.1.1 Using Zero-Crossings for the Commutation Instants

Consider an ideally commutating BLDC, with waveforms as shown in Figure 3.10. Assuming that phase A and B are conducting and phase C is not, we can write for the phase currents:

$$i_A = -i_B, i_C = 0$$  \hspace{1cm} (4.1)

Reconsider (3.4), the phase voltage in synchronous reference frame:

$$u^f_s = R_s \cdot i^f_s + \frac{d}{dt}\Psi^f_s + j\omega_s \Psi^f_s$$  \hspace{1cm} (4.2)

For phase C, this equation simplifies to:

$$V^f_C = \omega \Psi^f_C$$  \hspace{1cm} (4.3)

Or, the phase voltage in the unexcited phase is equal to the back-EMF. The zero-crossings of this signal are very useful for determining the commutation, because their relative position to the commutation instants is fixed and they occur precisely at rotor positions where the phase winding is not excited [Acaru and Watson, 2006]. However, the back-EMF zero-crossings do not correspond to those rotor positions where commutation should take place. Hence, the signals should be phase shifted by 30 electrical (el.) degrees before those moments can be used. Figure 4.2 shows intuitively how the zero-crossings of the back-EMF can be used to detect
CHAPTER 4. SENSORLESS CONTROL ALGORITHMS

Figure 4.2: Relation of current commutation to back EMF [Acarnley and Watson, 2006].

commutation instants for trapezoidal PM machines: This is by far the most popular method because the implementation is fairly straight-forward. In essence one simply compares voltages to obtain the commutation instants. The most obvious difficulty for this method is that the star-point of the machine must be available to measure the back-EMF. A second drawback of this method is that an accurate speed estimation is required to correctly implement the 30° phase shift in time and this at every operating speed.

4.2.1.2 Back-EMF Third Harmonic Integration

Moreira [1996] developed a position sensing method using the third harmonic of the back-EMF. It can be shown that, in case the resistances and inductances are three-phase symmetrical, the voltage between an added star-point $n$ and the actual star-point $s$ is equal to the mean of the three phase EMFs. Figure 4.3 shows the voltage $v_{ns}$ in its relationship to commutation instants. The frequency of $v_{ns}$ is three times that of the fundamental component of the back-EMFs, hence
the name “third harmonic” of the back-EMF, although it also contains harmonics of higher order. Because the zero-crossings of this waveform still lead the commutation instants by $30^\circ$ electrical, the waveform is shifted by integration. However, for every commutation cycle, the waveform passes three times through zero, such that some form of synchronisation must be applied in order not to commutate at a ‘false’ zero-crossing. An example of such a synchronisation could be the zero crossing of the fundamental of the back-EMF. Although at first sight the same integration has to be applied as with the simpler zero-crossings method, the integration is performed on a signal with a frequency three times higher than the fundamental. This improves the dynamic performance [Moreira, 1996]. Another advantage of this method is that current flow in the

![Figure 4.3: Schematic representation of the Third Harmonic Integration position-sensing method [Acarnley and Watson, 2006].](image)
unexcited phase does not influence the measurability of the third harmonic. It is even possible
to use the method with continuous current flow in the unexcited phase [Shen et al., 2004]. Note
that the initial assumption that the inductance is symmetrical for the three phases might not be
true, especially for machines with salient-pole machines such as machines with magnets buried
into the rotor. If this is the case, errors in the position estimation arise due to rapidly changing
phase currents [Acarnley and Watson, 2006]. A last remark is that this method relies on a
comparison between an extra star-point and the real star point of the machine. The internal
star point of the machine is often not available.

4.2.1.3 Integration of the Back-EMF

In this method, the commutation method is determined by integrating the back-EMF of the
non-energised phase when the back-EMF crosses zero. The idea behind this method is that the
integrated area of the back-EMFs shown in Figure 4.4 is approximately the same at all speeds.
A threshold is set to stop the integration that corresponds to a commutation instant [Tae-Hyung
Kim et al., 2005]. Field-weakening can be obtained by changing the threshold to a lower value,
advancing the commutation instants. This approach is less sensitive to high frequency noise and
automatically adjusts for speed changes, but at low speed the performance is poor due to error
accumulation [Tae-Hyung Kim et al., 2005]. Note that this method only provides a solution to
the 30° phase shift that is necessary for the back-EMF based methods, one still needs to obtains
the zero-crossings of the back-EMF.

4.2.1.4 Conduction of the Free-Wheeling Diodes

Although this method does not directly measure the back-EMF or its zero-crossings, it uses an
effect that occurs at the zero-crossings of the back-EMF. For a short period after reaching the
zero-crossing of the back-EMF in the non-energised phase, a small current will flow through the
diode when the back-EMF of the phase is smaller than $-V_D$, the voltage drop of the diode [Tae-Hyung
Kim et al., 2005]. This small current can be detected with comparator circuitry for each
diode. The main drawback of this method is that the circuitry to detect current flow through
each diode requires six additional power supplies. Again, since the back-EMF zero-crossings lead the commutation by $30^\circ$, a phase shift is needed to obtain the exact commutation instant.

### 4.2.2 Observer-Based Methods

![Principle of a closed-loop observer](https://example.com/figure4_5.png)

Observer principles have been applied to support the controller of PM machines: a mathematical model simulates the behaviour of the machine and is fed inputs and produces estimates of the outputs. These estimates are compared to the real output of the machine, in order to obtain an estimation error. The model parameters are updated using minimisation of the estimation error (see Figure 4.5). In a mathematical model, all states are accessible, including those which cannot be measured on a physical machine, which allows us to estimate unknowns, such as the rotor position. Observer-based methods all have the same limitation, namely that the estimates can only be as accurate as the model. The impact of modelling errors can be significant. Moreover, the mathematical model should not be too complex: if the calculation of the outputs takes too long, the model is useless for online use. Thus, the success of this method depends on the availability of an accurate, though not too complex model. Assumptions such as phase-symmetry may not be possible for certain machines. Accurate parameter measurements are not always possible. In [Takeshita et al., 2001] it appeared that for low speeds the position estimation error goes to zero, making it difficult to obtain an accurate position estimation. This problem could be solved by injecting voltage pulses at low speed, as proposed by the authors. This method is closely related to Inductance Variation, which is explained in the next section.

### 4.2.3 Methods Based on Inductance Variation

Methods based on inductance variation monitor the rates of change of winding current to estimate rotor position. The rate of change of current depends on the inductance of a winding,
and this is in turn a function of rotor-position and winding current. Hence, the rate of change of current in a winding contains information about the position. The main advantage of this method is that it can be used even at zero speed, since the absence of a motional EMF does not matter. Unfortunately this method is very difficult to implement in surface PM machines, since the inductance barely varies with rotor position. The rate of change of current is also mainly dominated by the back-EMF. An important drawback is illustrated in Fig. 4.6. The minimal value of the incremental inductance occurs at two rotor positions, both 0° and 180° electrical. This leads to an ambiguity in position sensing.

Figure 4.6: Flux linkages and incremental conductance as function of the position of the flux [Acarnley and Watson, 2006].

The inductance method addresses the problem of starting, by allowing to identify the initial rotor position [De Belie et al., 2010]. This is done by applying voltage pulses to the stationary machine, resulting in currents with amplitudes dependent on the incremental conductance. Note however that only salient or internal magnet type PM machines have a sufficiently large difference in inductances that are required to use this method effectively. With a smooth rotor, often used in high speed machines, inductances are almost equal for any rotor position. Voltage pulse injection at a high carrier frequency has no audible noise, but low amplitude current signals, whereas lower carrier frequency signals are more easily detectable but can introduce audible noise. A second problem arises in high speed applications. To obtain a frequent update on the rotor position estimation, the voltage pulses must have a frequency that is at least an order of magnitude larger than the fundamental frequency of the machine. For machines turning at high
speeds, it might be difficult to find an inverter that can switch at such high frequencies.

### 4.2.4 Methods Based on Flux Linkage Variation

The fundamental idea of the flux-linkage position sensing is based on the phase-voltage equation:

\[
v = R \cdot i + \frac{d\Psi}{dt}
\]  

(4.4)

which can be rearranged into

\[
\Psi = \int (v - R \cdot i) dt
\]  

(4.5)

Therefore, we can find the flux linkage by subtracting the resistive voltage drop from the phase voltage and integrating. This flux can then be used to calculate the generated torque, which is then in turn substituted into a mechanical equation to obtain the speed and position. One can see the Flux Linkage Variation method as a combination of the back-EMF method and the inductance method previously described. It does not present any more information than the combination of these two methods [Acarnley and Watson, 2006]. This method can be divided into two main approaches: with mechanical model and without a mechanical model. If a mechanical model is available it can be used to produce estimates of the speed and position:

\[
\frac{d\omega}{dt} = \frac{1}{J}(T - B\omega - T_L), \quad \frac{d\theta}{dt} = \omega
\]  

(4.6)

Of course, the model parameters and the load torque have to be known, which can be difficult for varying loads. To solve this problem, methods were developed that avoid using a mechanical equation. Figure 4.7 shows the basic principle of these methods: from the flux linkage and a stored flux linkage/postion/current characteristic both rotor position and current are estimated. The estimated current is fed back and compared to the actual current to generate an estimation error.

The drawbacks of using a mechanical equation are clear: to use use equation, the combined inertia of the machine and load (which can vary), the load torque and friction constant have to be known. For most applications, this is not an easy task and load characteristics will have to be studied case by case. Another drawback is that the integration to obtain the position from the speed can drift.

### 4.3 Motional Back-EMF in Detail

#### 4.3.1 Method of Preference for High Speed

Within the confines of this thesis, non-salient, surface permanent magnet machines combined with a PAM inverter are studied. The surface magnet machine is very suitable for high speed applications, because of its simple geometric form and easy construction, as explained in the introductory chapter. This excludes the use of sensorless methods that exploit the difference in saliency, such as Inductance Variation, since variation of the inductance with the rotor position is negligible. Furthermore, because of the high speeds at which the machine at hand
operates, complex algorithms such as Flux Linkage Variation or Observer-Based Methods might take too long to calculate on-line, especially for high speed machines where time frames are even smaller. Observer-based methods also require accurate machine models. Constructing models and finding the right parameters for these models can be very time consuming. Unfortunately, excluding these methods means that the estimation of starting position will be very difficult, because the remaining methods (Back-EMF-based, Flux Linkage Variation) are not possible at standstill. The methods using the Motional Back-EMF seem the most simple and robust, and without an upper bound concerning speed. This leaves us with the standard zero-crossing detection with 30° phase shift, integration of the third harmonic, integration of the back-EMF and conduction of the free-wheeling diodes. The latter will not be considered because measuring the conduction state of the free-wheeling diode is far more complicated than measuring the zero-crossing of the back-EMF by comparing voltages. Integration of the third harmonic requires the star-point of the machine, which is not available in most machines. Furthermore, this method also requires the standard zero-crossing method to determine which of the three zero-crossings of the third harmonic per cycle corresponds to the actual commutation instant. This method thus has no real advantages over the standard back-EMF method and is therefore also excluded. Integration of the back-EMF to determine the commutation instants also requires the zero-crossings and solely eliminates the need for the 30° phase shift. However, it requires some tuning to find the correct threshold and integrator drift might be an issue. The standard back-EMF method is therefore chosen as the method of preference because of its simplicity. In the remainder of this section, we highlight some practical difficulties that are inherent to this method, such as extracting the back-EMF, dealing with measurement noise, implementing a correct phase shift and dealing with the behaviour during commutation. A very useful variation on the standard method is given. This alternative technique, using line-to-line measurements as developed by Liu et al. [2016] alleviates some of the aforementioned problems.
4.3.2 Practical Difficulties

4.3.2.1 Unavailability of the Star-Point

As stated before one needs the star-point of the machine to extract the back-EMF. The star-point of the machine is often not accessible as most machines have only three terminals [Acarnley and Watson, 2006]. This can be solved by implementing a virtual start-point, which is made by connecting three resistors in wye between the phases.

4.3.2.2 Measurement Noise

Low-pass filters are used to remove high frequency components in the measurements. This is especially necessary in PWM-type inverters where the high switching frequency needs to be filtered out. Even more so when the star-point of the machine is used to obtain the back-EMF, because it contains a lot of common mode due to the switching of other phases Iizuka et al. [1985]. The main problem with low-pass filtering is that it inherently introduces a time delay. If not compensated for this delay, the commutation instants will not be correct.

4.3.2.3 Phase Shifting

Since the zero crossing points of the conventional back EMF method are inherently leading 30 electric degrees of the ideal commutation points, a precise velocity estimator and a phase shift circuit (algorithm, of digitally implemented) are needed to process the zero crossing signals so that accurate commutation points can be determined. In the original work of Iizuka et al. [1985] this phase shifting was implemented using RC-networks. Of course, using such fixed filters severely limit the speed range of the drive.

4.3.2.4 Behaviour During Commutation

The back-EMF is sensed in the phase that is not carrying current, because then the terminal voltage is exactly the back-EMF (4.3). However, in Section 3.6.1 we explained that during commutation, the free-wheeling diodes conduct for some time due to the inductive behaviour of the windings. If the conduction period becomes longer than 30\(^\circ\) the zero-crossings can no longer be detected. This problem is not easily solved. To detect zero-crossings, the phase current waveform has to be discontinuous, with no current flowing for a sufficiently long interval near zero-crossing of the back-EMF [Zhu et al., 1995]. This demands an appropriately designed motor where not only efficiency and flux density are design parameters, but also the inductances of the windings and by consequence the conduction angle of the diodes. In [Zhu et al., 1995] a complete design methodology for a high speed, 120 000 rpm motor is given. The authors show that whilst motors that have widely varying design parameters exhibit more or less the same performance, the diode conduction angle can vary by a factor of 2 or more. In Figure 4.8 the back-EMFs of two motors with similar efficiency are given. The motor with the back-EMF on the left has a relatively low stator flux density and a high stator core length-to-diameter ratio, using more iron and less copper, which results in fewer turns per phase, and consequently low winding inductances and a low diode conduction angle [Zhu et al., 1995]. Therefore, although
the motors have similar efficiency, only the left motor would be suitable for sensorless operation based on the detection of the zero-crossings of the back-EMF waveform.

Figure 4.8: Simulated back-EMFs of a motor with low winding inductance (left) and high winding induction (right). Based on [Zhu et al., 1995].

4.3.3 Line-to-Line Difference of Back-EMFs

In 2011, Kim et al. [2011b] presented a method that improves signal-to-noise ratio drastically while also eliminating the need for the classical 30° phase shift. Unlike the zero-crossing of the open phase, the zero-crossing of the line-to-line back-EMF difference corresponds exactly to the commutation instant. Indeed, the line-to-line voltages always lag over 30° with respect to the phase voltages, as illustrated in Figure 4.9.

From Figure 4.9b it is clear that the line-to-line back-EMFs have twice the amplitude as the phase back-EMFs. This allows to detect the back-EMF zero-crossings already at a much lower speed than with the conventional method. A phase shift is no longer required which simplifies the control. However, the commutation instants cannot be advanced with respect to zero-crossings of the back-EMF (unless one calculates them from the previous cycle). This means that letting the current vector lead the back-EMF to obtain field weakening is not possible. Important to note is that the coincidence of zero-crossings and commutation instants means that the zero-crossings of the line-to-line back-EMFs can be used to make emulated Hall-effect sensors. Thus, the output signals of the method can be directly applied to the conventional commutation table, as if they were obtained from the real Hall-effect sensors.
4.4 Conclusion

At the beginning of this chapter the need for sensorless control was established. Different sensorless techniques were explained and from these, the one using the motional back-EMF was chosen. This technique is fairly simple, does not require much additional circuitry and has no upper bound in rotational speed. However, the method requires the machine’s star-point (or a virtual one), needs signal filtering and an accurate phase shifting algorithm. These problems are alleviated with a variation on this technique based on the line-to-line back-EMFs. Concerning high speed, an important remark has to be made with regard to the conduction time of the free-wheeling diodes: if this duration is longer than 30° electrical, zero-crossings can no longer be discerned. This can only be solved by designing the motor appropriately. In the following chapter, we will discuss the test set-up on which we will implement a sensorless control algorithm.
CHAPTER 5

Description of the Test Set-Up
5.1 Introduction

Before developing the methodology of this thesis, we describe the test set-up on which the results of the remaining chapters were attained. Since the problems with high speed operation are mainly of a practical nature, it was important to implement the research on a physical set-up and not just simulate it. The test set-up was developed at Ghent University under supervision of prof. dr. ir. P. Sergeant. It consists of a rectifier, PAM inverter with a variable DC link, voltage and current measurement devices and control electronics. The use of a PAM inverter is justified by the considerations in Section 2.2.2. Jasper Beurms built the hardware of the set-up and conducted some basic tests [Beurms and Sergeant, 2014]. The hardware of the set-up was completed by Parmentier and Sergeant [2016] and an open-loop and closed-loop control using hall sensors for a low speed (3000 rpm) machine was developed in LabVIEW (cf. infra). The LabVIEW communication architecture used by Beurms and Sergeant [2014] as well as some components on the set-up were made by De Smaele et al. [2015] in their work for a didactic set-up using Space Vector Modulation. In this chapter an overview and some important details are given of both the hardware and software of the set-up.

5.2 Hardware

5.2.1 Overview

In Figure 5.1, an overview of the power electronics layout is given. From left to right, we have the 3-phase grid (230V), connected to a rectifier. The rectified voltage (≈ 310V) is fed to a buck converter, that lowers the voltage. The voltage is then filtered using a large capacitor. The current is then chopped up in the for BLDC control required 120° blocks by the inverter, which is connected to the motor. This simplified overview will be elaborated in the following sections. In Section 5.2.2 we introduce the two different motors that were connected to the set-up. The power electronics are explained in more detail in Section 5.2.3, while the control electronics that measure, process and send signals to the power electronics are discussed in Section 5.2.4. The last section is dedicated to the software that was used.

5.2.2 Motors

Tests were conducted on two motors, a Torcman TM685-40 hobby motor with outer rotor and a custom designed high-speed motor. The table below contains the nameplates of the two machines.

The high-speed machine was designed by prof. dr. ir. P. Sergeant. It has quite an unusual configuration, consisting of two machines, a motor and a generator mounted side-by-side on the same axle in one casing. The generator is connected to a rectifier and a resistor bank, allowing to easily load the motor with a variable load. One side is configured in wye with the standard three terminals. The other side has twelve terminals, which are the cable heads of six internal coils, that have to be connected in three pairs. This allows two things: firstly, the star-point is now externally accessible for measurements, a luxury that is not available in most machines. Secondly, one could connect three instead of six coils that form the stator windings, halving...
Figure 5.1: High level overview of the power electronics, consisting of a 3-phase rectifier (green), DC/DC buck-converter (red) and inverter (blue).

the magnetic flux density and thus resulting in a twice the nominal speed for the same nominal voltage. The axle of the machine extends outside the machine on both ends. On one side this allows for an encoder to be attached. The other side has a small external rotor with permanent magnets that allows the use of hall-effect sensors externally. In Figure 5.2, the lay-out of the motor configuration is sketched.

5.2.3 Power Electronics in Detail

Now that we have a general idea of the overall lay-out of the set-up, we review each of the power electronic components more thoroughly.

Rectifier, Inverter and Buck Converter Modules

The rectifier, buck converter and inverter are all integrated in two insulated-gate bipolar transistor (IGBT) modules from Semikron®, the SKiiP 14NAB066V1 (hereafter MiniSkiip). The modules contain six diodes for the three-phase rectifier, a diode and IGBT for use as a brake

<table>
<thead>
<tr>
<th>Torcman TM685-40</th>
<th>High-Speed Motor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Quantity</td>
<td>Value</td>
</tr>
<tr>
<td>Nominal Voltage</td>
<td>60 V</td>
</tr>
<tr>
<td>Nominal Speed</td>
<td>3000 rpm</td>
</tr>
<tr>
<td>Nominal Power</td>
<td>4.5 kW</td>
</tr>
<tr>
<td>Nominal Current</td>
<td>80 A</td>
</tr>
<tr>
<td>Pole Pairs</td>
<td>7</td>
</tr>
<tr>
<td>Nominal Frequency</td>
<td>350 Hz</td>
</tr>
</tbody>
</table>

Table 5.1: Nameplates of Torcman TM685-40 motor and custom high speed motor.
chopper and six IGBTs with free-wheeling diodes as inverter. The modules are mounted on test-boards, which allow easy access to all connection terminals shown in Figure 5.3a. However, due to the wiring of the test-boards, RECT+, B+ and DC+ are all connected and so are B− and DC− [Semikron International GmbH, 2014b]. This wiring prevents us from using only one module as rectifier, buck converter and inverter. Indeed, if we were to use the brake chopper IGBT as switch for the buck converter, connecting B to RECT+, we would have to connect B− to the DC+ side of the inverter. This is impossible since B− is already wired to DC−. Therefore, two modules are combined to acquire the correct topology. Figure 5.4 displays this configuration. The hard-wired connections of the test-boards are drawn as green lines. The blue lines represent externally made connections. In the red dashed rectangle we can identify the standard buck converter topology (compare with Figure 5.1). Initially, the brake chopper IGBT was used as the buck converter power switch, but any IGBT on the bottom of the inverter legs can be used.

Drivers

The digital logic signals generated by the control electronics (cf. infra) cannot be directly fed to the IGBT gates. Instead, the logic signals are passed to IGBT drivers that provide the correct voltages to the IGBT gates. The driver used for each pair of IGBTs is a SKYPER 32 R from Semikron®, mounted on a SKYPER 32 R adapter board. By soldering components to the adapter board, the gate charge of the IGBTs can be changed and the desaturation protection can
be calibrated. We refer to Beurms and Sergeant [2014] for the calculations of the gate resistors. The drivers can do some basic fault detection, such as interlock prevention and desaturation detection. Interlock prevention prohibits that both bottom and top IGBT of an inverter leg can be switched on at the same time which would short-circuit the supply. The desaturation protection (also called Dynamic Short Circuit Protection (DSCP)) monitors the collector-emitter voltage drop $V_{CE}$ across the IGBTs. A high $V_{CE}$ indicates that the IGBT is no longer in the saturated region and thus drawing a very high collector current. This is probably a sign of a short circuit or severe overload [Van den Bossche, 2017]. The DSCP compares $V_{CE}$ to a reference voltage $V_{CEref}$. $V_{CE}$ has a high transient right after switching on the IGBT, which should be ignored by the DSCP. For this reason, $V_{CEref}$ has an exponential shape, as shown in Figure 5.5. The steady-state value of $V_{CEref}$ is set to its maximum value by soldering an external resistor $R_{CE}$ to the adapter board. The blanking time ($t_{bl}$) is controlled by a capacitor $C_{CE}$. Values for

Figure 5.4: Configuration and wired connections of the two MiniSkiip modules.
the resistor and capacitor were previously found by Parmentier and Sergeant [2016], by simply increasing the current. Unfortunately, these values were set for much smaller peak currents. The resistors and capacitors were replaced, allowing for a peak current up to 25 A.

![Figure 5.5: Illustration of the exponential shape of \( V_{CE_{ref}} \), in relationship to \( V_{CE} \) (based on [Semikron International GmbH, 2014a]).](image)

### 5.2.4 Control Electronics

**Overview**

A high-level overview of the set-up is given in Figure 5.6. The power electronics are depicted as squares, with from left to right the rectifier, buck converter and inverter. On the left of the inverter a current measurement module measures two phases going to the motor and the DC-link current. In the dotted rectangle a voltage divider in delta-configuration lowers the line-to-line voltages. A Hall-sensor arrangement measures the position of the motor rotor. Red lines indicate information going to the controller, while green lines indicate signal flow from the controller. The signals from the controller are 3.3V logic level signals that are first shifted to 15V by level shifters. The 15V signals are then fed to the drivers that steer the IGBT modules. The controller is connected to a Windows machine via an Ethernet cable.

**CompactRIO**

The power electronics of course, need to be controlled and receive timely signals. The test-up is built around an embedded controller from National Instruments (NI), the CompactRIO™(cRio). The controller features a real-time operating system (RTOS), a reconfigurable FPGA and an Ethernet port. Input-output modules with a variety of ports can be inserted in a reconfigurable chassis. In this set-up two I/O modules are used: the NI 9223 and the NI 9402, depicted in Figure 5.7.
Figure 5.6: High-level overview of the control electronics lay-out.
CHAPTER 5. DESCRIPTION OF THE TEST SET-UP

NI 9223 To measure the Hall-effect sensor signals (cf. infra), line-to-line voltages and voltages from the current measurement module, three NI 9223 C Series modules are used. The NI 9223 is a 4 channel, high-speed analog measurement module that can sample at a rate of up to 1 MS/s on each channel [National Instruments, 2016a]. The input voltage range is $\pm 10$ V. The module houses a simultaneous 16-bit analog-to-digital converter, this corresponds to a resolution of around 0.3 mV. The extremely high sampling rate of this module can be considered overkill for this particular application, since the sampling rate is a thousand times higher than the fundamental frequency of the high-speed motor. Sampling at 100 times the fundamental frequency would probably suffice. Nevertheless, the set-up was over-dimensioned to allow some liberty in the research.

NI 9402 The pulses that control the IGBT drivers are generated by two NI 9402 modules. These modules have a maximum switching frequency of 16 MHz [National Instruments, 2016b]. Again, this is higher than necessary, since the SKYPER drivers do not allow switching pulses shorter than 1 $\mu$s. The control electronics were overdimensioned to also allow to implement field oriented control using PWM.

Hall Sensors

A Hall-effect sensor arrangement is attached to both motors. On the Torcman motor a printed circuit board (PCB) with three Hall-effects sensors that are displaced by $17^\circ$ (corresponding
CHAPTER 5. DESCRIPTION OF THE TEST SET-UP

to 120° el.) is mounted on the outside of the machine, because the Torcman is an outer-rotor type machine. On the high-speed machine, the PCB is mounted below the small external rotor outside the casing. The PCBs for both machines are fastened on holding plates with the use of arced slots, allowing to move the position of the sensors with respect to the rotor. Correct alignment of the sensor arrangement is of great importance, as we learned in Chapter 3. In order to obtain MTPA, the current blocks should coincide with the flat part of the trapezoidal back-EMF. Alignment can be done by driving the motor externally, or even rotating it manually and measuring both Hall-effect sensor signals and line-to-line back-EMFs. The PCB is moved until the flanks of the Hall-effect sensor signals are perfectly aligned with the zero-crossings of the line-to-line back-EMFs. Figure 5.8 shows the lay-out of the PCB with the Hall-effect sensors displaced over 120° el.

Figure 5.8: PCB with Hall-effect sensors.

Current Measurement Module

At Ghent University, a costum PCB was design by De Smaele et al. [2015], containing three current transducers from LEM. These current transducers are connected in parallel to resistors to convert the currents into voltage signals that are within the measurement range of the NI 9223 modules.

Level Shifters and Signal Distribution

The level shifters and signal distribution (on Figure 5.1 shown as ‘level shifter’) designed respectively by De Smaele et al. [2015] and Parmentier and Sergeant [2016] transfer the 3.3 logic output of the NI 9402 to the 15 V switching level of the IGBT drivers and ensure the correct pin outputs are matched.

Voltage Divider

A simple voltage divider is used to lower the line-to-line voltages of the machine to the input range of the NI 9223 modules. The resistors of the voltage divider should be high in order to
CHAPTER 5. DESCRIPTION OF THE TEST SET-UP

not draw a lot of current. More importantly, the ratio of the resistor values should be as such that the lowered voltages are in the measurement range for the NI 9223 modules, but still as high as possible in order to improve the resolution of the readings.

5.3 Software

5.3.1 NI LabVIEW

All of the software in this thesis is written in the NI LabVIEW 2016 development environment using NI’s visual programming language ‘G’. The code is compiled on a Windows host PC. After the compilation process, parts of the code are ran on one of these three: the host PC, the RTOS of the cRio or the FGPA integrated in the cRio, depending on the requirements of that specific part of the code. For example, the host PC is used to store data and monitor certain key parameters. The RTOS offers reliable, predictive behaviour for tasks with specific timing requirements. The FPGA excels at small parallel tasks that require high-speed logic and precise timing. Some communication architecture is also needed to guarantee correct data transfer between FPGA, RTOS and host.

In order to understand the next chapters of this thesis, a good understanding of the LabVIEW environment is required. National Instruments offers self-paced trainings on their website that cover basic LabView programming, making real-time applications and FPGA programming fundamentals. Since most of the code written for execution on the FPGA is optimised with regard to throughput, timing and numerical precision, it is advised to also read the NI LabVIEW High-Performance FPGA Developer’s Guide to understand some of the considerations made. The NI LabVIEW for CompactRIO Developer’s Guide explains the best practices for network communication between the cRio and host PC. An exhaustive communication architecture was designed by De Smaele et al. [2015] that follows this design pattern. Their thesis also contains a useful section on fixed-point math (cf. infra). In the remainder of this section, some more advanced programming concepts for LabVIEW FPGA are explained that will be used in the upcoming chapters. Although the reader might not be interested in programming in the LabVIEW environment, the underlying principles of the concepts presented here are generally applicable.

5.3.2 LabVIEW FPGA

In Chapter 2 was explained how FPGAs excel at performing parallel tasks at high speed due to their inherent parallelism that comes from synthesising code down to physical hardware. However, there are some conceptual differences and best practices to keep in mind when programming in parallel rather than sequentially. Firstly, the main requirement for a high speed application is throughput or ‘number of samples processed per second’. An example related the high-speed BLDC control might clarify this: Consider the high speed machine presented in Section 5.2.2. Its nominal fundamental frequency is 1 kHz. If one were to sample the back-EMF of this machine to obtain, for example, the zero-crossings of the back-EMF, a sampling rate of at least 100 kHz would be required to have an adequate resolution. Thus, calculations involving these samples should not take longer than 10 μs in order not to miss the next sampling instant. This places high requirements on the throughput. The next section explains how throughput
in LabVIEW FPGA can be improved. The subsequent section explains some techniques to obtain correct timing and synchronisation of loops programmed on an FPGA. We will find that optimisations in throughput can make it harder to control timing.

5.3.2.1 Optimising throughput

Throughput is the result of three factors:

1. The clock rate that is driving the application (cycles/second).
2. The number of samples that are processed each clock tick (samples/call).
3. The number of cycles that must elapse before the algorithm can be called again, referred to as the ‘initiation interval’ (cycles/call).

This can all be combined in the following definition:

\[
\text{Throughput} = \frac{\text{Clockrate} \times \text{SamplesPerCall}}{\text{InitiationInterval}}
\] (5.1)

Improving the clock-rate is the most straightforward way to improve throughput. The maximal clock rate of the FPGA is 40 MHz. The number of data samples per call is a fixed characteristic of the design of the code. However, algorithms can often be split in parts that can run in parallel, allowing that different parts execute within the same clock cycle. The initiation interval concerns code blocks that take multiple cycles to execute per input.

An example is given to illustrate how the throughput of a loop in LabVIEW FPGA can be improved using critical path reduction, pipelining and user controlled I/O. Consider the LabVIEW snippet of Figure 5.9, which is a typical signal processing loop. An I/O-module is read and samples are passed through a low-pass filter to remove noise. Some operations are applied to the signal that in turn switch a boolean indicator on or off.

A first step to improve the throughput is to identify the critical path. The critical path is the data path with the longest propagation delay. In this example, there is only one path and therefore it is critical. When the length of the critical path can be reduced, then the overall propagation delay is reduced which will improve throughput. In this example, the low-pass filter
CHAPTER 5. DESCRIPTION OF THE TEST SET-UP

Figure 5.10: A typical signal processing application with pipelining to improve throughput.

The block that takes the longest to execute. This block could be replaced by a simple average of the last $n$ samples to reduce the critical path length.

A second technique for increasing code throughput is pipelining. It involves inserting shift registers along the critical path to break it into shorter concurrent sections. In our example, we could break the path before and after filtering. By doing this, a new sample can already be read while a previous one is being filtered and the one before that is sent through the final operations, as in Figure 5.10. Note however that although pipelining improves overall throughput, the time for a single sample to complete the complete data path may actually increase because of the additional logic delays of the shift registers.

Lastly, User Controlled I/O Sampling can acquire data faster from I/O modules than standard I/O nodes. In a typical I/O Node, operations happen serially: when a Node is called, the node tells the module to acquire data and then transfers the data from the module. The I/O Node cannot be called again until both operations are completed, so the acquisition rate is determined by the time it takes the module to acquire the data and transfer the data back to the FPGA. With User-Controlled I/O Sampling, operations can happen in parallel. The Clock node controls when the module acquires data and the Read Node controls when data is transferred. Since these are separate operations the Clock Node can run while the data is transferring from the module. The sample rate is now determined by the time it takes for the module to have the data ready for transfer, which is 1 MS/s in case of the NI 9223.

5.3.2.2 Timing control

It can sometimes be desired that operations run in a predetermined sequence. LabVIEW FPGA has so-called Flat Sequence Structures (the film reel-like structure in Figure 5.12) that consist of frames. The frames execute sequentially from left to right. If synchronisation is required between two loops that do not exchange data directly, you can use occurrences. An occurrence allows you to block one loop using the Wait for Occurrence function until another loop has completed a specific operation and calls the Set Occurrence function, as is shown in the diagram in Figure 5.12. Information can be exchanged between concurrent loops using local variables. Local variables are stored in flip-flops on the FPGA. Variables are not buffered, so they are only
Figure 5.11: A typical signal processing application with pipelining and User Controlled I/O to improve throughput.

useful when only the latest written value is needed. Only a single writer is allowed (to prevent race-conditions) but multiple readers are allowed.

Figure 5.12: Synchronisation between two loops using an Occurrence.

5.4 Conclusion

In this chapter, we established an overview of the hardware of the set-up. Special attention was given to the motors that are used, the important components of the power electronics and all of the control electronics. The schematic overview of the control electronics lay-out in Figure 5.6 might be useful to revisit in the coming chapters. Manuals and theses that are required to fully
understand the software in this thesis were referred to in the section on software. Subsequently, we explained more advanced concepts with regard to writing software for (LabVIEW) FPGA, such as throughput optimisation and timing control. These concepts will prove necessary in the next chapters.
CHAPTER 6

Methodology
In this chapter, we briefly underpin the decisions made within the research of this thesis with justifications from the literature and findings during the experiments. Due to the experimental and exploring nature of this thesis, most decisions were made based on problems that were encountered during the research.

6.1 Familiarisation with the Set-Up

The first main part of the research consisted of familiarising with the existing hardware and software of the set-up. A closed-loop speed control with Hall-effect sensors made by Parmentier and Sergeant [2016] was already present, although due to some changes made to the set-up, the speed control did not work. A step-by-step approach was chosen to debug the set-up and to understand the various parts. Firstly, the grid rectifier was replaced by a variable DC-source to ensure safety during the tests. Secondly, some changes were made to the buck converter and its correct working was verified. Thirdly, the inverter was tested and the motor phases were identified with the driver signals, level shifter signals and the ports on the cRIO 9402 module. A commutation state-machine was then programmed on the FPGA. This allowed for the Torcman motor to be driven open-loop, by simply increasing the frequency and voltage. A commutation table based on the Hall-sensor signals (as presented in Chapter 3) was then constructed and implemented on the FPGA. The speed control program of Parmentier and Sergeant [2016] was then slightly adapted and also implemented on the FPGA. The PI-controllers of the current and speed feedback were tuned until the dynamic response was satisfactory.

6.2 Choosing an Appropriate Sensorless Technique

It became clear in Chapter 4 that sensorless techniques based on the back-EMF are most suitable for high speed machines due to their simple and robust implementation. Instead of using the standard technique with 30° phase shift, an alternative technique based on the line-to-line voltages was chosen, because it eliminates the need for an accessible star-point (or the construction of a virtual one), the phase shifting and because it gives a better signal-to-noise ratio. The only drawback of this technique is that is does not allow field weakening because the commutation instants cannot be advanced. However, this was not considered an issue since the machines under consideration operate nominally at high speed, and do not attain high speed by field weakening.

6.3 Development on a Low Speed Machine

A first sensorless control was implemented on a the Torcman motor (see nameplate in Table 5.1). This motor is very suitable for tests with regard to high speed: although the nominal speed is ten times smaller than the nominal speed of the high speed motor designed at the University of Ghent, the nominal electrical frequency is less then 3 times smaller than that of the high speed motor, due to the large number of pole pairs. Therefore, the motor gives quite a good indication whether or not acquisition rates are sufficiently high and calculation times on the FPGA are short enough to use the sensorless algorithm on a high speed machine. A few first attempts were
made towards sensorless commutation. Although these were not successful, they are important to consider as they gave valuable insight in the requirements of a digital sensorless algorithm. The solution to most of the problems encountered in the first attempts was to use a detection interval, such that only parts of the back-EMF are searched for zero-crossings. An algorithm was implemented to detect the type of zero-crossing (positive to negative or negative to positive) to set the emulated hall signals either 'high' or 'low'. This allowed sensorless commutation to be successfully implemented.

6.4 Advanced Development on a High Speed Machine

As soon as the sensorless control worked on the low speed machine, the high speed machine was connected to the set-up. It became clear that the sensorless algorithm designed for the low-speed machine needed some adaptations in order to work on the high-speed machine. The cause of failing of the algorithm were the long diode conduction periods, which were predicted in Section 3.6.1. We first looked into ways to shorten this conduction period. A first method was to synchronise the DC-link voltage with the commutation instants of the inverter; by temporarily increasing the DC-link voltage, one could possibly counter the free-wheeling current. Unfortunately, it became clear that the response time of the DC-link is an order of magnitude too slow to have a significant impact. A second method was to test the effect of delaying or advancing the commutation instants. Under controlled conditions in simulation we showed that although delaying the commutation instants increased the diode conduction time, advancing the commutation instants did not significantly shorten the conduction time. The detection interval was then adapted to deal with the long conduction time. Firstly, using a delay to ignore the conduction period of the diodes and secondly, by measuring where the current is zero. It became clear that the first method requires some tuning to get it to work properly at different speeds. The main drawback of this method is that if the conduction period of the diodes were to become longer then the fixed delay, due to increased load or speed, the algorithm would fail to detect the zero-crossing. This could be solved by measuring how long the current is close to zero. If this duration falls below a safety threshold, the speed or load should be lowered. The algorithm appears to be robust with regard to missed crossings, meaning that algorithm does not fail to detect subsequent crossings when a crossing is missed. The second algorithm adapts the detection interval automatically by establishing the detection interval as the period where the current is zero. However, since the current waveform is a direct consequence of the commutation instants, a missed zero-crossing will result in failing to detect all subsequent crossings. A more robust sign detection was designed because the simple sign detection used on the low machine was not completely accurate. This completed the zero-crossing detection, allowing to run the high-speed machine sensorless at different speeds.
CHAPTER 7

Development on a Low Speed Machine
CHAPTER 7. DEVELOPMENT ON A LOW SPEED MACHINE

7.1 First Look at the back-EMF Waveform

In order to find a good first approach to the zero-crossing detection of the line-to-line back-EMF of the low-speed Torcman motor, a look is taken at the waveform of the back-EMF. The waveform was captured on a DLM2000 Yokogawa oscilloscope using differential probes and is depicted in Figure 7.1. In the following chapters, every oscilloscope capture will be indicated by (DLM2000).

Figure 7.1: Waveform of the line-to-line voltages at 2300 rpm (DLM2000 capture).

At first glance we notice distinctive peaks at every commutation instant, except where the back-EMF reaches its extremities. These peaks, which we will refer to as ‘commutation spikes’ in this chapter, are actually very short conduction periods of the free-wheeling diodes. We also see that the Hall-effect sensors are not perfectly aligned in this case. The commutation (indicated by the arrow) happens before the back-EMF reaches zero, which means the Hall-effect sensors are a few degrees advanced.

7.2 Exploratory Tests

7.2.1 First Zero-Crossing Detection

Although the first few attempts at sensorless commutation were not successful, some important insights were gained from them. In this section, we explain some of these attempts to understand what is needed for a robust commutation algorithm.

LabVIEW FPGA’s built-in zero-crossing detection (ZCD) block was used as a first test. This block can detect zero crossings in either direction, crossings that go from negative to positive
and vice versa. To capture the boolean signals of this block, each time a zero-crossing was detected the NI 9402 module sent a 5$\mu$s pulse to the oscilloscope. In Figure 7.2 we see that the block detects the spike just after commutation, and then multiple zero-crossings near zero.

![Figure 7.2: Zero-crossing detection pulses, with zoomed-in view in inset (DLM2000 capture).](image)

To reduce the number of pulses, a low-pass filter with a cut-off frequency of 20 kHz was applied to the signal. The settings of the zero-crossing block were then changed to only detect negative to positive crossings. This lowered the number of detected crossings, but did not eliminate that some zero-crossings were detected at completely wrong instants due to measurement noise.

It was then decided that after a first crossing is detected, a delay is to be inserted such that no other crossings could be detected. This allowed to make a relatively stable emulated Hall-sensor signal. However, when this Hall-sensor was used, it appeared that on the falling side of the line-to-line voltage, zero-crossings were not well detected. From Figure 7.4 we can deduce why this happened: the line-to-line voltages had a very small positive DC-component. Since the algorithm used exactly zero as threshold value for a crossing, it failed to detect that the back-EMF is near zero. Only due to a measurement error the line-to-line voltage would appear to pass through zero. This was solved making a custom ZCD block that used $\pm0.01$ as threshold and not zero.
7.2.2 Rejection of Commutation Spikes

As the research progressed, the commutation spikes became more pronounced due to shifting of the Hall-effect sensors, caused by mechanical vibrations. The commutation spikes often passed...
through zero, causing a false commutation signal. In Figure 7.5 every detected zero-crossing is indicated by a red star. It is clear that the algorithm needs some adaptation to reject these false crossings.

![Figure 7.5: Erroneous detection of the commutation spikes as zero-crossings (DLM2000).](image)

To reject the commutation spikes that did not correspond to a zero-crossing of the line-to-line voltage, an interval was created around these instants wherein no commutation was allowed. The commutation instants were programmatically delayed to create this interval, as illustrated in Figure 7.6.

![Figure 7.6: Rejection interval around commutation instants.](image)

This was implemented in LabVIEW FPGA using Occurrences. In the commutation table, an Occurrence was set when a commutation signal was received from the Hall-sensors (or later a ZCD algorithm). The Occurrence set a boolean indicator to 'high' that blocked detection. The commutation signals were then delayed for a few microseconds. After several tens of microseconds, the blocking boolean indicator was set to 'low' again, allowing detection of the crossings again. During the testing of this new algorithm, it became clear that only minor improvement was gained. The commutation delay had to be significantly large to reject the commutation spikes completely, which meant the machine was no longer operating optimally. The Occurrences also made it more difficult to keep track of the timing of the different loops.
CHAPTER 7. DEVELOPMENT ON A LOW SPEED MACHINE

7.2.3 Requirements for a ZCD Algorithm

From the above experiments we can learn what the requirements are for a robust ZCD algorithm should be. First, the algorithm should not be triggered by crossing exactly zero, instead there should be a small ‘band’ around zero that triggers the ZCD. Second, due to the different loops all running at different rates, it is very difficult to keep track of the timing of events, especially when using Occurrences. A fixed sample period should be set and the various loops in the algorithms should be synchronised with this sample rate. Third, the commutation spikes should be eliminated completely. Efforts to mask the commutation spikes locally, as in Section 7.2.2 were not effective enough to completely eliminate them. Furthermore, some masking methods of the commutation spikes led to deterioration of the detection of the actual zero-crossings. Fourth, as filtering gave only minor improvements, but introduced significant delays, filtering should only be used as a last resort.
7.3 Sensorless Commutation Algorithm using Detection Intervals

7.3.1 User-Controlled I/O and Sample-and-Hold

To make the timing of the different loops consistent, each loop is given the same loop timer. The I/O Nodes are replaced by User Controlled I/O (see Section 5.3.2.1) to sample at the same fixed rate as the loop timer. A custom sample-and-hold algorithm is made because User Controlled I/O returns zero when one tries to access samples at a time that doesn’t perfectly coincide with the moment a sample is produced. Figure 7.8 illustrates how the Sample-and-Hold is implemented in LabVIEW FPGA: each time the Read I/O Method produces zero, the previous sample is kept. When a value different from zero is read, this value overwrites the previous value. Note that this algorithm ignores the fact that the signal can actually be precisely zero. However, since the resolution of the NI 9223 module is 0.3 mV, the number of times the signal would be exactly zero is small, so this is not considered an issue.

![LabVIEW FPGA sample-and-hold implementation.](image)

7.3.2 Creating a Detection Interval

Consider again Figure 7.1. One can see that the commutation spikes can be easily rejected if measurements are only made in an interval where the line-to-line voltage is close to zero for an extended time. This allows to optimise the ZCD within this interval, without having to consider what effect the optimisations would have on detecting ‘false’ crossings outside this interval. Concretely, this means that zero-crossing detection only starts when the line-to-line voltage is within a certain interval, for example between $\pm 10\%$ of its maximum value. The interval is constructed using a debounce filter. A debounce filter processes the raw sampled data and only pass through a change in state whenever the sampled data has remained constant for a defined period of time. Any pulses shorter than the filter time constant are removed and not passed through. In LabVIEW FPGA it is straightforward to implement such a debouncing algorithm. The diagram shown in Figure 7.9 compares the new input signal to the current output signal and looks for a change in state whenever the sampled data has remained constant for a defined period of time. When the counter exceeds the limit (‘Cycles to Filter’ in the diagram) it
passes the new input value to the output and resets the counter to 0. Note that the loop in Figure 7.9 is timed with the same sample period as the sample-and-hold loop. This means that the time constant of the debounce filter is a multiple of the sampling period, making it easy to interpret. For example, if the sampling frequency is set to 1 MS/s, the 'Cycles to Filter' constant corresponds to microseconds.

![Figure 7.9: Debounce filter implementation in LabVIEW FPGA.](image)

If the time constant of the debounce filter is set correctly, it will reject the short instants where the line-to-line voltage is within the interval due to the commutation spikes, but will accept the periods where the line-to-line voltage is within the interval for prolonged periods. Figure 7.10 illustrates this. Waveform (a) is the line-to-line voltage with exaggerated commutation spikes. (b) is a boolean value that is 'high' where (a) is within the indicated interval. A debounce filter is then applied with the time constant shown in red, resulting in the debounced signal (c).

![Figure 7.10: Illustration of the debounce filter applied to a threshold on the line-to-line voltages.](image)

The detection interval made with this debounce filter is very stable. In Figure 7.11 an oscilloscope still is shown with the line-to-line voltage and the detection interval, which was generated by the NI 9223 to test it.
7.3.3 Emulated Hall Signal

The detection intervals allow to easily detect single zero-crossings. However, whether a crossing goes from negative to positive or positive to negative was still to be determined, to uniquely define the emulated Hall signals. This is done by measuring whether the line-to-line voltage is positive or negative at the beginning of the sensing interval. For example, if the emulated Hall signals are supposed to be 'high' when the line-to-line voltage is positive, as shown in Figure 7.12, the Hall signal should be high at the beginning of a positive-to-negative zero-crossing. To enforce this, the emulated Hall signal is explicitly set to high at the beginning of the detection interval. This changes nothing if the Hall signal was already correct, but corrects the Hall signal if it were 'low' due to a missed zero-crossing.

Within the detection interval, the emulated Hall-signal is simply changed with a 'NOT'-gate when the line-to-line voltage is within a very small band. Subsequently, further detections are blocked with a delay. The delay should be long enough for the line-to-line voltage to be far from
zero and preferably outside of the detection interval, to prevent accidental ZCD. Figure 7.13 shows how this algorithm was implemented in LabVIEW FPGA.

![Image of LabVIEW FPGA implementation](image_url)

Figure 7.13: Emulated Hall-effect signal implementation in LabVIEW FPGA.

### 7.4 Conclusion

After some initial attempts to detect zero-crossings and to reject unwanted noise and commutation spikes, the key requirements for a ZCD algorithm were identified:

- Consist and controllable timing
- Measuring only in specific intervals to reject peaks and noise
- A 'band' around zero that triggers the ZCD

The first requirement is met by using User-Controlled I/O, combined with a sample-and-hold algorithm. The interval where zero-crossings are detected is determined by a **debounce filter**. At the beginning of this interval, the sign of the back-EMF is measured to determine the type of zero-crossing. An emulated Hall-sensor signal is then constructed, using the measured sign of the back-EMF and by detecting zero-crossings within a small band around zero. No further tests under load conditions were performed because problems encountered during the development on the low-speed machine might not have been relevant for the development of a high-speed sensorless algorithm.
CHAPTER 8

Extended Development on a High Speed Machine
8.1 Back-EMF with Long Diode Conduction Period

In Section 3.6.1 we explained the behaviour of BLDCs during commutation: The non-energised phase keeps conducting for a while after commutation through the free-wheeling diodes, due to the inductive behaviour of the windings. The diode conduction period (DCP) is influenced by the winding resistance, the back-emf, the winding inductance, the load and the electrical pulsation. It is thus to be expected that the conduction period can be quite long for high-speed machines, because of the high electrical pulsation. As we saw in Section 4.3.2.4, if the DCP becomes longer than $30^\circ$ electrical, zero-crossings of the back-EMF can no longer be detected.

Consider the two oscilloscope captures of line-to-line voltages and phase currents of the high-speed machine developed at Ghent University given in Figure 8.1. We see that under no-load conditions (8.1a), the conduction period is already longer than for the Torcman motor from the previous chapter, with around 12 degrees conduction after commutation. Under partial load, in (8.1b), we see that the conduction period of the free-wheeling diodes is already far longer than $30^\circ$ electrical, making it impossible to detect zero-crossings.

8.2 Shortening the Diode Conduction Period

In order to extend the range where the machine can be operated sensorless, we would like to be able to shorten the conduction period of the diodes. Unfortunately, this proves to be rather difficult, as will be shown in the following two sections.

8.2.1 Synchronisation between Inverter and DC-DC Converter

A first idea to shorten the conduction period is to increase the DC voltage right after commutation. How this shortens the conduction period can be understood from Figure 8.2. The commutation is the same as in Figure 3.12, U is energised, while the commutation changed the
energised phase from V to W. Due to the remaining current in phase V, the phase is temporarily connected with positive DC-bus \((DC^+\)) through the top free-wheeling diode. We assume that the star-point of the machine can be approximated by \(\frac{2}{3}DC^+\), if \(DC^-\) is the reference. The voltage over the phase (red arrow) then counters the current (blue arrow) and forces it to exponentially diminish to zero. The only way we can improve the rate with which the current reduces to zero is by imposing a higher \(DC^+\) voltage.

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The DC-bus voltage can be altered by changing the duty ratio of the buck-converter. Unfortunately, the buck-converter has a rather slow response time due to the large inductor that it uses to limit the current ripple. A simulation of the buck-converter was made in Simulink®, using values from the actual test set-up for components: the inductor of the buck converter has an inductance of 3.88 mH, the motor is modelled as an RL-load with \(R = 1\Omega\) and \(L = 0.7\) mH, the capacitor over the output voltage has a capacitance of 660 \(\mu\)F, the DC-voltage source is 200V and the switching frequency is 10kHz. At time step 5ms, the duty cycle is suddenly increased from 25% to 95%. We see in Figure 8.3 that the output voltage has a typical second order response with a settling time of around 10ms. This is too slow to have any significant influence on the \(DC^+\) voltage during commutation for the high-speed machine, since the fundamental period at a speed of, for example 20000 rpm (± 1.5ms), is an order of magnitude smaller.

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### 8.2.2 Advancing or Delaying Commutation Instants

A second option could be to advance (i.e. field-weakening) or delay the commutation instants with respect to the back-EMF zero-crossings. At first glance, it is not clear whether or not this will shorten the DCP. Therefore a BLDC machine was partially simulated with Simulink®, to measure the conduction period for both advanced and delayed commutation instants under controlled conditions. A machine with a resistance of 0.5Ω and a inductance of 1mH was simulated (see Appendix A for the simulation diagram). The back-EMF amplitude is 2V and the DC bus voltage is 6.5 V. These low values were required to keep the simulation stable. The inverter operates in six-step block commutation. The phase and line voltages, phase currents, as well as the diode conduction angles were measured while the commutation instants were delayed.
with an angle $D$, where a negative $D$ indicates advancement. The results are summarised in Table 8.1 and some simulation measurements are displayed in Figures 8.4 through 8.6. The results show that an advancement of the commutation instants doesn’t influence the conduction period of the diodes much, however delaying the commutation instants increases the conduction period significantly. We can therefore conclude that field-weakening operation doesn’t interfere with a sensorless algorithm based on the back-EMF, but doesn’t extend the range of operation of the sensorless algorithm either. One can take this into account when installing Hall-effect sensors: since perfect alignment isn’t always possible, it is better to slightly advance the sensors to ensure the DCP is minimal when switching to sensorless commutation. Likewise for a sensorless algorithm: it is better to slightly advance the commutation instants to obtain optimal results.

### 8.3 Detection Interval Adaptation

The debounce filter introduced in Chapter 7 is no longer sufficient to eliminate the commutation ‘spikes’, since the conducting periods are almost of the same length as (or even longer than)
the non-conducting periods. The interval where the back-EMF is sensed has to be adapted to ignore the periods of diode conduction. The first attempts involved choosing the interval such that the symmetric threshold of the interval would be just between the forward voltage of the free-wheeling diode (around 1V) and zero. However, this did not provide stable detection intervals. Two other options were tried: inserting a delay or measuring the phase currents to obtain a detection interval that rejects the DCPs.

8.3.1 Delay During Diode Conduction

Using the same debounce filter as in the previous chapter, a sensing interval is determined, which now includes the DCP. We then create a new interval of which the beginning is delayed and the ending coincides with the ending of sensing interval, as illustrated in Figure 8.7. The delay
should correspond to a fixed number of electrical degrees, and thus depend on the speed of the machine. For example, one could always delay the detection interval with $20^\circ$ el. to ensure the DCP is ignored. It is then of course necessary to prevent that the conduction period becomes longer than $20^\circ$, otherwise the algorithm will fail to correctly detect the zero-crossing. This can be done using current measurements: if the current is zero for a shorter time than $10^\circ$, the speed or load should be lowered.

Figure 8.6: Phase voltage, line-to-line voltage and phase current of the high-speed machine for $D = 10$ (simulation).

Figure 8.7: Illustration of the adapted sensing interval.
8.3.2 Current Measurements

If accurate and fast current measurements are available, as is the case on the test set-up, a simpler algorithm can be implemented that requires less tuning. Consider Figure 8.8: if we use the interval where the current is near zero in the phase that is non-energised as detection interval, the DCP is completely ignored. We can again use a debounce filter to make sure this interval is not intermitted by measurement errors. The striped yellow line in Figure 8.8 shows the boolean signal generated by a debounce filter with a time constant of 50µs. The advantage of this method is it solidly rejects the DCP with very little tuning. If the DCP changes due to load or speed changes, the detection interval will adapt smoothly with it. However, if the algorithm fails to detect a zero-crossing, there is little possibility for recovery, since the current waveform is a consequence of the commutation instants. If commutation doesn’t happen because of a missed zero-crossing, the current waveform will considerably deviate from the one during normal operation. As a consequence, the next detection interval will be wrong or even be absent, such that no more zero crossings can be detected. The motor will then stall. This could be solved by forcing a timed commutation at for example 65° el. after the last commutation if no commutation has happened after 62° el.

8.4 More Robust Sign Detection

In Section 7.3.3 we explained how the sign of the back-EMF is measured right after the start of the sensing interval, in order to set the emulated Hall-effect sensor 'low' or 'high'. During the experiments on the high speed machine it appeared that this was no longer accurate since the damped oscillations after the DCP (caused by the dynamic behaviour of the free-wheeling
diode) caused the algorithm to measure a wrong sign. This was solved using another debounce filter that keeps track of the sign of the back-EMF. Only when multiple consecutive samples are positive (or negative), the debounce filter will change a variable \textit{IsBackEMFPositive}.

### 8.5 Complete ZCD

In the LabVIEW diagram below, we see how the zero-crossing detection of the previous chapter, the new sensing interval and more robust sign detection are brought together to detect the crossings on the high-speed machine. If the back-EMF is within the adapted sensing interval (indicated by \textit{Delayed}, or for the algorithm using current measurements indicated by \textit{WithinCurrentInterval}), the algorithm will await a \textit{CommutateSignal}. This signal can be the first zero-crossing (of course with a detection ‘band’ around zero, as in the previous chapter) or several consecutive crossings, determined by another debounce filter. If \textit{CommutateSignal} is ‘high’, the emulated Hall-effect sensor \textit{simHall} will be set ‘high’ or ‘low’, depending on whether the back-EMF is positive or negative, as determined by the sign detection of 8.4. A delay is then added to avoid detecting successive zero-crossings. Note that \textit{CommutateSignal} is also implemented with a debounce filter, allowing to select if one or more zero-crossings are required to change its value.

![LabVIEW FPGA diagram of the zero-crossing detection loop.](image)

### 8.6 Results and Tuning

Both the algorithm using the delay during the DCP and the algorithm using current measurements were tested on the test set-up. It was not possible to safely test the algorithms above 12000 rpm. Measurements were made at 500, 2000, 5000 and 10000 rpm.

#### 8.6.1 Method Using Delay

The method using a delay to determine the sensing interval needed some tuning for different speeds. The tuning follows a general trend: for higher speeds the time constants of the debounce
filters have to be shorter. For low speeds the ZCD should use a low threshold value, such as 0.02 such that commutation instants are not advanced too much. For higher speeds, the threshold value should be a bit higher, for example 0.05 in order not to miss a zero-crossing. The amount of samples (time constant of debounce filter) that is used to determine the sign of the back-EMF has to be lowered for higher speeds, as well as the number of samples that is used for determining the detection interval. A first test of the sensorless algorithm at 2300 rpm shows that it works correctly. The real Hall-effect sensors are slightly delayed to make the difference between sensored and sensorless commutation more pronounced. In Figure 8.10 we see that the advancement of the commutation instants slightly shortens the DCP, as we could expect from the simulation results in Section 8.2.2.

![Diagram](image1.png)

(a) Commutation instants determined by Hall-effect sensor.

![Diagram](image2.png)

(b) Commutation instants determined by sensorless algorithm.

Figure 8.10: Line-to-line voltage, current, Hall sensors and emulated Hall sensors at 2300 rpm, no-load (DLM2000).

At 5400 rpm and at 10000 rpm we note similar results. It becomes more difficult at higher speeds to find the right settings for the different thresholds and time constants, again with the general trend of shorter constants and higher thresholds. Results at 5400 rpm are depicted in Figure 8.11. Important to note is that, in contrast with the current measurement method, a missed zero-crossing is not catastrophic. The back-EMF largely retains its shape, because of the continued movement of the rotor due to the inertia of the motor and load. Because the back-EMF remains the same, the next detection interval is activated and zero-crossings will be detected.

### 8.6.2 Method Using Current Measurements

With the method using current measurements, we tried to find the lowest speed from which the algorithm worked. It was found that the algorithm already stably works from 500 rpm, 1.6% of the nominal speed, as shown in Figure 8.12.

The algorithm was also tested at 10000 rpm (Figure 8.13). As for the method using the delayed detection intervals, the ZCD threshold had to be slightly higher in order not to miss zero-crossings. The general trend concerning time constants also applies here: time constants have to be shorter for higher speeds.
Figure 8.11: Line-to-line voltage, current, Hall sensors and emulated Hall sensors at 5400 rpm, no-load (DLM2000).

(a) Commutation instants determined by Hall-effect sensor.
(b) Commutation instants determined by sensorless algorithm.

Figure 8.12: Line-to-line voltage, current, Hall sensors and emulated Hall sensors at 500 rpm, no-load (DLM2000).

(a) Commutation instants determined by Hall-effect sensor.
(b) Commutation instants determined by sensorless algorithm.

The algorithm was also tested under a 0.08Nm load, with an RMS phase current of 3.2A, 40% of the nominal current. The results are displayed in Figure 8.14. We see that the DCP is indeed longer under load and that switching between the delayed Hall-effect sensor signals and the accurate sensorless signals slightly shortens the diode conduction period, as we could expect from the simulation results.

8.7 Outlook

Although both algorithms work until 10000 rpm, the detection interval using the delay has a clear preference since failure to detect a zero-crossing does not cause an immediate stall of the motor.
Due to technical problems with the set-up, it was not possible to feed the motor with more than 2A. Experimenting with the different delay and parameter settings caused the motor to sometimes draw excessive current, limiting the operating speed. It is however clear that at higher speeds, it becomes more difficult to tune the delays and thresholds, causing the algorithm to sometimes miss a zero-crossing. Before testing over the full speed range, a safety should be build-in that forcibly commutates if a commutation is missed by the algorithm, to prevent sudden stalls of the motor. As this thesis was limited in time, such safety was not implemented. Future research can work towards making a completely fail-safe algorithm.

It is clear that current measurements can complement an algorithm based on the back-EMF. However, the core working of the algorithm should not be based on current measurements, since the current waveform is a consequence of the commutation instants. Instead, current measurements can be used to check whether the DCP is not becoming longer than allowed for
a correctly working back-EMF based sensorless algorithm.
At the beginning of this work, the reader was introduced to high-speed machines and their important role in making electrical motor driven systems more efficient. We highlighted the emerging FPGA technology and its possibilities for use in digital control of high-speed machines. A literature study on PMSMs, optimal control of PMSMs and sensorless control was then performed. This literature study showed there is a need to control high-speed machines sensorless and that the sensorless methods based on the back-EMF are favourable for high speed machines. An digital sensorless algorithm based on the line-to-line back-EMF method from Kim et al. [2011b] was developed. The sensorless algorithm uses a debounce filter to establish 'detection intervals' where the back-EMF is measured for zero-crossings. The detection intervals are constructed in such way that they ignore the long diode conduction period that is inherent to high-speed machines. On variation creates a detection interval by using a delay during the DCP. Another variation measures when the phase current is zero and uses this information to construct the interval. Both variations were experimentally verified on a test set-up from 500 rpm until 10000 rpm, 1.7% and 33% of nominal speed respectively, demonstrating the validity of fully digital sensorless control. On one hand, the sensorless method based on the delay alone can recover from missed zero-crossings. However, there is a risk of the DCP becoming longer than the fixed delay, which will result in failure of the algorithm. On the other hand, the method using current measurements smoothly adapts the detection interval to the DCP. Unfortunately, it is likely the algorithm will not detect subsequent crossings after missing a zero-crossing, since the current waveform is a consequence of the commutation instants. We thus propose that current measurements should not be the core of a sensorless back-EMF based algorithm, but can play a supporting role in detecting when the DCP becomes longer than the delay.

Before testing over the full speed range, a safety should be build-in that forcibly commutates if a commutation is missed by the algorithm, to prevent sudden stalls of the motor. The algorithm itself shows promise to work over the full speed range. Research can still be done to make the algorithm completely fail-safe. The use of FPGAs allow great flexibility to improve the algorithm in future research.


APPENDIX A

Model of a high-speed machine in Simulink and SimScape.
APPENDIX A. MODEL OF A HIGH-SPEED MACHINE IN SIMULINK AND SIMSCAPE.
APPENDIX B

Model of the buck-converter with set-up parameters in Simulink and SimScape.
APPENDIX B. MODEL OF THE BUCK-_CONVERTER WITH SET-UP PARAMETERS IN SIMULINK AND SIMSCAPE.