Contactor driver using the VNH7100BAS

Introduction

There are two types of primary circuits for high voltage contactors: contactors with a two-coil primary, and contactors with a single coil primary. The dual coil primary contactor has a high current leg for pull-in (or pick-up) and lower current leg for lower hold currents to minimize steady state power dissipation/consumption in the primary.

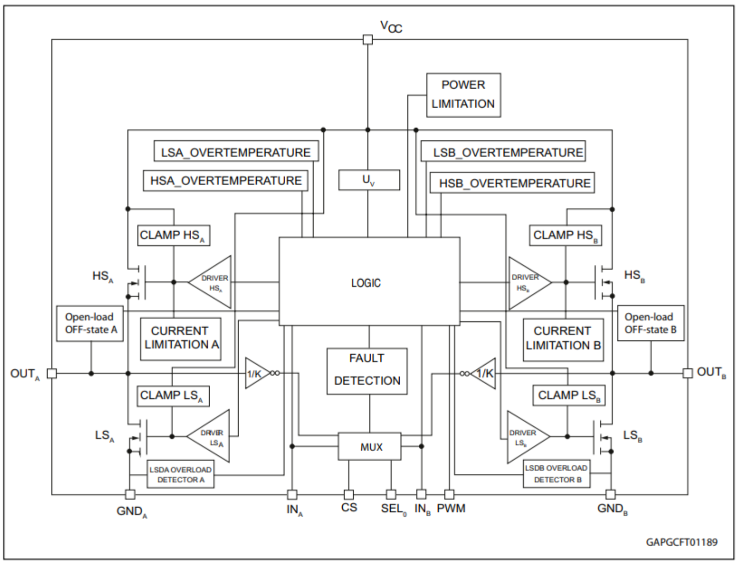
The contactors with single coil primaries require a variable drive mechanism that provides a higher current fast pull-in, lower power hold current, and higher voltage flyback for faster turn-off. These three states of operation combined with a need for safe actuation (no inadvertent actuations) require something more than a simple high side or low side driver.

This solution needs:

- An independent high side driver to drive the upper end of the primary coil
- An independent low side driver to drive the lower end of the coil separately
- A freewheeling element to recirculate inductive current during steady state PWM operation
- A fast recirculation path to quickly reduce the coil current and disable the contactor rapidly.

Some contactors include an “economizer” circuit that provides all of the above. All that is needed is to provide a voltage on the primary circuit. For those that do not have an economizer, the VNH7100BAS can accomplish all of these elements with only a microcontroller.

Figure 1. VNH7100BAS block diagram



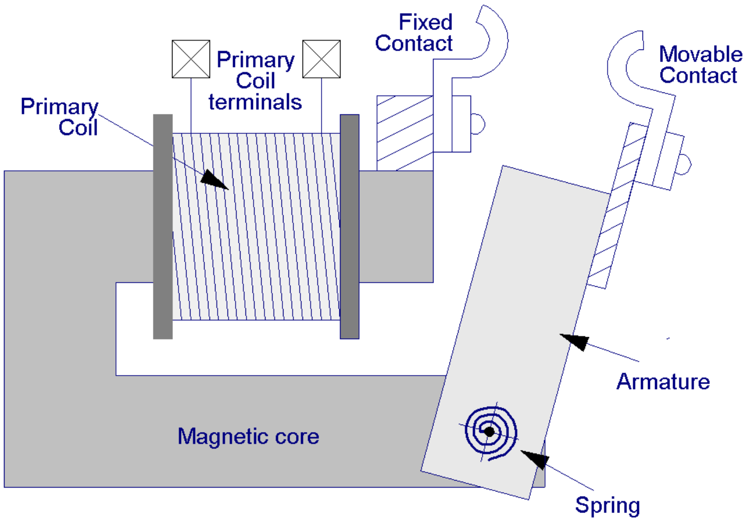
There are two ways to accomplish the pull-in and hold currents. One way is somewhat open loop by monitoring the battery voltage and calculating the desired duty cycles. This is the least expensive while it does take some microcontroller effort. This method is discussed in detail in the first part of this application note.

The second method monitors the motor current and forces a PWM to regulate the current to the desired levels. This eliminates the microcontroller requirement while it does add several components to the BOM.

1 Contactor components

The single coil contactor consists of a primary coil, a magnetic core, fixed and movable contacts, and an armature on some sort of sprung hinge. In a normally open contactor, the spring keeps the contacts apart in the “at rest” condition. Contact is made when current flows in the primary creating a magnetic field. The magnetic field overcomes the spring forcing the armature to rotate causing the movable contact to connect with the fixed contact, completing the circuit.

Figure 2. **Single coil contactor construction (simplified)**



The current required to pull the armature and close the contacts is greater than the current required to maintain a reliable connection between the two contacts. The pull-in current requirement can be as high as 5 A. The required minimum hold current can be in the tens of milliamps.

2 Application schematics

Figure 3. **Simplified application circuit, micro controlled**

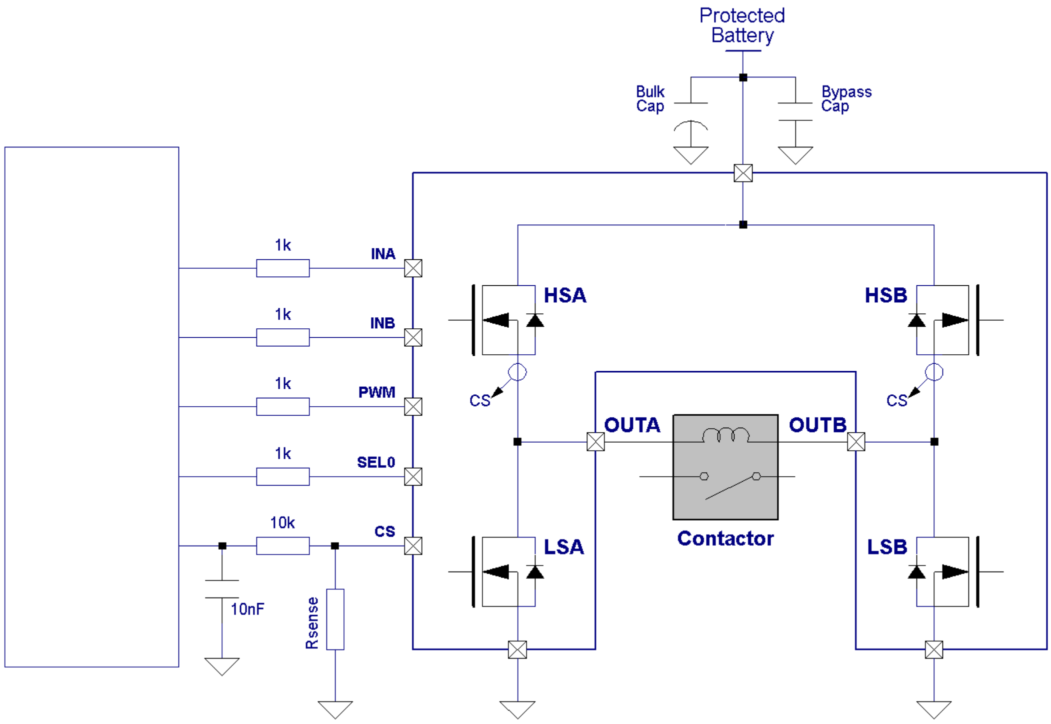
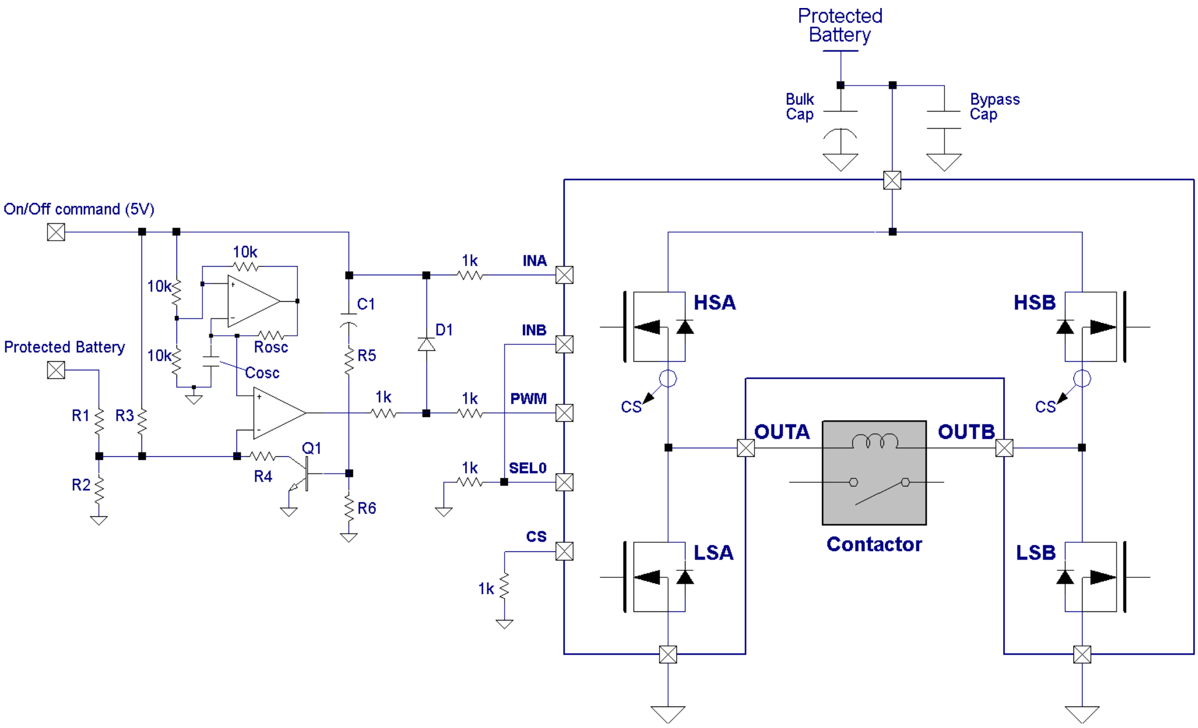


Figure 4. **Simplified schematic, micro independent**



3 Sequence of operation

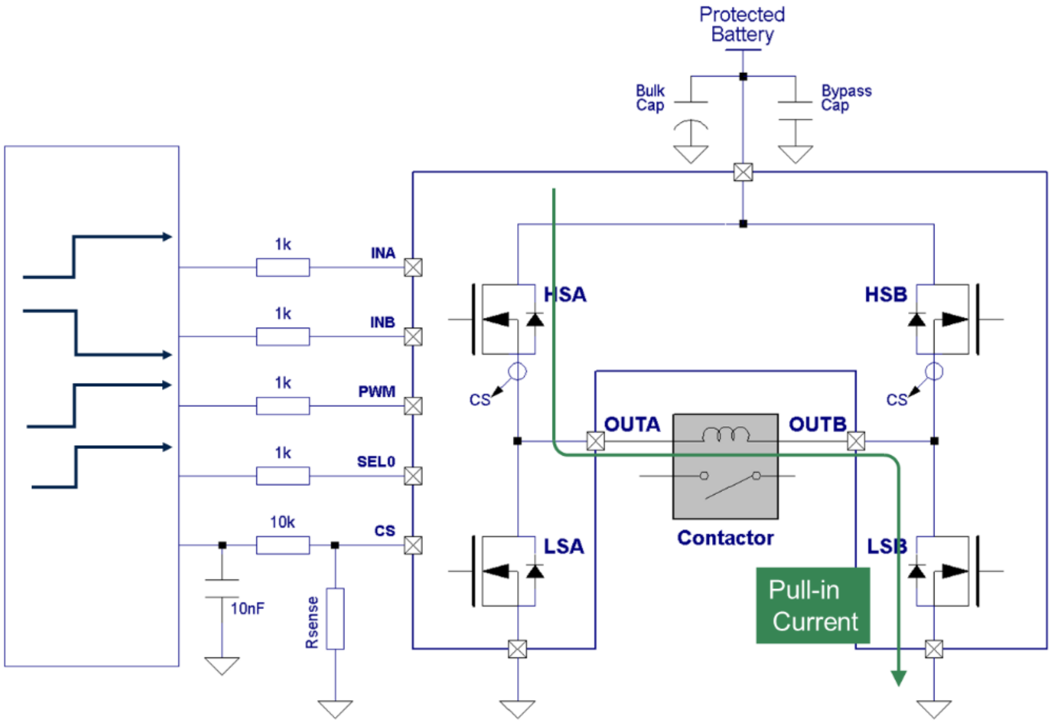
There are actually four phases of operation:

1. Pull-in
2. Slow recirculation to the hold current
3. Hold current regulation
4. Fast recirculation to remove quickly the coil current at turn-off.

3.1 Pull-in

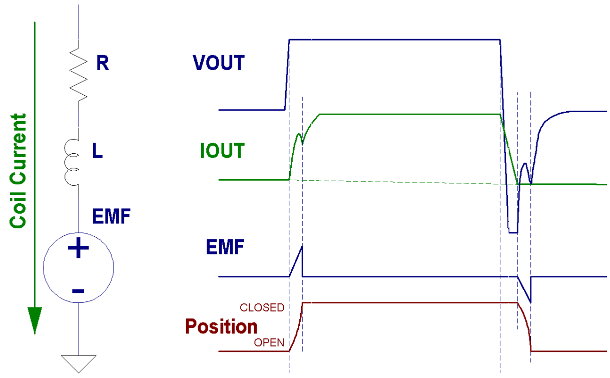
A high current pull-in to actuate initially the contactor can be as simple as full voltage across the primary coil. In this example, the pull-in current is flowing from HSA to LSB with no PWM.

Figure 5. Pull-in current flow



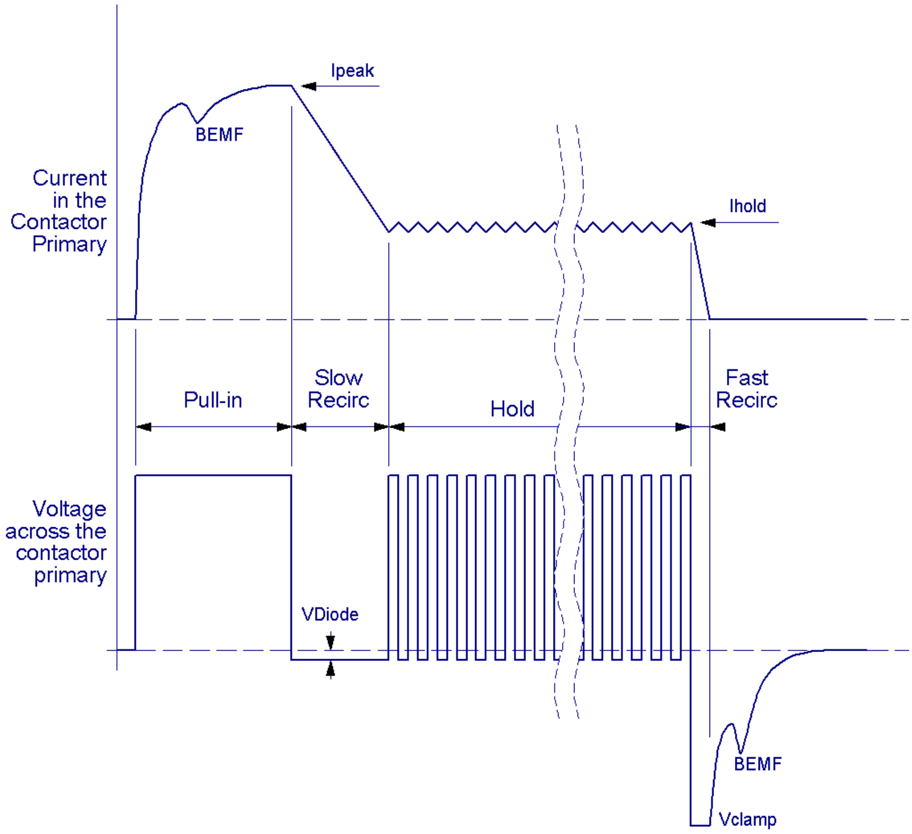
During the pull-in phase, the armature moves to close the high voltage contacts. This armature movement may generate a back EMF effect on the current. There is no BEMF when the armature is stationary (fully closed or fully open). This is similar to what occurs in a permanent magnet motor. The electrical equivalent circuit looks like the Figure 6. This figure illustrates the effect of BEMF voltage on the pull-in current and fast recirculation voltage.

Figure 6. Contactor primary electrical model



BEMF appears as a dip in the pull-in current typically as it is still rising (see Figure 7). During pull-in, the SEL0 pin can be held High so that the current sense pin (CS) reflects the current in HSA. That current can be monitored to verify that the expected back EMF event occurred. This provides a secondary confirmation that the contactor behaved as expected.

Figure 7. Contactor voltage and current waveforms

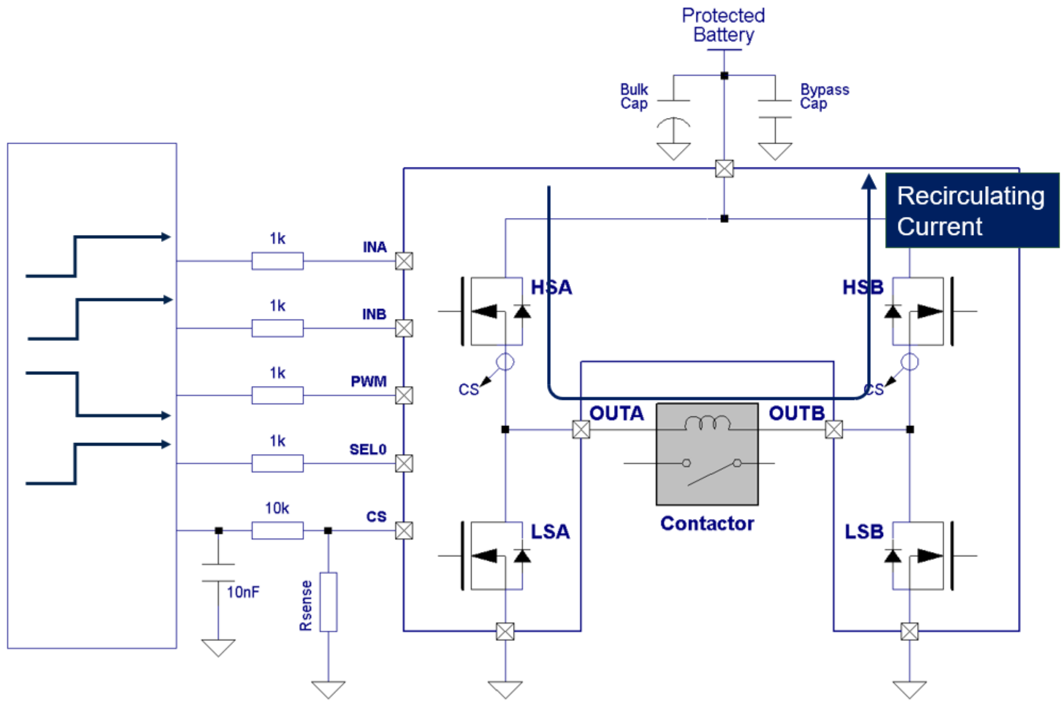


The pull-in current can be regulated as well if desired. The equations provided in the Section 3.3 Hold current below apply to pull-in current regulation as well.

3.2 Slow recirculation

The slow recirculation time between the end of the pull-in cycle and the beginning of the hold current regulation (see Figure 7) can be determined by knowing the battery voltage, load inductance, and resistance.

Figure 8. Slow recirculation path



During this time both high side drivers (HSA and HSB) are active, and the current is just recirculating.

Off time from pull-in to hold

$$t_{OFF}(V_{Batt}) = \frac{L_{Coil(min)} \ln\left(\frac{V_{Batt}}{I_{Hold} R_{Coil(max)}}\right)}{R_{Coil(max)}} 90\% \quad (1)$$

The slow recirculation time (t_{OFF}) needs to be shorter than the shortest possible t_{OFF} time. Having a longer than needed slow recirculation time would allow the coil current to fall below the minimum hold current. This may cause the contactor to open. Having a shorter slow recirculation time has no harmful effects on the operation and provides some margin to the system.

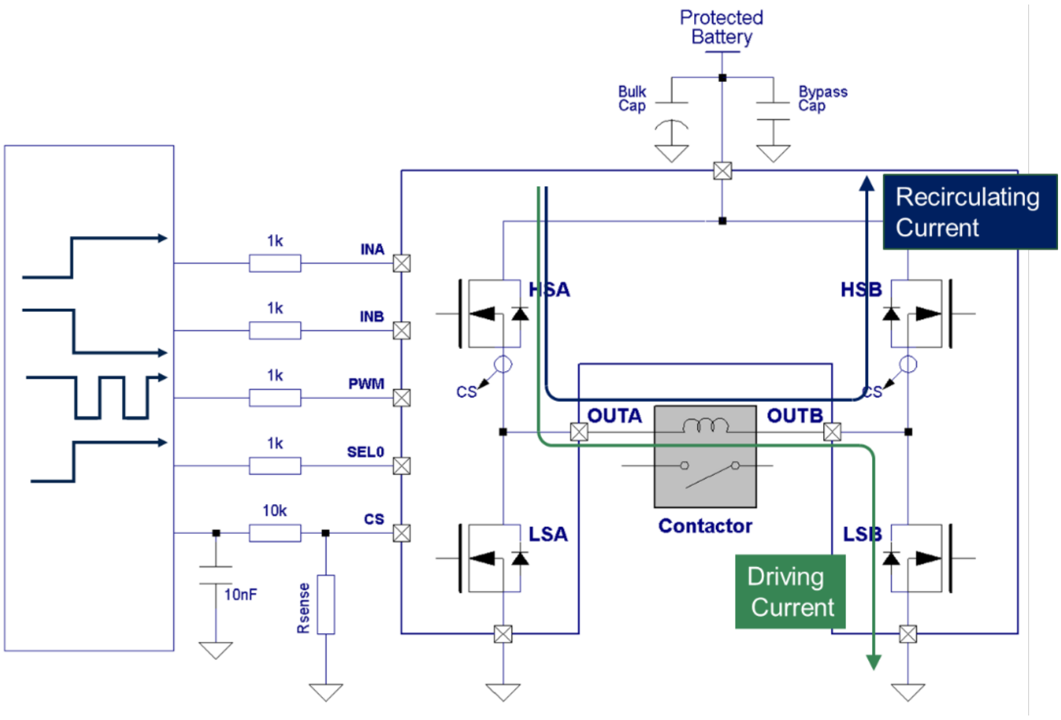
To calculate the shortest OFF time, the lowest coil inductance and the highest coil resistance should be used. This hold time can be calculated based on battery voltage as it impacts the slow recirculation time. Adding the 90% can ensure that the hold current minimum will not be violated.

Alternatively, a simpler solution could be just enabling PWM immediately after pull-in, skipping the “slow recirculation” time altogether. Doing so just slows the ramp down from the pull-in current to the hold current.

3.3 Hold current

During the hold time the low side switch (LSB) is in PWM to reduce the current in the primary windings, see the Figure 7. The micro, using the coil resistance and the battery voltage, calculates the duty cycle. The coil inductance is heavily relied upon to regulate the primary current during PWM.

Figure 9. Hold current flow



PWM duty cycle is not as simple as the ratio between the battery voltage and the hold voltage. The rate of change of current in an inductor is proportional to the voltage across its inductance. So, the duty cycle calculation must include this difference between the ON and recirculation circuits.

Basic inductor equation

$$V_{Ind} = L\frac{\Delta I}{\Delta T} \quad (2)$$

3.3.1 Hold current PWM duty cycle calculation

Looking at the Figure 6, the voltage across the coil inductance is affected by the current through the coil resistance as well. So, V(inductor) is not just the battery voltage. When there is current in the coil, the inductance sees the battery voltage minus the voltage drop across the internal resistance.

PWM ON time calculation

$$V_{Batt} - I_{Hold}R_{Coil} = L\frac{\Delta I}{t_{ON}} \quad (3)$$

When the coil is recirculating, the inductor sees the same voltage drop across the coil resistance plus the voltage drop across the recirculating element. In the case of the VNH7100BAS that is the HSB body diode (V_f = 0.7 V).

PWM OFF time equation

$$V_f + I_{Hold}R_{Coil} = L\frac{\Delta I}{t_{OFF}} \quad (4)$$

Understanding that the duration in recirculation (t_{OFF}) is just the remaining time left in the PWM period after t_{ON}.

The relationship between t_{OFF} and t_{ON}

$$t_{OFF} = \frac{1}{f_{PWM}} - t_{ON} \quad (5)$$

Putting these three equations together, and a little hand waving, it is possible to calculate an expected duty cycle:

Duty cycle calculation

$$Duty = \frac{V_f + I_{Hold}R_{Coil}}{V_{Batt} + V_f - I_{Hold}R_{HBr}} \quad (6)$$

3.3.2 Hold current tolerance

Looking at Eq. (6) it is put in evidence that the duty cycle is independent of frequency (f_{PWM}), inductance (L), and ripple current amplitude (ΔI). There are two parameters that do make a difference in the calculated duty cycle: they are the supply voltage (V_{Batt}) and the coil resistance. The supply voltage is typically known and can be used when calculating the appropriate duty cycle from Eq. (6). Coil resistance is not dynamically known and may impact to maintain current error.

3.3.2.1 Coil resistance

The coil resistance is very dependent on the coil temperature. As a result, coil temperature affects the actual coil current at a given PWM duty cycle. Unfortunately, coil temperature is not so easily determined. Some estimations can be given if ambient temperature is provided. Also in automotive applications, the charging system voltage is typically a function of ambient temperature. However, getting to that level of complexity may not be needed.

From the temperature coefficient of copper resistivity, it is possible to estimate what range of coil resistance to expect. The temperature coefficient of copper resistivity is:

Resistivity of copper

$$\alpha = 0.393\%/°C \quad (7)$$

The equation to calculate the difference in the coil resistance over temperature is (given that the reference resistance is at 25 °C):

Coil resistance over temperature

$$R_{Coil} = R_{Coil@25°C}[1 + \alpha(T_{Coil} - 25°C)] \quad (8)$$

Variation in the hold current due to coil resistance can be calculated by solving Eq. (6) for coil current. Then varying the coil resistance from -40 °C to 125 °C as well as accommodating for the +/-5% tolerance of the coil resistance specification. This leads to the worst case variation in coil current.

Coil current calculation

$$I_{Hold}(R_{Coil}) = \frac{V_f - Duty(V_{Batt} + V_f)}{R_{Coil} + Duty \cdot R_{HBr}} \tag{9}$$

3.3.3 Ripple current

Frequency and inductance influence the ripple current amplitude. To be sure that the ripple current is not excessive, it is possible to calculate the ripple current amplitude for a given inductance and PWM frequency.

Ripple current calculation

$$\Delta I = \frac{V_{Batt} - I_{Hold}(R_{Coil} + R_{HBr})}{L_{Coil}} \frac{Duty}{f_{PWM}} \tag{10}$$

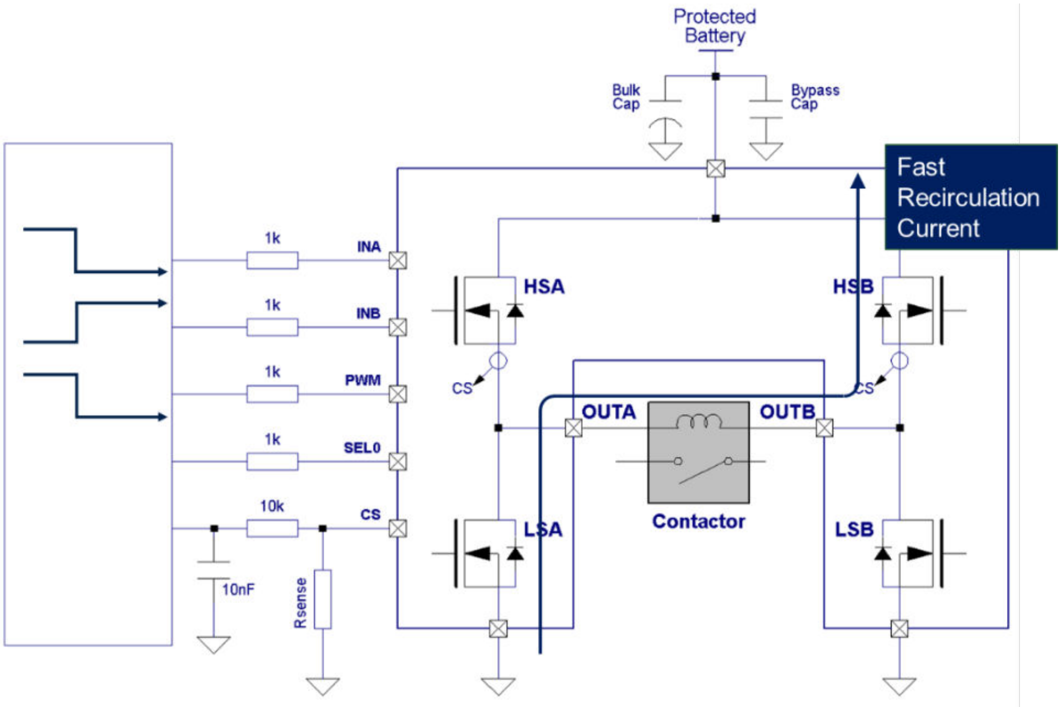
Most automotive applications require a 20 kHz or greater PWM frequency to limit audible noise. The worst-case condition would be at the minimum duty cycle, which would be at max battery (V_{batt} = 16 V).

Anything less than 10% would be more than sufficient to keep the contactor reliably closed (and quiet) as long as the current is always above the minimum required.

3.4 Fast recirculation

When the contactor is disabled, the polarity of the primary coil is reversed due to inductive flyback. The simplest way is to reverse the polarity of the inputs (INA and INB) and set the PWM pin low. This turns off all the low side drivers and turn-on HSB. The inductance of the coil automatically inverts the polarity across the coil and current flow from the ground to the supply. This remains as long as there is current due to the coil inductance.

Figure 10. Fast recirculation path



The VNH7100BAS has a cross conduction protection that delays switching polarity between HS and LS switches. During the cross-conduction delay time (t_{cross} < 350 μs) the entire bridge is tri-stated. It may be that this cross-conduction delay time is longer than the fast recirculation time. This does not inhibit the fast recirculation. During this tri-stated condition, the current just flows through the body diodes of the LSA and HSB switches.

Two trials were implemented to verify the correct behavior of the contactor when driven by the VNH7100BAS.

The contactor that has been used to implement the measurements shows a coil inductance L = 52 mH with a series resistance R_{series} = 11 Ω.

In Figure 11 the fast decay after the hold phase is implemented. Setting a hold current of about 500 mA (modulating the PWM with a duty cycle of 32%), the fast decay time is less than 8 ms: in the Figure 12, CH2 and CH6 show that the contactor is open before the condition I_{coil} = 0 A is reached in the primary windings.

Figure 11. Hold phase and fast decay time

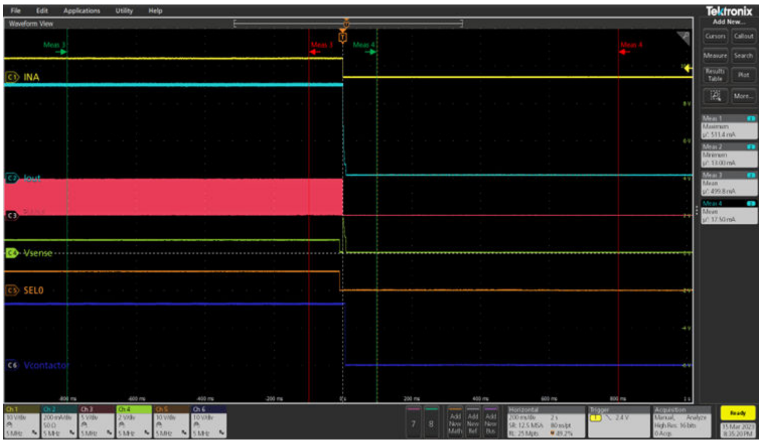
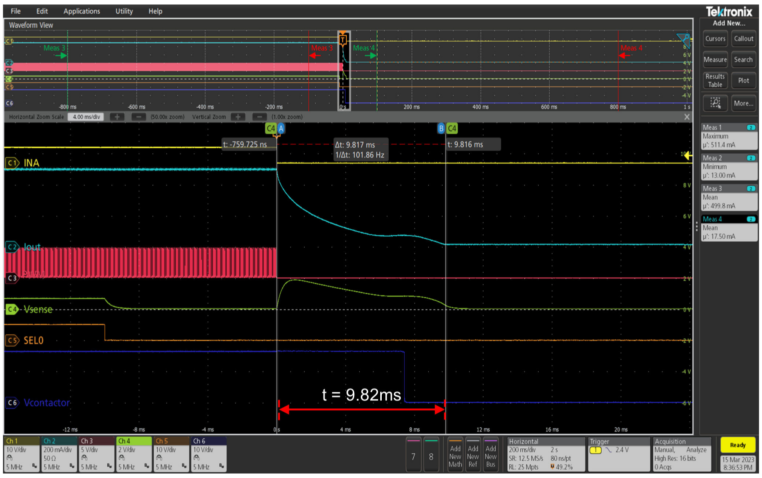


Figure 12. Zoom Hold phase and fast decay time



Fast decay phase has been implemented also during the pull-in phase, simulating a worst-case condition where a fault occurs immediately after contactor closure. In this case, the device drives the contactor with a DC current of about 1.5 A. This trial is shown in the Figure 13, where it is put in evidence that the device works correctly being able to sustain the required demagnetization energy. In the Figure 14, it is shown a fast decay time around 10 ms.

Both cases put in evidence that the actual clamping method provided by the VNH7100BAS is more than sufficient for operating this contactor.

Figure 13. **Pull-in phase and worst-case fast decay time**

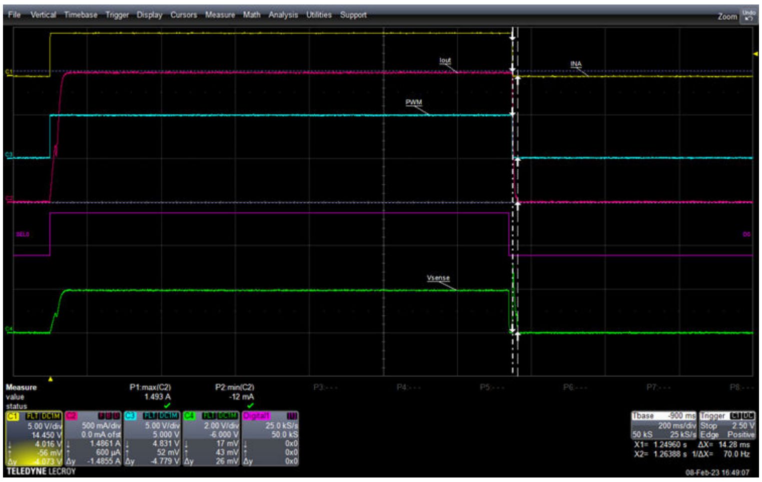
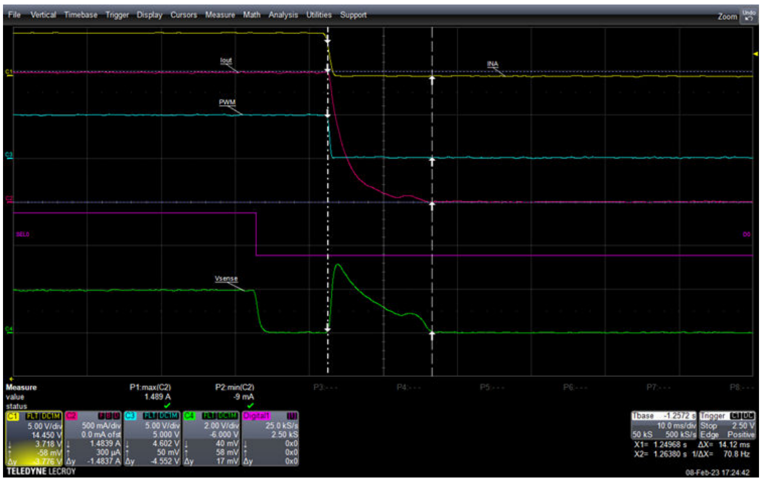


Figure 14. **Zoom worst-case fast decay time**



3.4.1 Protection against loss of battery

During fast recirculation, the current is flowing from ground to supply. During this time, there needs to be a place for the current to go. Typically, that means a path back to the battery. If the current path returning to the battery is interrupted or blocked by a reverse battery protection circuit, then the current finds another path. That usually means the fast recirculation voltage climbs higher until it finds an alternate path back to ground. There are a few options for that recirculation path. Fast recirculation current can be shunted to an external component like a bulk capacitor or a TRANSIL (zener) or it can break the H-Bridge by exceeding the breakdown voltage of the part.

The first line of defense is typically a bulk capacitor. When just using a capacitor, the energy in the primary coil is dumped or transferred into the bulk capacitor. This energy transfer looks like:

Bulk capacitor energy equation

$$\frac{1}{2} L_{Coil} I_{Coil}^2 = \frac{1}{2} C_{Bulk} \Delta V_{Bulk}^2 \quad (11)$$

Where:

- L_{Coil} = coil inductance
- I_{Coil} = coil current at loss of battery
- C_{Bulk} = bulk capacitance (see Figure 3)
- ΔV_{Bulk}= the rise in voltage across the bulk capacitor

The upper limit to the voltage on the capacitor is based on the breakdown voltage of the H-Bridge. The VNH7100BAS absolute maximum supply voltage rating is $V_{CC(max)}$ = 38 V. So, the range of voltage on the bulk capacitor starts at the battery voltage and ends at $V_{CC(max)}$.

Bulk capacitor max voltage change

$$\Delta V_{Bulk} < V_{CC(max)} - V_{Batt} \tag{12}$$

Simplifying the Eq. (11) and inserting the Eq. (12) it is possible to find the bulk capacitor equation as follows:

Bulk capacitance calculation

$$C_{Bulk} > \frac{I_{Coil}^2 L_{Coil}}{\left(V_{CC(max)} - V_{Batt}\right)^2} \tag{13}$$

The worst-case condition is where the coil current is the highest, during the pull-in phase. The pull-in current is essentially the battery voltage divided by the coil resistance.

If the bulk capacitance required to protect the H-Bridge is too excessive (too large or expensive), then either use a sufficiently robust TRANSIL to keep the voltage below 38 V or a combination between a TRANSIL and capacitance. Bulk capacitance is needed anyway on a module for supply stability.

A TRANSIL has resistance associated with it. So, be careful in choosing a TRANSIL that has an initial clamping voltage that is above a double battery (jump start) and below the max V_{CC} of the VNH7100BAS when conducting the peak coil current.

4 Controlling the VNH7100BAS using a microcontroller

The VNH7100BAS can be controlled entirely by a microcontroller. It takes up to five pins from the micro. 4 digital I/O and one analog input (ADC). The VNH7100BAS I/O works with either a 5 V or a 3.3 V system. Two pins are optional depending on the level of monitoring required. These would be the SEL0 pin and the current sense (CS) pin.

The current sense pin (CS) has two functions. The first is to reflect the selected high side output current reduced (divided) by a K factor. The second is to report a fault where half of the bridge is latched off.

The current sense feedback output requires a sense resistor (Rsense in Figure 3). The resistor value is based on the ADC max voltage and the k-factor reduced coil current. Details on how the current sense feedback works and how to calculate the proper sense resistor size can be found in AN5026.

Fault indication by the current sense pin is further described in detail in the application note AN5026.

The table below summarizes the steps in each phase of contactor operation.

Table 1. VNH7100BAS truth table

State	INA	INB	PWM	SEL0	Comment
Pull-in	H	L	H	H	Current flowing from HSA to LSB, monitoring HSA current
				L	
Slow recirc	H	H	X	H	Current flowing from HSA to HSB, monitoring HSA current
				L	Current flowing from HSA to HSB (LSB OFF)
	H	L	L	H	Current flowing from HSA to HSB, monitoring HSA current
				L	Current flowing from HSA to HSB (LSB OFF)
Hold	H	L	20 kHz	H	Current flowing from HSA to LSB/HSB at PWM frequency, monitoring HSA current
				L	Current flowing from HSA to LSB/HSB at PWM frequency, monitoring HSB current
Fast recirc.	L	H	L	X	Current flowing from LSA to HSB, only HSB is ON

Related links

See AN5026:VIPower M0-7 H-Bridges for Automotive DC Motor Control

5 Designing the micro independent solution

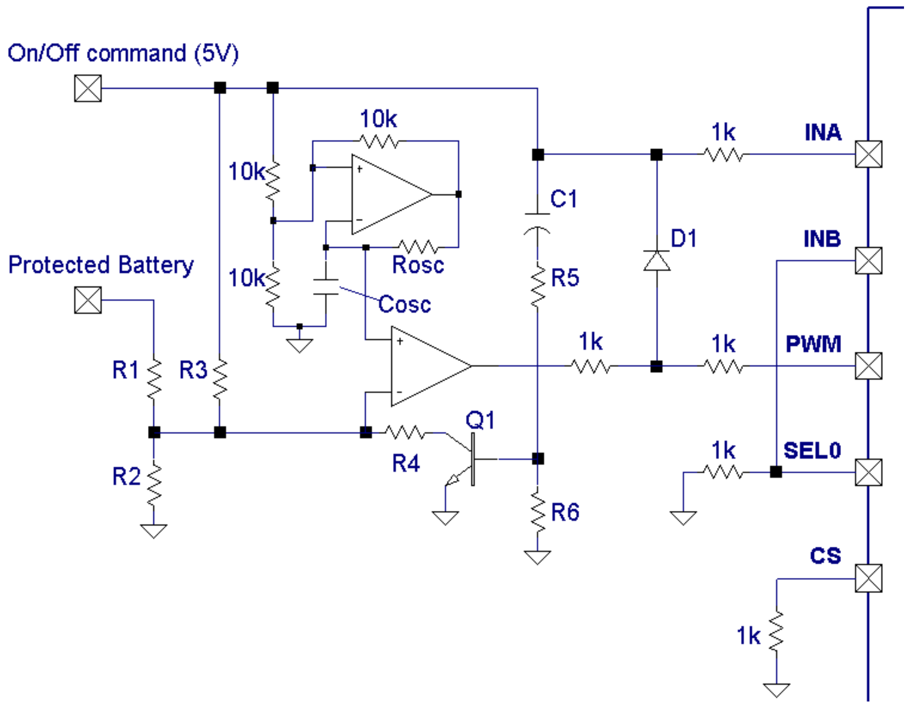
The VNH7100BAS can be used to drive a contactor without the advantage of a microcontroller. There are a few functions or features that are lost when forgoing the microcontroller. They are:

1. Fault reporting
2. Positive diagnostic feedback such as load integrity checking.
3. Accuracy in both timing and current regulation

The functions that must be recreated are the pull-in timing, the coil voltage control, and the recirculation control (slow/fast). The current regulation benefits from the current requirements shown in Section 5.2. The current is not flat over the temperature range. This is mostly due to the change in resistance of the coil.

During pull-in the PWM circuit can either be completely overridden or set to provide a pull-in level of current regulation. Slow recirculation occurs naturally while fast recirculation is a result of completely tri-stating the outputs.

Figure 15. **Standalone PWM hold current control circuit**



The limitations of this standalone circuit include:

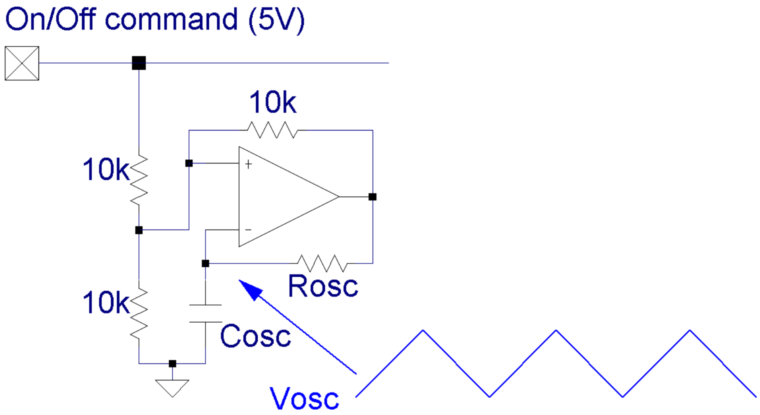
- No analog feedback of regulated current for diagnostics.
- No off time between pull in and hold currents. The system switches directly from pull-in mode to hold PWM mode. This changes the current profile between pull-in and hold considerably.
- The accuracy of the regulated current is limited to the tolerance of the resistors and the stability of V_{IN}.

5.1 Oscillator

The oscillator is a simple 1/3 – 2/3 oscillator. It generates a simple triangle waveform starting at 1/3 of the supply and ending at 2/3 of the supply. This keeps the input voltages well within the common-mode range for most op-amp and the voltage slope to be fairly linear.

Three of the four oscillator resistors are simple 10 kΩ resistors. This makes for the basic 1/3 - 2/3 thresholds. When the output of the op-amp is low, the voltage on the + terminal is 1/3. When the output in the op-amp is high, the voltage on the + terminal is 2/3.

Figure 16. Simple 1/3-2/3 oscillator



The triangle waveform shown for Vosc (Figure 16) is generated by the basic charge–discharge of C_{OSC} by R_{OSC}. The basic equations for the voltage of an RC circuit can be described by:

Charge and discharge CR equations

$$V_{discharge}(t) = V_{max} e^{-t/RC} \quad or \quad V_{charge}(t) = V_{max}\left(1 - e^{-t/RC}\right) \tag{14}$$

Using these two equations and describing the voltage from 1/3 to 2/3 and back the oscillation frequency can be calculated as follows:

Oscillator frequency equation

$$f_{OSC} = \frac{1}{2 R_{OSC} C_{OSC} \ln 2} \tag{15}$$

Setting C_{OSC} and solving for R_{OSC}.

Oscillator frequency

$$R_{OSC} = \frac{1}{2 f_{OSC} C_{OSC} \ln 2} \tag{16}$$

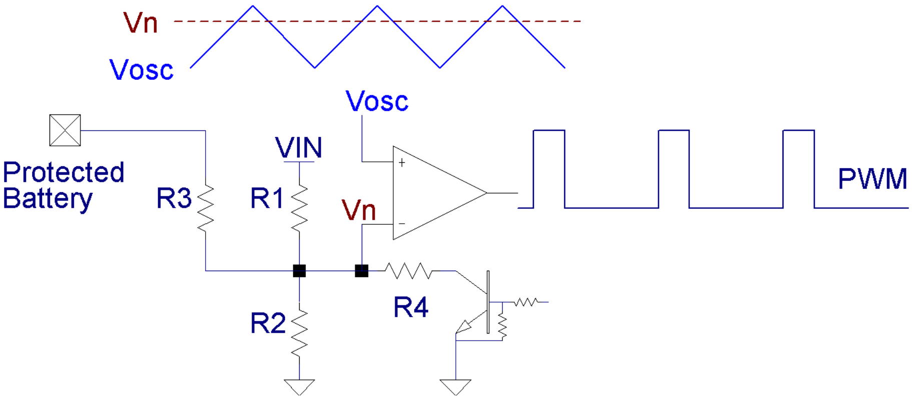
The frequency equation is simple in both case due to the symmetry of the triangular waveform in the oscillator above, and does not depend on the supply voltage.

5.2 Coil voltage control

The coil driving requirements are voltage centric. This means that this circuit needs to provide a steady voltage over the operating voltage range.

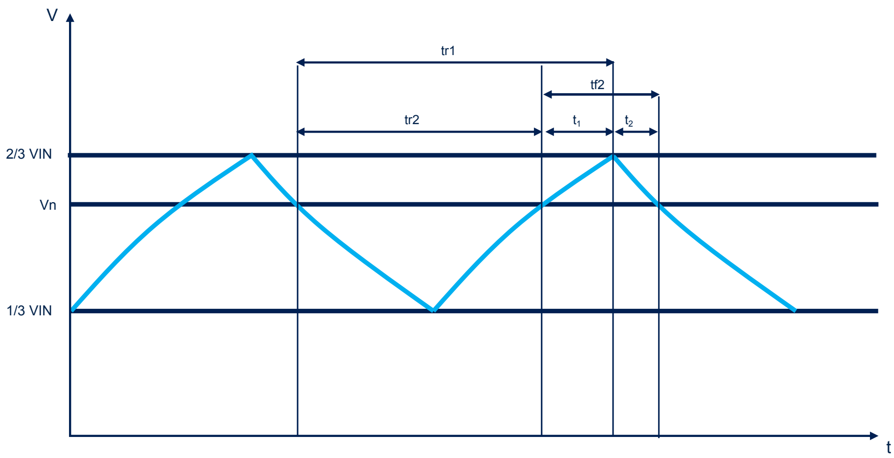
The voltage control circuit compares the supply voltage with the fixed triangle waveform generator and adjusts the duty cycle inversely proportional to the battery voltage. It is not a true RMS converter. As a result, the duty cycle is initially determined by the Eq. (6) while being controlled by the battery voltage.

Figure 17. **PWM generator circuit**



First, it is necessary to find the voltage (Vn) that generates the correct duty cycle at 9 V and at 16 V (by using Eq. (6)). Having these bounds, it is possible to figure out the resistor values in the Figure 17.

Figure 18. **Dissecting the oscillator waveform**



The equation for determining duty cycle requires calculating known sections of the theoretical RC waveform and subtracting other portions of the waveform. For instance, the Figure 18 illustrates that the rising edge duration above the red dashed line (Vn) can be described as:

Rising edge duration

$$t_1 = t_{r1} - t_{r2} \tag{17}$$

And the falling edge duration above the red dashed line (V_n) can be described as:

Falling edge duration

$$t_2 = t_{f2} - t_1 \tag{18}$$

All of the time above the V_n line is the ON time for the PWM generator. Such that the duty cycle decreases as the battery voltage increases. This can be expressed as:

Duty cycle

$$Duty = (t_1 + t_2) f_{OSC} \tag{19}$$

These can all be expressed in terms of the basic RC rise and fall equations shown in Eq. (14). With that, the duty cycle generated by a specific V_n can be described:

Duty cycle with respect to the voltage at V_n

$$Duty\left(V_n\right) = f_{OSC}R_{OSC}C_{OSC}\ln\left(\frac{2V_{IN}}{V_n} - 2\right) \tag{20}$$

V_n needs to be translated back to the battery voltage at the coil. So, starting with a simple nodal equation from Figure 17 of the control voltage.

PWM control node equation

$$\frac{V_{IN} - V_n}{R_1} + \frac{V_{Bat} - V_n}{R_3} = \frac{V_n}{R_2} \tag{21}$$

Solving Eq. (21) for V_n.

V_n with respect to V_{Bat}

$$V_n = \frac{R_2R_3V_{IN} + R_1R_2V_{Bat}}{R_1R_2 + R_1R_3 + R_2R_3} \tag{22}$$

Between Eq. (20) and Eq. (21), it is possible to generate an equation for the duty cycle as a function of the battery voltage.

Duty cycle with respect to V_{Bat}

$$Duty\left(V_{Bat}\right) = f_{OSC}R_{OSC}C_{OSC}\ln\left(\frac{2V_{IN}}{\frac{R_2R_3V_{IN} + R_1R_2V_{Bat}}{R_1R_2 + R_1R_3 + R_2R_3}} - 2\right) \tag{23}$$

Setting R_2, taking Eq. (22), inserting $V_n(Duty)$ for Vn and solving for R_1:

Calculating R_1

$$R_1 = \frac{R_2R_3\left(V_{IN} - V_n(Duty)\right)}{R_3V_n(Duty) - R_2\left(V_{Bat} - V_n(Duty)\right)} \tag{24}$$

Setting the Eq. (24) to equal itself, with one side using the duty cycle needed at 9 V to generate V_n and the other side using the duty cycle needed for 16 V to generate V_n (for $V_n(Duty)$ see Eq. (6)), solving for R_3 and simplifying gets:

Calculating R_3

$$R_3 = -\frac{R_2}{V_{IN}}\frac{V_n(Duty_{9V})(V_{IN} - 16V) + V_n(Duty_{16V})(9V - V_{IN}) + V_{IN}(16V - 9V)}{V_n(Duty_{9V}) - V_n(Duty_{16V})} \tag{25}$$

Where V_{IN} = the input voltage (it is better to have this fixed to a known value (5 V)).

5.3 Pull-in control

The pull-in function has two possible parameters: pull-in duration and pull-in current regulation (if desired).

5.3.1 Pull-in duration

The pull-in duration timing is governed by C_1, R_5, R_6, and the V_{BE} of Q_1 shown in the Figure 15. This circuit has a lot of over temperature variations. However, pull-in time can vary widely (within reason) without adverse effects on the function or reliability of the contactor. Only the minimum pull-in time needs to be respected.

The current through R_6 is fixed by the V_{BE} of Q_1. The current through R_5 is determined by the V_{BE} and the charging of C_1. With that, it is possible to calculate the time it takes to charge up capacitor C_1 to where the V_{BE} of Q_1 is no longer sufficient to keep Q_1 ON. At that point, there is no longer any base current to turn on Q_1. The Eq. (6) finds the point where R_6 takes all of the charging current from C_1.

Pull-in timing equation

$$\frac{V_{BE}}{R_6} = \frac{(V_{IN} - V_{BE})e^{-t_{Pull-in}/R_5C_1}}{R_5} \tag{26}$$

Solving Eq. (26) for the pull-in time gives us:

Pull-in timing equation

$$t_{Pull-in} = R_5C_1\ln\frac{R_6(V_{BE} - V_{IN})}{R_5V_{BE}} \tag{27}$$

The shortest time to charge the capacitor (C_1), to where V_{BE} is no longer respected, is when V_{BE} is the highest. That depends on the ambient temperature and transistor used. In the collector current range under consideration (6 mA-7 mA), a simple BC848B NPN small signal bipolar transistor has a V_{BE} range of ~450 mV (at hot) to ~850 mV (at cold).

Setting R_5 and R_6 to be equal and solving for R_5 finds the optimum resistance needed to stay within the bounds of the pull-in time.

Solving for R5

$$R_5 = \frac{t_{Pull-in}}{C_1\ln\frac{(V_{BE} - V_{IN})}{V_{BE}}} \tag{28}$$

5.3.2 Pull-in current regulation

The pull in current can be completely unregulated between 10 V and 16 V. A simple pulldown of the voltage at V_n can force the PWM generator to work at 100% duty cycle. This makes selecting the value of R_4 simple. It can be 0 Ω. However, if you would like to reduce the duty cycle above a certain voltage, then R_4 can be calculated.

Using Eq. (21) keeping R_1 and R_2 fixed and setting V_n at the peak (1/3 V_{IN}) of the oscillator triangle waveform, it is possible to calculate what R_2 needs to be. From that it is put in evidence that this new value for R_2 is a parallel of the previously calculated R_2 and the introduced resistor, R_4.

PWM control node equation with R_2 and R_4 in parallel

$$\frac{R_2R_4}{R_2 + R_4} = \frac{V_{IN}R_1R_3}{3V_{Bmax}R_1 - V_{IN}R_1 + 2V_{IN}R_2} \tag{29}$$

Solving for R_4:

PWM control node equation for R_4

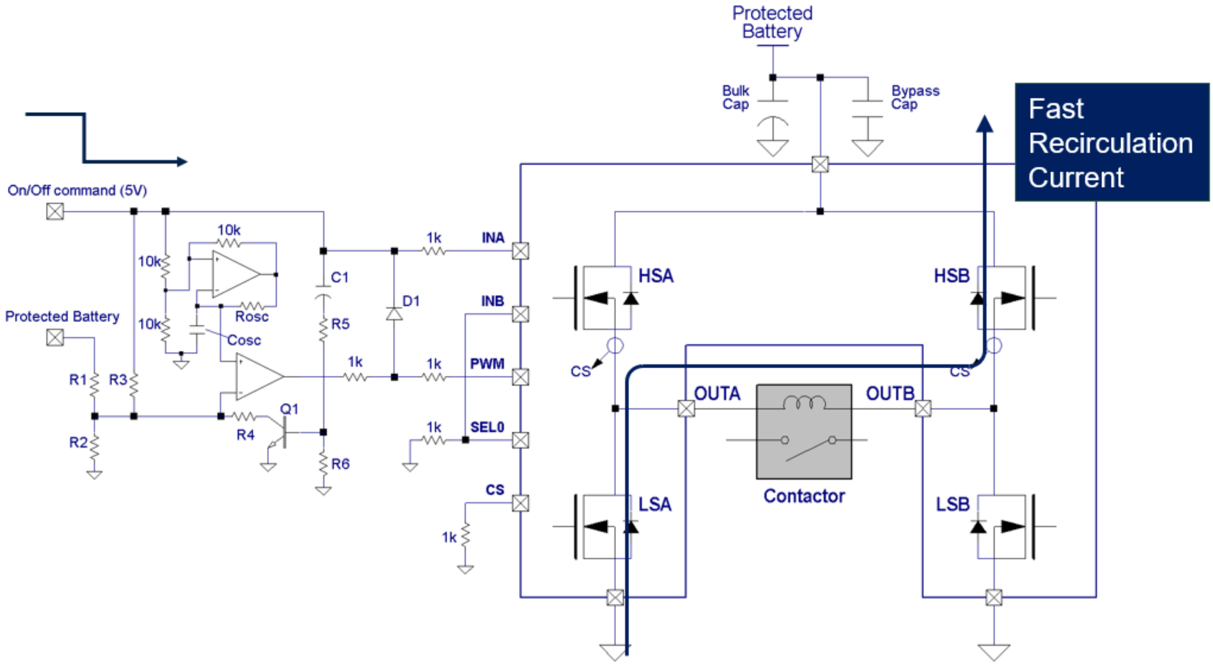
$$R_4 = \frac{V_{IN}R_1R_2R_3}{3V_{Bmax}R_1R_2 - V_{IN}(R_1R_2 + R_1R_3 - 2R_2R_3)} \tag{30}$$

Using this value for R_4 causes the PWM engine to start PWMming at $V_{Bmax.}$ or whatever voltage you want to start PWMming the coil.

5.4 Turning off the contactor

To disable the contactor, the On/Off command pin goes low. This pulls the INA and PWM pin to the ground. INB is already low so this action completely tristates the bridge and forces the current to flow from LSA to HSB, generating a fast turn-off.

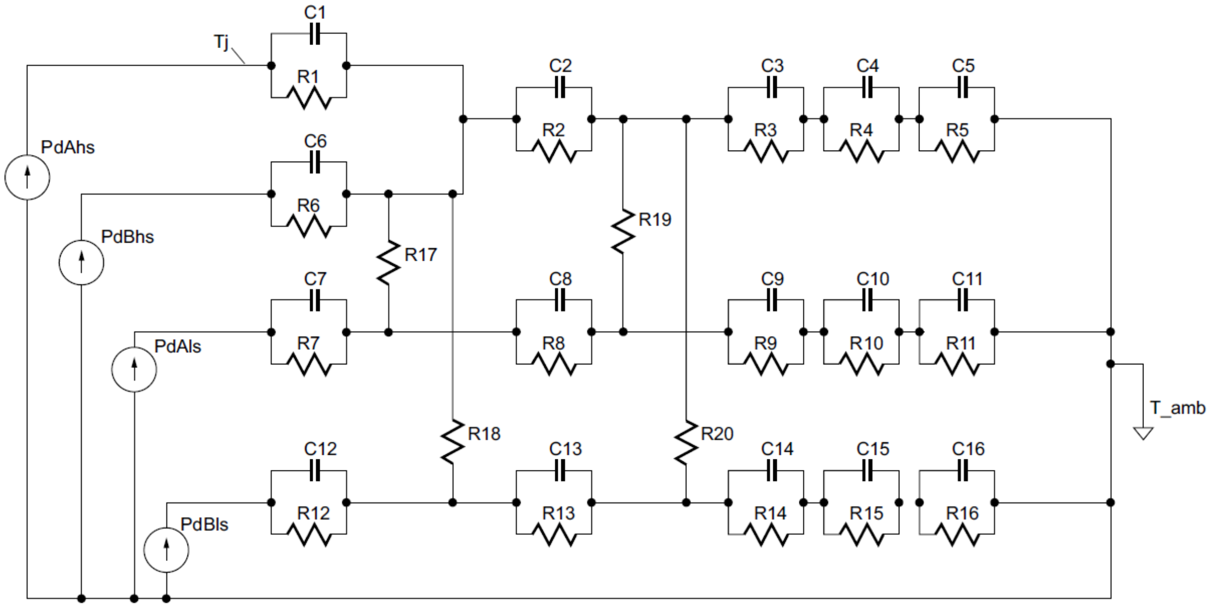
Figure 19. Fast recirculation at turn-off



6 Power dissipation

There is a number of ways to determine if the VNH7100BAS can handle the power it dissipates while driving the contactor. There is a multi-element Foster thermal model in the datasheet illustrating the complexity of the VNH7100BAS thermal paths. This model can be incorporated into SPICE where the output structures are emulated either by equation or by schematic.

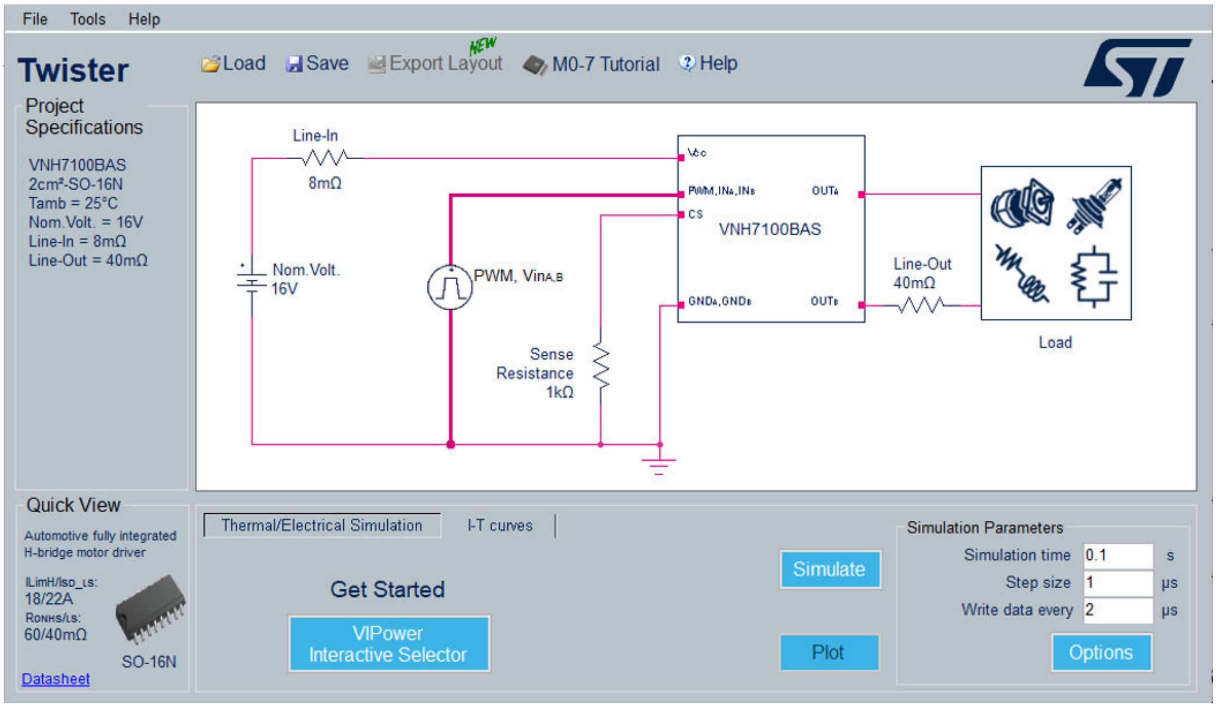
Figure 20. Foster thermal model for the VNH7100BAS



The above (Figure 20) model is incorporated into a thermal simulation tool provided by STMicroelectronics called TwisterSIM. TwisterSIM is available by download from the ST.com website. You can download the register TwisterSIM for free.

Since the pull-in time is short, the thermal time constants play a large role in determining the junction temperature. At least for pull-in it is recommended that the TwisterSIM be used. It provides the most accurate thermal modeling. This would also apply to the fast recirculation period.

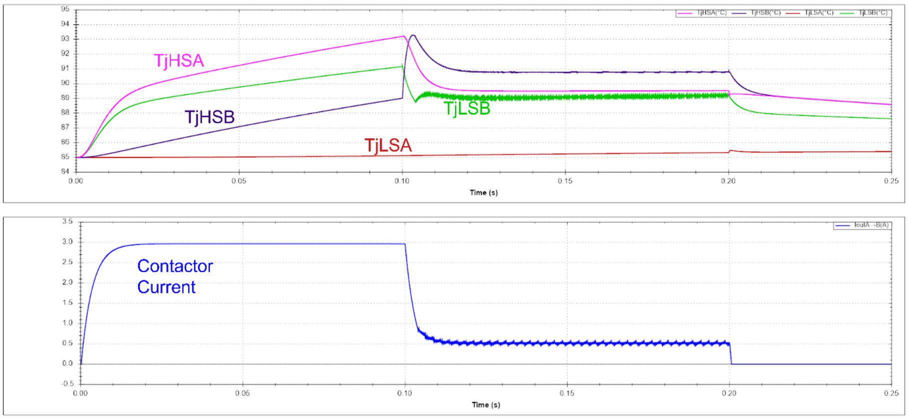
Figure 21. TwisterSIM



We can spend a lot of time calculating the average power dissipation by the various switching elements in the different stages of contactor driving. However, the volume of calculations needed is extensive and still doesn't match the accuracy of TwisterSIM.

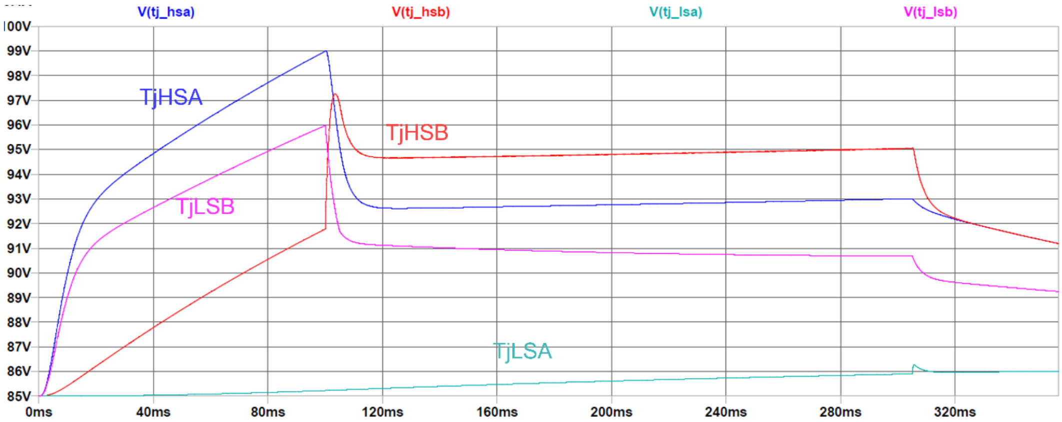
The simulator does not emulate the coil perfectly in that there is no BEMF generated by the movement of the armature. That lack of BEMF does little to hinder the accuracy of the thermal simulation.

Figure 22. Twister thermal simulation results



SPICE model thermal simulation results correlated quite well to the twister model.

Figure 23. SPICE model thermal simulation



In both simulations, the junction temperature rise is minimal. Long-term (SPICE) simulations indicate that the maximum junction temperature is less than 100 °C, a 15 °C rise.

Revision history

Table 2. **Document revision history**

Date	Revision	Changes
27-Mar-2023	1	Initial release.

Contents

List of figures