Adaptive Control Scheme for a Practical Bidirectional DC-DC Converter with a 80 kHz Switching and a 10 kHz Sampling Frequency

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Keywords: bidirectional DC-DC converter, high switching frequency, adaptive control, hybrid electric vehicle.

Abstract

This paper discusses an adaptive control scheme for a bidirectional boost converter used in a hybrid gasoline-electric vehicle. The converter transforms the voltage of the low-voltage battery, approximately 96 V, to 400 V on the DC-link of the motor inverter. The switching frequency of the inverter is 80 kHz, which enables the usage of rather compact 64 µH inductors. However, the sampling frequency of the control scheme is 10 kHz, so the duty cycle of the switches will only be updated after 8 consecutive switching periods. Due to the large dead-time of the switches in comparison with the switching frequency, the response of the current to changes in the duty cycle is highly non-linear. The low sampling frequency of the control scheme combined with the highly non-linear current response makes it impossible for a simple PI-controller to regulate the current in a satisfactory way. At first the cause of the non-linear response is investigated. Based on that knowledge, a control scheme is developed that adjusts the duty cycle when necessary and switches between two PI-controllers, one for small currents and one for large currents. This current controller is implemented on two interleaved bidirectional boost converters. The current controller's set point is regulated by the DC-link voltage controller.

1 Introduction

A series hybrid electric go-cart was built at the K.U.Leuven for didactical and research purposes [1]. The vehicle is equipped with a three-phase 4 kW induction motor to drive the rear axle. The induction machine is connected to an inverter which is supplied by a low-voltage battery and a gasoline generator. The portable gasoline generator is rated at 1.6 kVA. The generator produces a 230 V_{AC} single-phase voltage, which has to be converted into a 400 V_{DC} voltage for the DC-link of the inverter. A unidirectional power-factor-correcting (PFC) converter is used to operate the generator at unity power factor. In this way the generator can deliver up to 1.6 kW of active power. The low-voltage battery is made up of 8 series-coupled 12 V/30 Ah lead-acid absorbed glass-mat (AGM) batteries and is connected to the inverter through a bidirectional DC-DC converter [2] which controls the voltage of the 400 V DC-link.

The control scheme of the vehicle consists of 5 parts: The first part collects the different input and output signals, the second part controls the torque of the induction motor, the third part controls the power delivered by the generator through the control of the PFC converter, the fourth part controls the bidirectional DC-DC converter and the fifth part controls the power flow of the battery and gasoline generator, based on the battery state of charge and the power consumed by the induction motor. The control scheme runs on a single-board computer (SBC) and is connected to a FPGA which is responsible for generating the control signals for the IGBTs of the motor inverter, the PFC boost converter and the bidirectional DC-DC converter. The sampling and switching frequency of the motor inverter and PFC boost converter is 10 kHz. Due to the 12 mH inductance of the PFC boost converter and 100 mH phase inductance of the induction machine, the current ripple at 10 kHz through these components is small. The current flowing through the PFC boost converter is limited to 10 A, resulting in a relatively thin wire and corresponding limited weight and volume of the iron core inductor. The current originating from the low-voltage battery can have a peak value as high as 60 A and exhibits a strongly peaked shape. This would expose the buffer capacitors and the battery to very high stresses. Therefore, the bidirectional DC-DC converter is implemented as two distinct interleaved converters operating in parallel, each carrying a current of 30 A through a 64 µH ferrite core inductor. To limit the current ripple of these converters, while keeping the size of the inductors relatively small, the switching frequency was set at 80 kHz. The control scheme controlling the boost converters, however, runs at 10 kHz, which implies that the duty cycle calculated by the control scheme is applied 8 consecutive times before the SBC is able to compute a new duty cycle. Because the inductors are sized for a switching frequency of 80 kHz, the dynamics of the system are very fast compared to the sampling frequency. In order to establish a stable control scheme in all circumstances an adaptive control scheme was implemented that changes the parameters of the PI-controller depending on the magnitude of the currents.

2 PLECS model of the DC-DC converter

At first a model of a bidirectional DC-DC converter is made with the PLECS toolbox in Simulink. The model uses 2 ideal switches and diodes, together with a lossless inductor. The primary side of the converter is connected with a variable voltage source representing the battery, of which the voltage reflects the state-of-charge. The secondary side of the converter is connected to a 4 mF capacitor and a variable current source which represents the current drawn by the motor. This model is shown in figure 1.

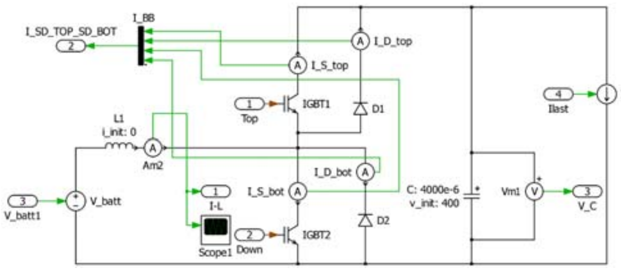

figure 1: PLECS model of the bidirectional boost converter

The dead-time of the switches must be incorporated in the simulation. First of all, the duty cycle, which is defined for the top switch and ranges between 0 (open) and 1 (closed), is transformed into a reference signal ranging between -1 (open) and +1 (closed). The bottom switch state is complementary to the top switch state. Secondly, the reference signals which are compared to the carrier are raised and lowered for the top switch and bottom switch respectively.

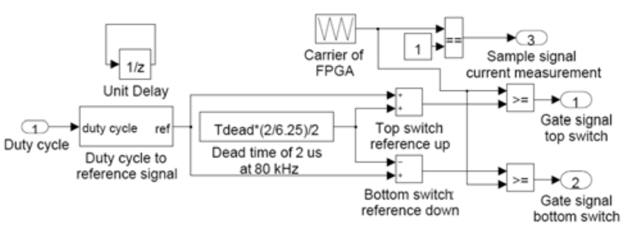

figure 2: Algorithm for transforming one duty cycle into two gate signals with dead-time.

The value by which this occurs is calculated as follows: The 80 kHz carrier signal is a sawtooth with a period of 12.5 µs, which means it rises 2 units (from -1 to +1) in 6.25 µs, which is clearly visible in figure 3. The dead-time of 2 µs thus corresponds with a change in reference value of 2*(2/6.25). The dead-time is equally distributed between the switches, which results in a change in reference value of 0.32 (16 % of full range) for each switch. The adjusted reference signals are compared with the carrier (figure 3). When the reference signal is larger than the carrier, the gate signal is high and vice versa. The algorithm that transforms one single duty cycle in 2 distinct gate signals for the IGBTs is depicted schematically in figure 2. This results in an advanced switching-off of the current carrying switch and a delayed switching-on of the complementary switch.

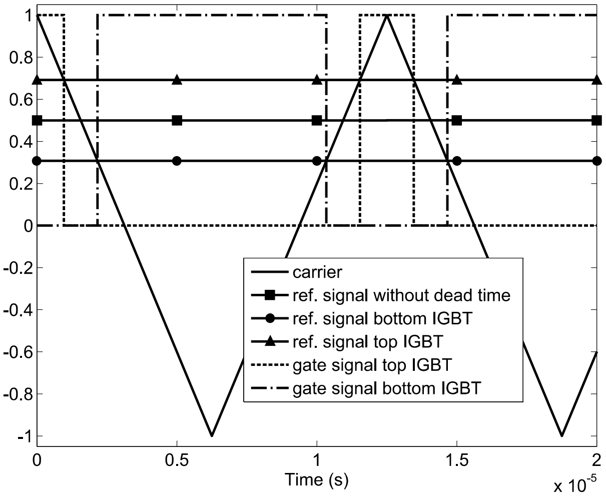

figure 3: Reference signals for the switches and corresponding gate signals

The difference between the switching frequency and sampling frequency is incorporated in the simulation by generating the gate-signals in continuous time using an 80 kHz PWM carrier frequency while implementing the controller as a discrete-time system with a sample frequency of 10 kHz. The duty cycle calculated by the controller is passed on 8 times to the algorithm depicted in figure 2 before an update occurs.

3 Discontinuous currents caused by dead-time

In the case of currents in the vicinity of zero and at low switching frequencies, e.g. 10 kHz, the top switch takes over the current from its anti-parallel diode when the current passes from a positive to a negative value. This ensures that the converter operates in continuous current mode and as an additional benefit, the converter operates under zero-voltage switching conditions in this specific region. In this case however, the dead-time of the IGBTs of the converter is 2 µs, while the switching period is 12.5 µs. This implies that the dead-time represents a rather large portion of the switching period. In the case of a positive average current with the lower peak near zero, the large dead-time results in the switching off of the top switch during the period it is supposed to take over the current from its anti-parallel diode (figure 4). This results in a discontinuous current through the converter. If the duty cycle of the bottom switch rises, the battery voltage is applied longer to the inductor, resulting in a longer current carrying period of the bottom switch and top diode. However, the discontinuity still exists and only becomes smaller by applying the greater duty cycle, this can be seen by comparing both current samples of figure 4. Thus, in this situation the current will barely react on changes in the duty cycle.

The situation ceases to exist when the duty cycle of the bottom switch is high enough to allow it to take over the current from the anti-parallel diode of the top switch. The current becomes continuous and in the case that the average current is positive, only the bottom switch and top diode will conduct. The same considerations are valid for a negative current of which the upper peak is near zero. As soon as the current becomes continuous, the reaction to small changes in the duty cycle is very large. This is due to the steep increase of the current through the small inductance, so that any rise in the duty cycle immediately results in a larger average current. This is illustrated in figure 5, where the current barely changes if the deviation of the duty cycle is less than 0.16 from its equilibrium value, while the current is extremely sensitive to changes in the duty cycle above that value. The reaction of the current to changes in the duty cycle is clearly non-linear. The equilibrium value is defined as the duty cycle of the ideal bidirectional boost converter, i.e. $\delta = 1 - V_{in}/V_{out}$.

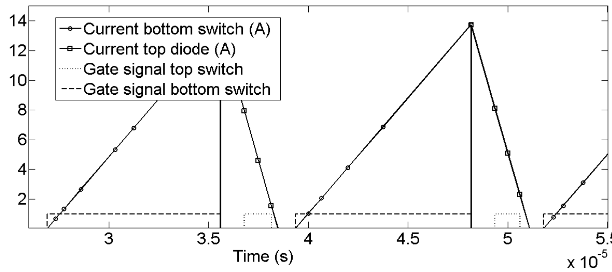

figure 4: Discontinuous currents due to dead-time

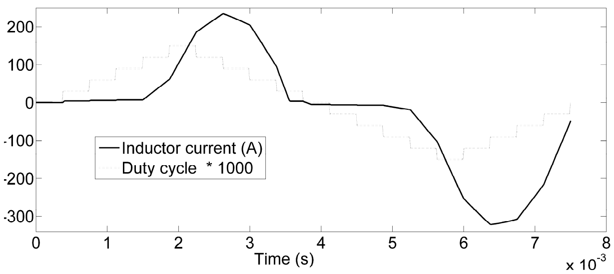

figure 5: Non-linear respons of inductor current on duty cycle

4 Adaptive current controller design

Due to the extremely non-linear response of the current to changes in the duty cycle, the current can not be controlled sufficiently accurately by a simple PI-controller at 10 kHz. On the one hand it is possible to design a fast PI-controller that copes with the currents in the vicinity of zero, but the response of this controller for large currents is unstable. On the other hand it is possible to design a mild PI-controller that copes with fast varying, i.e. larger currents, but this controller is unacceptably slow for small currents. In order to cope with this problem an adaptive controller is developed.

The first problem is to make the distinction between large and small currents. In this particular case, small currents are those currents that barely react on the changes in the duty cycle, while large currents are extremely sensitive to those changes. This situation is explained for positive currents. As explained in section 3, currents will barely react on changes in the duty cycle as long as the discontinuity exists. The discontinuity disappears when the lower peak of the current is at zero. This is visible for the second current sample of figure 6. A small change in the duty cycle will now result in an immediate rise of the current, which is visible for the third and fourth current sample of figure 6. So the distinction between small and large currents is determined by the boundary between continuous and discontinuous currents. At this boundary, the lower peak of the current is equal to zero. The peak-to-peak value of the current through the inductor is determined by the input voltage V_{in} of the converter, the duty cycle δ of the bottom switch, the inductance L of the inductor and the switching frequency f_{sw}, resulting in the following formula: $I_{P2P} = V_{in}\delta / (Lf_{sw})$. The input voltage of the converter is assumed to be 96 V, due to the series connection of eight 12 V batteries. The duty cycle is determined by the input and output voltages according to $\delta = 1 - V_{in}/V_{out} = 0.76$. The inductance and switching frequency are 64 µH and 80 kHz respectively. This results in a peak to peak value of the current of 14.25 A. The current is measured when the value of the carrier is equal to 1. According to figure 3, this instant is halfway the on time of the top switch. Due to the triangular waveform of the current through the inductor, this means that at this instance the instantaneous current through the inductor is equal to the average inductor current. Thus, the measured current at the boundary between continuous and discontinuous currents is the average of a triangular current with a peak-to-peak value of approximately 14 A, which corresponds to 7 A.

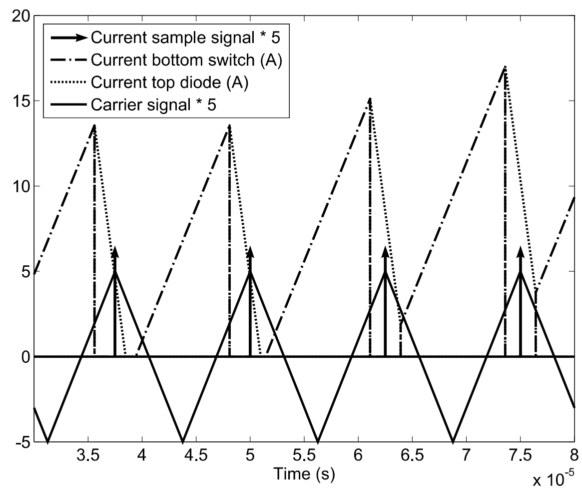

figure 6: Sampling of currents and distinction between small and large currents.

The first step of our solution is to design a current controller that can regulate the currents with an absolute value below 7 A. This controller is a PI-controller (figure 8) with anti-windup. The values of P and I are 1.5e-3 and 8 respectively. This controller needs to be very fast to compensate for the slow reacting system, i.e. the controller changes the duty cycle with an order of magnitude of 0.1 in the region between -7 and 7 A. This ensures that the set point of the current is reached in approximately 10 ms or 100 periods of the controller.

The effect of the P- and I-action of the controller can be examined in figure 7, which represents the simulated response of 2 current controllers. The figure displays the P- and I-action of two controllers with identical values for P, but the I-value of the second controller is 4 times larger. The P- and I-action and the inductor current of the second controller are marked with an "□". Both controllers react on a step change in the set point of the current from -5 to +5 A. Since both P-values are identical, the initial reaction of the P-part of the controllers is the same and changes the duty cycle 0.04. If the controller would be a purely proportional, the duty cycle would stabilize at 0.01, causing the current to stabilize at 3 A and leaving a 2 A steady state error. This behaviour is approximated by the first controller between 3 and 3.1 ms. The remaining steady state error is integrated by the slow I-action of the first controller, causing the duty cycle and current to rise slowly. Increasing the value of the I-action causes the controller to react much quicker on the error between the measured current and set point. Although the initial reaction of both P-actions is identical, the duty cycle response of the second P-action shows a much steeper decrease and will fall to zero after 0.3 ms due to the stronger I-action. After the initial response of the P-action on the step change, the duty cycle is entirely determined by the I-action. A further increase of the I-action is unnecessary, since the response is already quick enough to deal with the power demands of the motor. If an increase of the I-value occurs, the response will show an overshoot and become unstable.

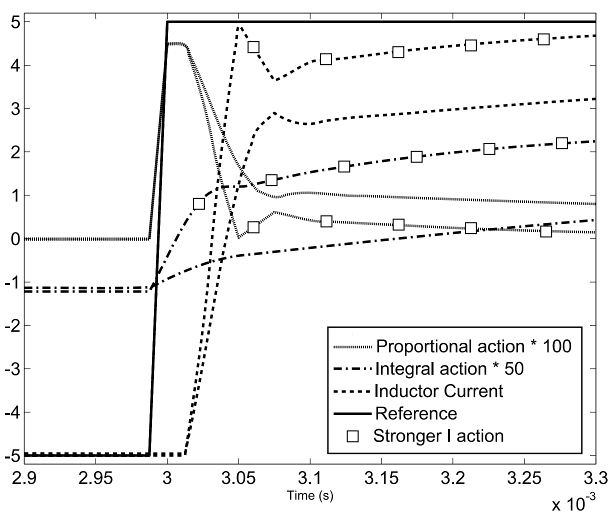

figure 7: Reaction of the 2 current controllers to a step change

Currents above 7 A can not be regulated by this controller, because the controller will change the duty cycle too much when an error between the set point and measured current is detected. This results in overcurrents, which trip the security settings.

The P- and I-values of the actual controller are determined by connecting the boost converter with the batteries at the low-voltage side and with a 400 V DC source at the high-voltage side. Both voltages are measured and based on the formula $\delta_{FF} = 1 - V_{in} / V_{out}$, the duty cycle feed forward term δ_{FF} is determined. The δ_{FF} term is added to the response of the PI-current controller to obtain the duty cycle of the bottom switch of the boost converter. At first, the I-value is set to zero and the P-value is determined until the response of the P-action is quick, but without any instability. The steady state error must be dealt with by the I-action, which is slowly raised. Raising the I-action will cause the current to rise more quickly towards the set point, but must be stopped when overshoot occurs. Making a controller that works without an overshoot, allows to implement a software-based current protection, by putting a saturator at the set point of the current controller.

The second step is to design a second current controller of the same configuration as the first one, but with different values for P and I, such that it can regulate the currents with an absolute value above 7 A. The determination of the values of P and I is performed in the same way as in the case of the controller for the small currents. The obtained values are much smaller however, with P- and I-values of 4.5e-4 and 0.4 respectively. This controller needs to be very mild in order to cope with the fast reacting system, i.e. the controller changes the duty cycle with an order of magnitude of 0.001 to regulate the currents with an absolute value between 7 and 30 A. This controller is far too slow to deal with the currents below 7 A, e.g. a step change from 0 to 5 A takes the fast controller for small currents 5.5 ms to reach 4.5 A, where the mild controller for large currents needs 60 ms to reach the same current value.

The third step is to make an algorithm, the so-called Platform Avoidance System (PAS), that determines which of the controllers should be used and brings the duty cycle in the vicinity of the correct duty cycle. The latter is based on a test where the voltages at the input and output of the boost converter are kept constant and the duty cycle is controlled manually to determine at which value of the duty cycle the current reaches 7 A. The test revealed that this duty cycle was approximately 0.16 above the feed forward duty cycle. The algorithm first determines the current value of the duty cycle. If the duty cycle is below -0.16, the current is considered a large negative current , between -0.16 and 0 the current is considered a small negative current , between 0 and 0.16 the current is considered a small positive current and above 0.16 the current is considered a large positive current. The values of the duty cycle δ are shown in the first column of table 1 and table 2. In a next step the value of the set point of the current is taken into consideration, which is shown in the first row of table 1 and table 2.

$\delta\downarrow$ $I_{Ref}\rightarrow$	$I_{Ref} > 7$ A	$I_{Ref} < -7$ A
$\delta > -0.16$	/	$\varepsilon < -2$ A => $\delta_{SP} = -0.16$
$\delta < 0.16$	$\varepsilon > 2$ A => $\delta_{SP} = 0.16$	/

table 1: Determination of duty cycle set point δ_{SP} for currents with an absolute value above 7 A.

$\delta\downarrow$ $I_{Ref}\rightarrow$	$I_{Ref} > 0$ A	$I_{Ref} < 0$ A
$\delta > 0.16$	$\varepsilon < -2$ A => $\delta_{SP} = 0.08$	$\varepsilon < -2$ A => $\delta_{SP} = -0.08$
$\delta < -0.16$	$\varepsilon > 2$ A => $\delta_{SP} = 0.08$	$\varepsilon > 2$ A => $\delta_{SP} = -0.08$

table 2: Determination of duty cycle set point δ_{SP} for currents with an absolute value below 7 A.

The first part of the algorithm is shown in table 1 and determines the set point δ_{SP} of the duty cycle for large currents. When the current duty cycle indicates that the current is small or of an opposite sign of the set point and the set point I_{Ref} indicates that the set point of the current is changed into a large current, the algorithm will intervene by

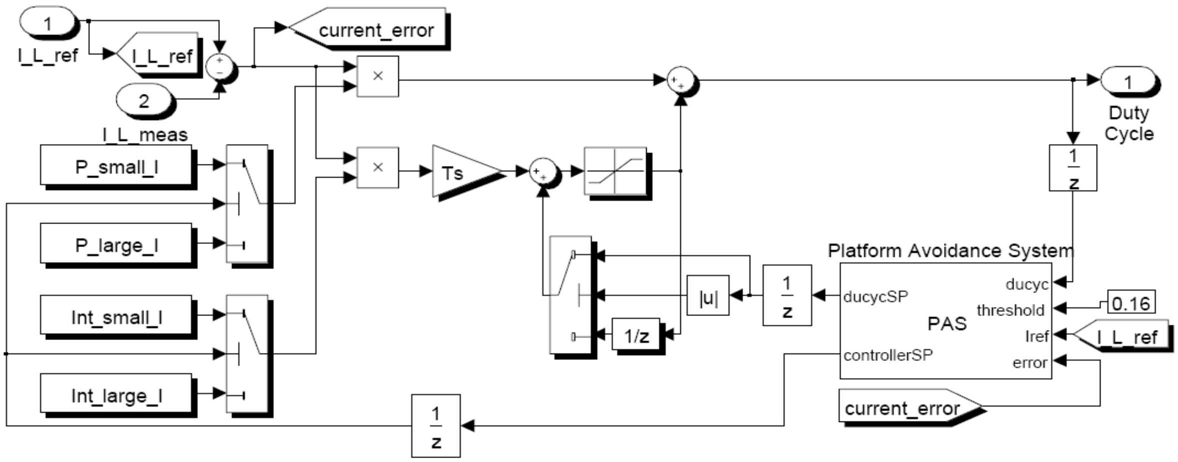

figure 8: Algorithm of the adaptive current controller with PAS

resetting the integrator of the I-action to the value defined in table 1 if the absolute value of the error ε is larger than 2 A.

The second part of the algorithm is shown in table 2 and determines the set point of the duty cycle for small currents. Based on the current duty cycle, the algorithm will first determine whether the present current is a large positive or a large negative current. In a next step, the algorithm will reset the integrator of the I-action to the value δ_{SP} defined in table 2, based on the sign of the set point I_{Ref} and taking into account the magnitude and sign of the current error ε.

The third part of the algorithm determines which P- and I-values of the PI-controller are used, based on the set point of the current. When the set point of the current changes to a small respectively large value, the P- and I-values turn the controller into a fast respectively mild controller.
The structure of the PI-controller with anti-windup and the integration of the algorithm, which determines δ_{SP} and is implemented in an embedded Matlab function, in the integrating part of the control loop is presented in figure 8.

5 Voltage controller and interleaved operation of parallel converters

The adaptive current controller will in practice be used on two parallel bidirectional boost converters. These are equipped with 2 Fairchild HGTG30N60A4D IGBTs that can block up to 600 V and conduct up to 60 A, but are derated to 30 A due to the high switching frequency. The converters are operated in an interleaved mode, which reduces the current ripple drawn from the battery. Each parallel branch has a separate current controller, as shown in figure 9. If the duty cycles of both converters were to be controlled by one single current controller, small differences in the hardware of the converter, such as different resistance values of the wires and different inductance values of the inductors, would result in strongly deviating currents through the converters. This problem can not occur when each converter has its own current controller.
The voltage of the DC-bus is controlled by a voltage controller which operates in cascade with the two current controllers, i.e. the voltage controller uses the deviation from the voltage set point to calculate a current set point for the current controllers. This voltage controller is a simple PI-controller with anti-windup. The values of P and I are 1 and 3 respectively, which is the order of magnitude to be expected. This means that a voltage deviation of 1 V will result in a P-action of 1 A and I-action of 3 A if the deviation would remain stable at 1 V for 1 second.
When a discontinuous current is flowing through the converter, the measured current deviates from the actual average current, because the instantaneous current measured halfway the on time of the top switch is lower than the average current. This does not pose a problem for maintaining the DC-link voltage at the set point, since the voltage controller will command a higher current set point for the current controllers to adjust the duty cycle to the appropriate

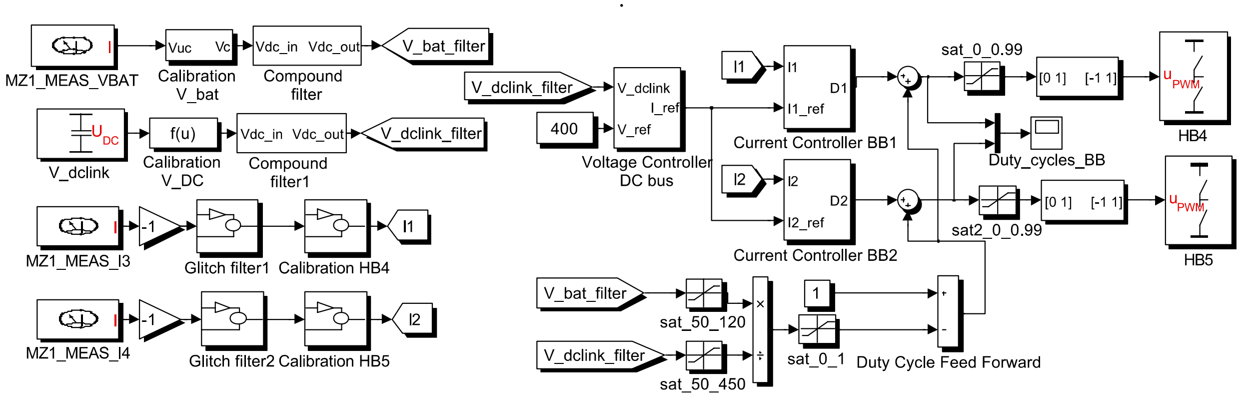

figure 9: Measurements and control scheme of the two interleaved bidirectional boost converters

value.
In total 4 measurements are required for this control scheme. Two voltage measurements measure the battery and DC-link voltage. Both voltages are filtered with a compound filter, composed of a low-pass filter and a band-stop filter. The low-pass filter has a cut-off frequency of 25 Hz and suppresses the measurements noise. The band-stop filter has its -3 dB points at 80 and 120 Hz. This filter blocks the 100 Hz originating from the rectifier of the PFC boost converter. Without this filter the DC-link voltage starts oscillating due to the voltage ripple caused by the PFC rectifier. Two current measurements measure the current through the inductor of the converters, as indicated in by measurement *Am2* in figure 1. These measurements are filtered with a glitch filter to suppress spikes in the current measurements

6 Experimental results

The first test is a step change of the current reference from -5 A to 30 A and back to -5 A (figure 10). As soon as the step change of the reference value is applied, the PAS resets the I-action of the PI-controller (i.e. part of the duty cycle) from approximately -0.16 to +0.16, while the P-action adds an additional 0.02 to the duty cycle. This step change of the duty cycle is clearly visible in the figure and causes the current to jump from -5 A to 12 A, after which the PI-controller adjusts the duty cycle to enable the current to rise to the 30 A reference value in 10 ms. The duty cycle originating from the PI-controller is added to the feed forward value of the duty cycle to produce the total duty cycle. The step change back to -5 A will occur in a similar fashion: the current jumps to 5 A, after which it is adjusted to the -5 A reference value.

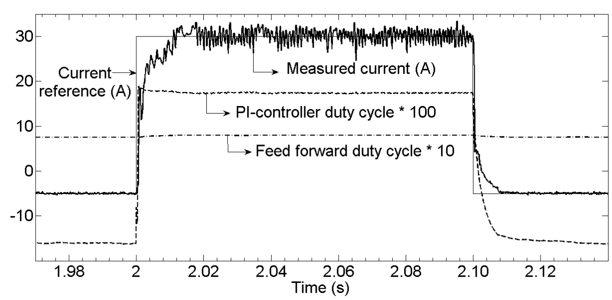

figure 10: Response on step change in current reference from -5 to 30 and back to -5

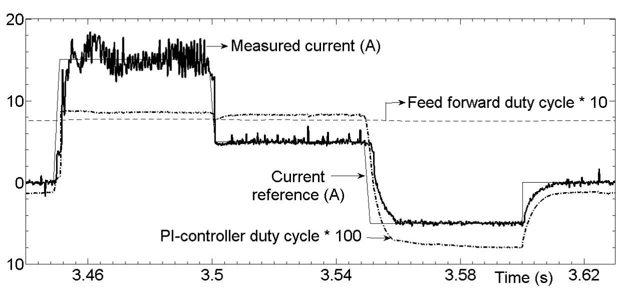

figure 11: Response on step change in current reference from 0 to +15 over +5 to -5

A second test (figure 11) examines smaller steps in the current reference; a first step from 0 A to 15 A, a second step to 5 A, a third step to -5 A and back to 0 A. In the first two steps the current reference passes the 7 A boundary between large and small currents and the PAS will cause the current to jump to the vicinity of its set-point. In the last two steps there is no need for the PAS to intervene and the PI-controller will adjust the current until it reaches its set-point.

The tests prove that the PAS allows the current to change rapidly without being hindered by the non-linear response of the current on the duty cycle. The tests also reveal that the two separate controllers for large and small currents are very effective. When the current reference changes from 5 A over -5 A to 0 A, the current will reach its set-point in 10 ms, while the large currents reach their set-points in the same time-span. The large currents oscillate round their set point due to the lack of resolution of the duty cycle, which is limited to 0.002. This causes the current to switch between two distinct values, e.g. a 15 A current will switch between 14 A and 16 A.

The test on the go-kart (figure 12) implies an acceleration of the induction motor from standstill to 5000 rpm at full throttle in 0.8 s in no-load condition. This translates in a power demand that rises from 0 to 4.5 kW in 0.15 s, followed by a decrease to 3.2 kW and back to 4.5 kW. In order to cope with this power demand the two interleaved converters draw about 50 A (25 A each) out of the battery. Due to the quick response of the current controllers, the maximum deviation from the 400 V DC-link set point is 20 V, which is more than satisfactory, given the fact that the space vector modulation of the motor control scheme can cope with voltage levels down to 330 V and the built-in over-voltage protection will only shut the inverter down when the voltage level goes above 450 V.

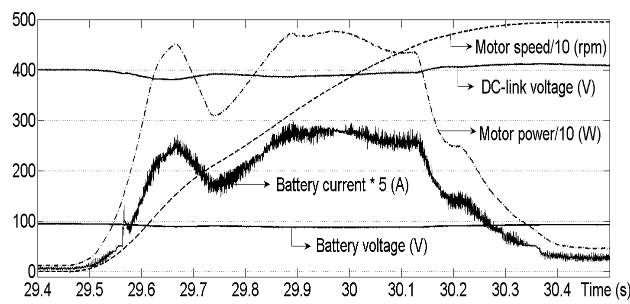

figure 12: Response of DC-link voltage and battery current on full acceleration of the induction motor

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