TORQUE CONTROLLER
BRUSHLESS DC MOTOR DRIVER MODULE
100V/40A, 250V/40A

FEATURES:

- True Four Quadrant Complementary Switching.
- Fully integrated 3-Phase Brushless DC Motor Control Subsystem includes power stage, non-isolated driver stage, and controller stage
- MOSFET Output Stage, with low Rdson.
- 40A peak Phase Current with 10V to160V Maximum Bus Voltage
- Internal Precision Current Sense Resistor
- Cycle by cycle current limiting.
- Fixed frequency PWM.
- Closed-loop SERVO Control of Motor, Current Loop Closed Internally
- Direction and motor current are controlled through command polarity and value.
- Tacho output with frequency proportional to speed
- Enable/Disable input
- Direction Output
- Adjustable Current Limit
- On board –11.5V supply
- On board +5V supply
- Hermetic or non-hermetic packaging available (3.10" x 2.10" x 0.385")

APPLICATIONS:

- Servo positioning systems
- Actuation systems
- Hoists

DESCRIPTION:

SMCT6MXX-XX is a completely self-contained 4-quadrant motor controller that converts an analog input command signal into a motor current. The motor current is internally sensed, processed, and used as an output for the closed current loop feedback control. Error amplified input and output are available for closed loop compensation. Proportional or proportion/integral (PI) compensation can be implemented to optimize motor and system performance. SMCT6MXX-XX is best used as a four-quadrant torque controller for controlling servo systems.

The SMCT6MXX-XX is a fully integrated three-phase brushless DC motor control module housed in a 43 Pin power flatpack. Many integral control features provide user flexibility in adapting the SMCT6MXX-XX to specific system requirements. Current control or closed loop speed control can be easily implemented.

The small size of this complete motor control module makes it ideal for commercial aerospace and military applications.

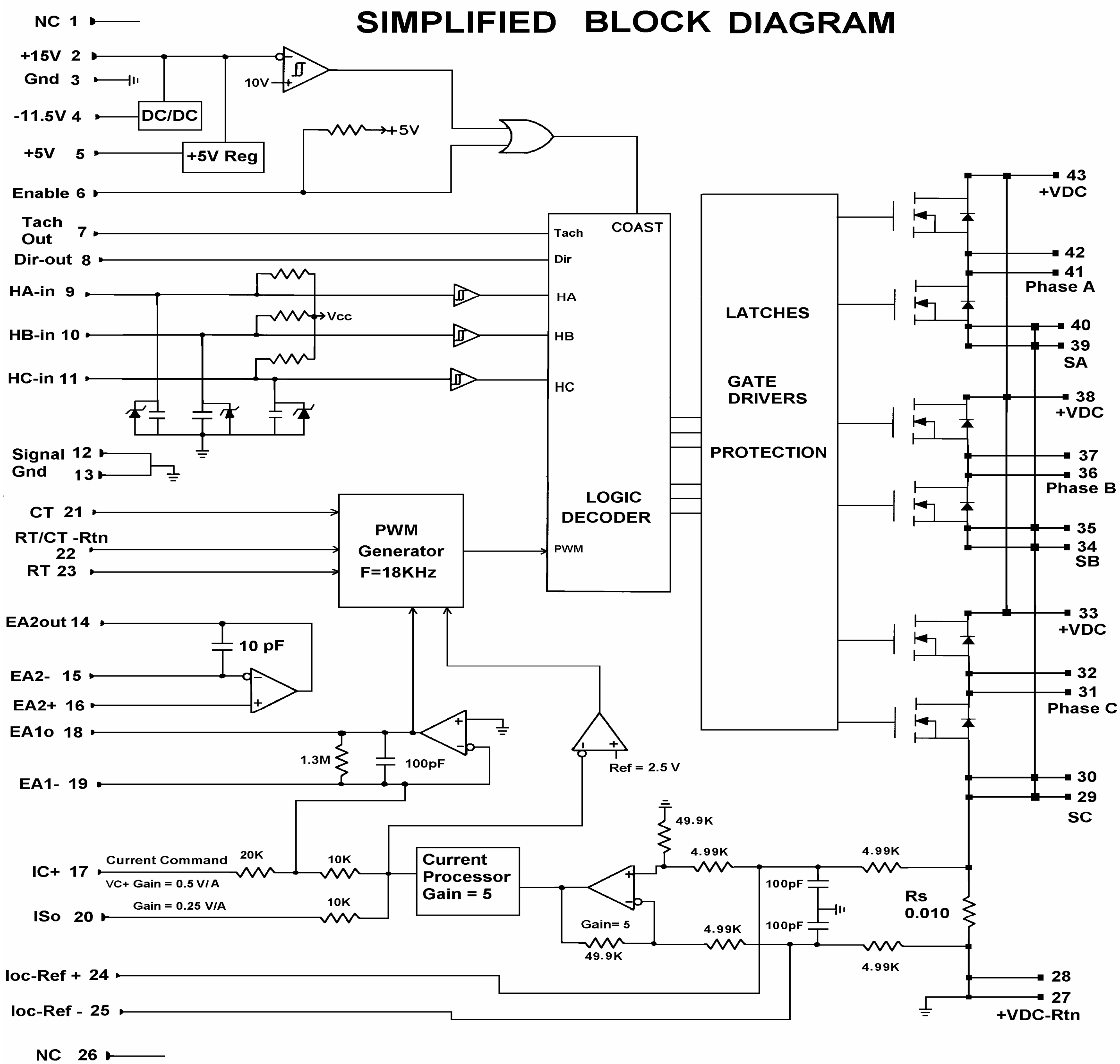


Fig. 1: Block Diagram

1- Switching frequency is internally set to 18 kHz, typically.
2- It is preferred not to have external connection between Signal Gnd at Pins 3, 12, and 13, and +VDC Rtn at Pins 27 and 28.
3- Over-current limit is internally set to 10A peak.

COMMUTATION TRUTH TABLE

This table shows the Phase Output state versus the state of the Hall-Effect and Command Polarity Inputs. The commutation coding shown reflects Hall-Effect sensors that are spaced at 120^o mechanical increments. Also, internal protection logic disables all three Phase Outputs when the Hall-Effect Inputs are set to an illegal condition (i.e., all logic low or all logic high).

Hall Sensors			Command Polarity =Positive Forward Rotation			Command Polarity =Negative Reverse Rotation		
			PHASE OUTPUTS			PHASE OUTPUTS		
HA	**HB**	**HC**	**PhA**	**PhB**	**PhC**	**PhA**	**PhB**	**PhC**
1	0	1	Source	Sink	Hi-Z	Sink	Source	Hi-Z
1	0	0	Source	Hi-Z	Sink	Sink	Hi-Z	Source
1	1	0	Hi-Z	Source	Sink	Hi-Z	Sink	Source
0	1	0	Sink	Source	Hi-Z	Source	Sink	Hi-Z
0	1	1	Sink	Hi-Z	Source	Source	Hi-Z	Sink
0	0	1	Hi-Z	Sink	Source	Hi-Z	Source	Sink
0	0	0	Hi-Z	Hi-Z	Hi-Z	Hi-Z	Hi-Z	Hi-Z
1	1	1	Hi-Z	Hi-Z	Hi-Z	Hi-Z	Hi-Z	Hi-Z

ABSOLUTE MAXIMUM RATINGS

(T_C=25 ^{o}C) unless otherwise noted

Characteristic	Maximum
Maximum Peak DC Bus Supply Voltage SMCT6M40-10, SMCT6M40-25	 100 V 250 V
Maximum Operating DC Bus Supply Voltage SMCT6M40-10 SMCT6M40-25	 60 V 150 V
RMS Output Current for SMCT6M40-10, T_C=90^{o}C for SMCT6M40-25, T_C=90^{o}C	40 A 30A
Peak Output Current for SMCT6M40-10, T_C=90^{o}C for SMCT6M40-25, T_C=90^{o}C	50 A 40A
+15V Supply Voltage	+15.5 V
-11.5V Output Load Current	30mA
+5V Output Load Current	30mA
Logic Input Voltage	-0.3 V to +5.5 V
Error Amplifier Input (EA1-)	+/- 10 V
Error Amplifier Output Current	±8 mA
Spare Amplifier Input Voltage (EA2+/EA2-)	+/- 10 V
Spare Amplifier Output Current	±8 mA
Operating & Storage Junction Temperature	-55^{o}C to +150^{o}C
Power Devices Thermal Resistance R_{thjC}	0.60 ^{o}C/W

ABSOLUTE MAXIMUM RATINGS (continued)

(T_C=25 ^oC) unless otherwise noted

Characteristic	Maximum
Pin-to-Case Voltage Isolation for SMCT6MXX-XX for SMCT6MXX-XX-1	 600V DC 1000V DC
Lead Soldering Temperature, 10 seconds maximum, 0.125" from case	300°C

RECOMMENDED OPERATING CONDITIONS

(T_C=25 ^oC) unless otherwise noted

Characteristic	Maximum
Operating Supply Voltage SMCT6M40-10 SMCT6M40-25	 50 V 130 V
RMS Output Current for SMCT6M40-10, T_C=80^{o}C for SMCT6M40-25, T_C=80^{o}C	40 A 30A
Peak Output Current for SMCT6M40-10, T_C=80^{o}C for SMCT6M40-25, T_C=80^{o}C	50 A 40A
+15V Supply Voltage	+ 14 V +/-5%

PARAMETER SYMBOL CONDITIONS [(1)]	MIN.	TYP.	MAX.	UNITS
Power Output Section				
Drain-Source Leakage Current IDss at 0.8VDss Diode Reverse Recovery Time trr I_F = 40A, di/dt = -100A/usec Diode Forward Voltage V_F at I_F = 40 A SMCT6M40-10-1 SMCT6M40-25-1			250 300 1.0 1.4	uA nSec V V
Drain-to-Source On-Resistance Rds(on) I_D =40A [(2)] Per switch SMCT6M40-10-1 SMCT6M40-25-1			 5 35	 mΩ
Control Section				
Input Supply Current Icc at +14V supply without any external load on –11.5V or +5V -11.5V Output Range +5V Output Range	 -11.0 4.7		80 -12.0 5.3	mA V V

RECOMMENDED OPERATING CONDITIONS (continued)

(T_C=25 °C) unless otherwise noted

PARAMETER SYMBOL CONDITIONS [1]	MIN.	TYP.	MAX.	UNITS
Error Amplifier EA1				
Output Offset Voltage Amplifier Output Voltage Range	-20 -11	- -	20 11.5	mV V
Spare Amplifier Sections EA2				
Output Offset Voltage, VCM=0V Amplifier Input Common-mode Voltage Range	-5 -8	- -	5 10	mV V
Current-Sense Amplifier Section				
Amplifier Voltage Gain Out DC Offset	0.24 -10	0.25 0	0.26 +10	V/A mV
Current Limit Set Point Voltage Corresponding Current Limit with Rs=0.010 Ohm	+/-2.3 -	+/-2.5 +/-10.0	+/-2.7 -	V A
Logic Input Section				
HA, HB, HC, and EN High Level Input Voltage Threshold Low Level Input Voltage Threshold	 2.0 -	 - -	 - 1.0	 V V
Tachometer				
Tachometer Output High Level Voh Tachometer Output Low Level Vol Tachometer Frequency Ratio To Hall Input Frequency	4.5 - 3	4.8 - 3	5.2 200 3	V mV -
Oscillator Section				
Oscillator Frequency f_S	16	18.0	20.0	kHz

NOTES:

1. All parameters specified for Ta = 25°C, Vcc = +14Vdc, and all Phase Outputs unloaded. All negative currents shown are sourced by (flow from) the Pin under test.
2. Pulse Test: Pulse Width < 300 μSec, Duty Cycle < 2%.

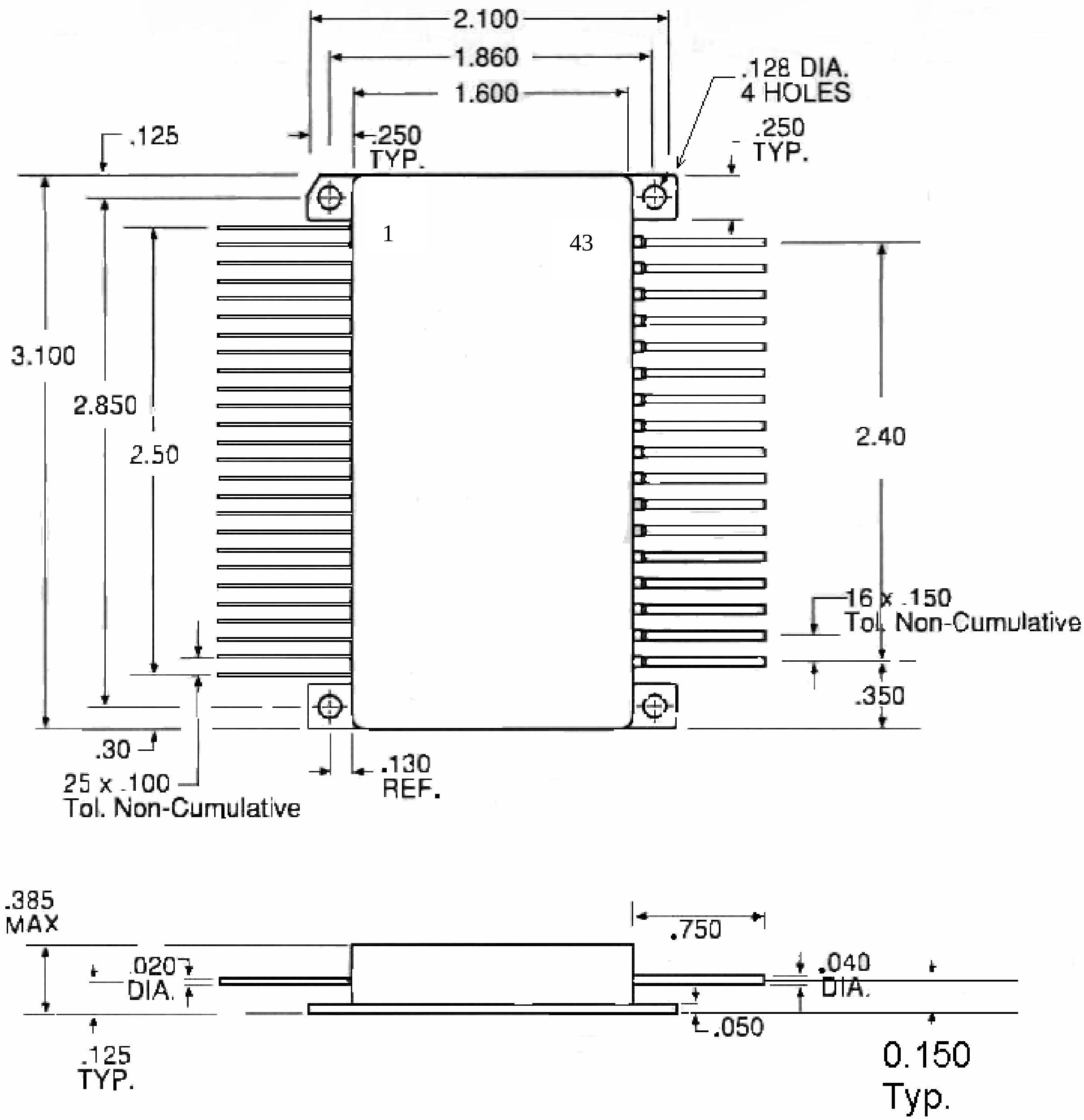


Fig. 2: Mechanical Outline for Hermetic Package, SMCT6MXX-XX

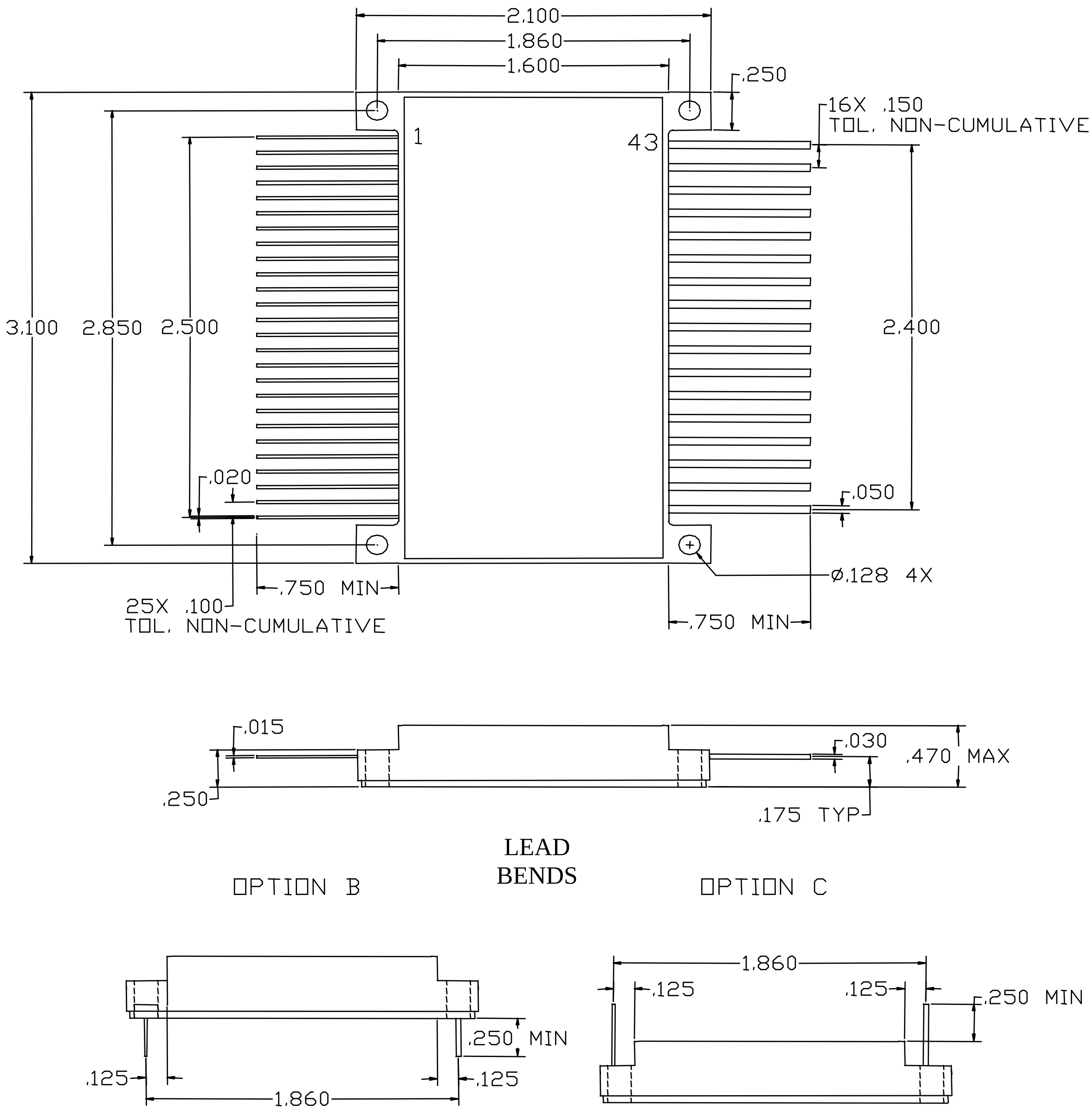


Fig. 3: Mechanical Outline for Non-Hermetic Package, SMCT6MXX-XX-1

PIN OUT

PIN NUMBER	NAME	DESCRIPTION
1	NC	Not connected
2	+15V Input	The +15V power supply connection for the controller. Under-voltage lockout keeps all outputs off for Vcc below 9 to 10.5V. The return of +15V is Pin 3. **The input current requirement is 80mA without any external loads on Pins 4 and 5**. Recommended input range is 14V min, 15.0V max.
3	Signal Gnd	Return for +15V supply, and –11.5V, +5V outputs
4	-11.5V Output	-11.5V **Output**. The return of –11.5V is Pin 3. The maximum output current is 30mA. This Pin should be by-passed to Gnd with 3-5uF capacitor. The range of this output is -11V to –12.0V.
5	+5V Output	+5V **Output**. . The maximum output current is 30mA. The return of +5V is Pin 3. This Pin should be by-passed to Gnd with 3-5uF capacitor. The range of this output is 4.7V to 5.3V.
+15V supply should be an isolated power supply		
6	Enable Input	Digital input that disables all outputs once pulled low. This input is internally pulled high to +5V by a 10kΩ resistor. This input can be used as an enable/disable input using an open collector switch. If the switch is opened, the controller is enabled. If the switch is closed to Gnd, the controller is disabled.
7	Tachometer Output	Variable frequency output proportional to the motor speed. The output frequency is three times that of hall sensors. The pulse duty cycle is 50%. There are 3 pulses every 360 electrical degrees. The number of pulses per motor revolutions is P*3/2. The Tachometer output frequency is $ft = \frac{P.n}{40}$ Hz Where P is the number of poles, n is the motor speed in rpm.
8	Direction Output	A logic output. High corresponds to motor CW rotation. Low corresponds to motor CCW rotation.
9	HA Input	Hall input of Phase A
10	HB Input	Hall input of Phase B
11	HC Input	Hall input of Phase C

PIN OUT (continued)

12	Signal Gnd	Reference ground for all control signals of the device. All bypass capacitors and loop compensation components must be connected as close as possible to Pins 12 and 13. This ground is internally connected to the +VDC Rtn. It is preferred not to have external connection between Signal Gnd and +VDC Rtn at Pins 27 and 28.
13	Signal Gnd	
14	EA2 Output	Output of the spare amplifier
15	EA2- Input	Inverting input of the spare amplifier
16	EA2+ Input	Non-inverting input of the spare amplifier
17	IC+	Current Command input. The input command gain is 0.5 V/A.
18	EA1 Output	Output of the error amplifier and is internally connected to the PWM comparator.
19	EA1- Input	Error amplifier inverting input. This input is available for current loop PI loop calibration.
The error amplifier non-inverting input, EA1+, is internally connected to signal ground.		
20	Iso	Output of the current sense amplifier for external monitoring. This output is internally feeding an over-current limit circuit. It is recommended to have the over-current limit 20-30% higher than the target peak motor current. The gain of Iso is internally set to 0.25 V/A.
21	C_T	PWM oscillator programming capacitor in nF. Insert a capacitor between this Pin and Pin22 to reduce switching frequency. Use a high quality capacitor for best results. The voltage on C_T is symmetrical about the signal ground. The peak voltage is internally set to +/-**7.2V.** The voltage across C_T is internally applied to the PWM comparator.
22	R_T/C_T- Rtn	Return Pin for the PWM oscillator programming capacitor at Pin 21, and oscillator programming resistor at Pin 23.
23	R_T	PWM oscillator programming resistor in K ohms. Insert a resistor between this Pin and Pin22 to increase switching frequency. $f_S = \frac{200(10+R_T)}{10R_T(C_T+1)}.$ KHz Where R_T is in K Ohm, and C_T is in nF
Refer to Fig.9 for R_T, C_T connections		

PIN OUT (continued)

24	Ioc-Ref +	Over-current limit reference adjustment.
25	Ioc-Ref -	Over-current limit adjustment. Connect a resistor Rg KOhms between Pins 24 and 25 to decrease the current amplifier gain and increase peak current limit. The current amplifier gain attenuation due to Rg will be $Kc = \frac{Rg}{Rg + 4.99}$ The output signal gain at Pin 20 will be $0.25 {}^{*} K_C$ V/A. The command input gain at Pin17 will be $2/K_C$ A/V
26	NC	Not connected
27, 28	+VDC Return	Motor supply DC bus return.
29, 30, 34, 35, 39, 40	Source C	These pins are the source terminals of the three arms of the three-phase bridge. These Pins shall be shorted together externally using a low impedance bus to minimize power loss, as shown in Fig. 18.
31, 32	Phase C Output	Phase C terminals. Both terminals shall be used.
33 38 43	+VDC	These pins are the motor input power supply positive terminal. These Pins shall be shorted together externally using a low impedance bus. +VDC bus should be bypassed to +VDC Rtn with adequately voltage-rated low ESR capacitor.
36, 37	Phase B Output	Phase B terminals. Both terminals shall be used.
41, 42	Phase A Output	Phase A terminals. Both terminals shall be used.
Case	NC	Not connected

Application Report

By
Deib A Deib, PhD
Senior Design Engineer
Sensitron Semiconductor

Control Modes and Closed Loops

Typically, speed regulation is achieved by regulating the average input voltage to the motor, while torque regulation is achieved by current control. Voltage and current control loops may be combined to achieve a specific speed-torque performance.

SMCT6MXX-XX can be easily used in many applications, including position control, speed control, or torque control. Fig. 4 shows the block diagram of a servo position control application. In this diagram; current, velocity, and position loops are implemented for best system performance.

Definitions:

TF = Transfer Function
CA = Current Sense Amplifier
E_b = Motor Back EMF
I_m = Average Motor Current
R_m = Motor Resistance
ω_m = Motor Speed in rad/sec
V_S = Oscillator Ramp Peak Voltage
f_{ci} = Current Loop Crossover Frequency
R_S = Current Sense Resistor Value
V_{dc} = DC Supply Voltage
EA = Error Amplifier
K_b = Motor Back EMF Constant
K_T = Motor Torque Constant
L_m = Motor Inductance
n_m = Motor Speed in rpm
f_S = Switching Frequency
f_{cV} = Voltage Loop Crossover Frequency
PM= Phase Margin in degrees

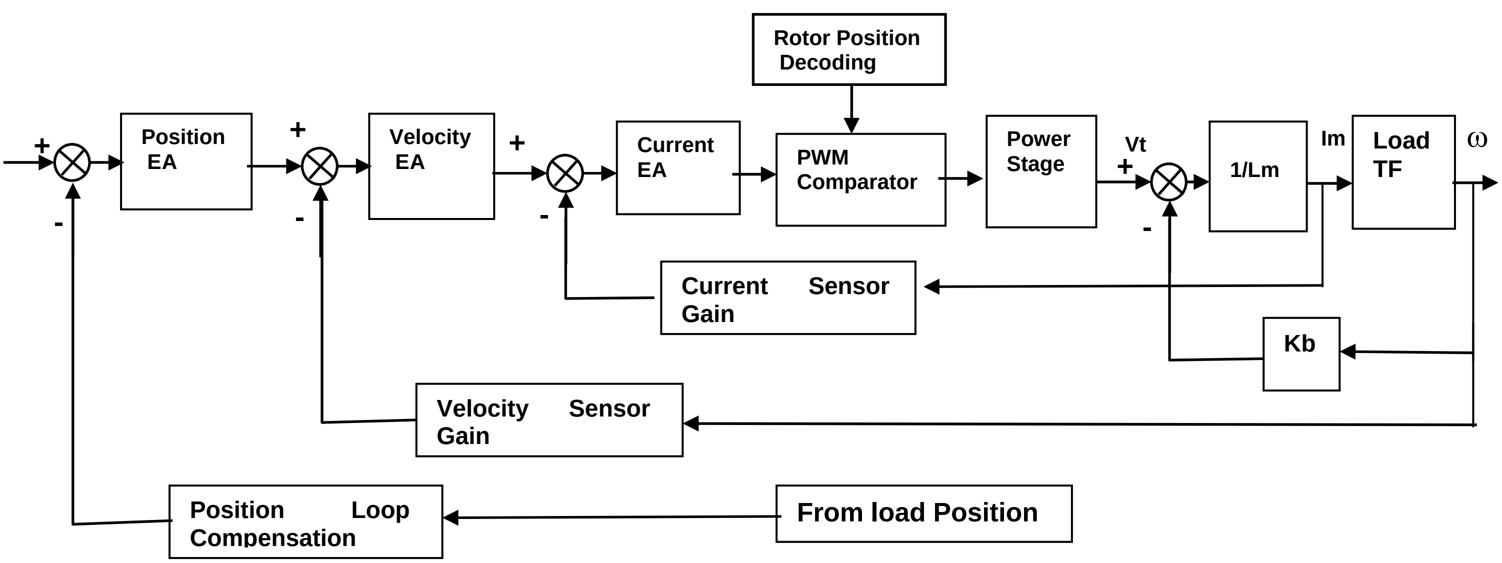


Fig. 4. Typical Closed loop Servo System Block Diagram

Cycle-by-cycle current limiting

The block diagram of the current control loop is shown in Fig1 and Fig. 6.

The current sense resistor R_S is chosen to establish the peak current limit threshold, which is typically set 20-30% higher than the maximum current command level to provide over-current protection during abnormal conditions. Under normal circumstances with a properly compensated current loop, peak current limit will not be exercised.

The input divider network provides attenuation, with Rg selected to accommodate the current command signal range. Connect a resistor Rg K Ohms between Pins 24 and 25 to decrease the current amplifier gain and increase peak current limit. The current amplifier gain attenuation Kc will be

$$Kc = \frac{Rg}{Rg + 4.99} \qquad 1$$

The internal current sense resistor Rs is 0.010 Ohm. This resistor is rated up to 20A maximum. For higher current operation an external sense resistor shall be added between Pins 27, 28 and Pins 29, 30.This will be in parallel with the internal 0.010 Ohm.

The output signal gain at Pin 20 will be

$$G_O = 25 R_S K_C \ \text{V/A} \qquad 2$$

Go is internally set to 0.25 V/A .

The command input gain at Pin17 will be

$$G_C = \frac{1}{50 R_S K_C} \ \text{A / V} \qquad 3$$

Gc is internally set to 2 A / V

The peak current limit Ip is

$$I_P = \frac{2.5}{25 R_S K_C} \ \text{A} \qquad 4$$

Ip is internally set to 10 A with Rg open (Kc=1), and no external current sense resistor.

In four quadrant operation and bidirectional current sensing, the current sense amplifier input (I_{RS}) and output (I_{SO}) are shown in Fig. 5.

The current sense amplifier filter time constant is internally set to less than 1 usec. It is necessary to have limited filtering by the current sense amplifier to maintain the high speed of the over-current limit comparator.

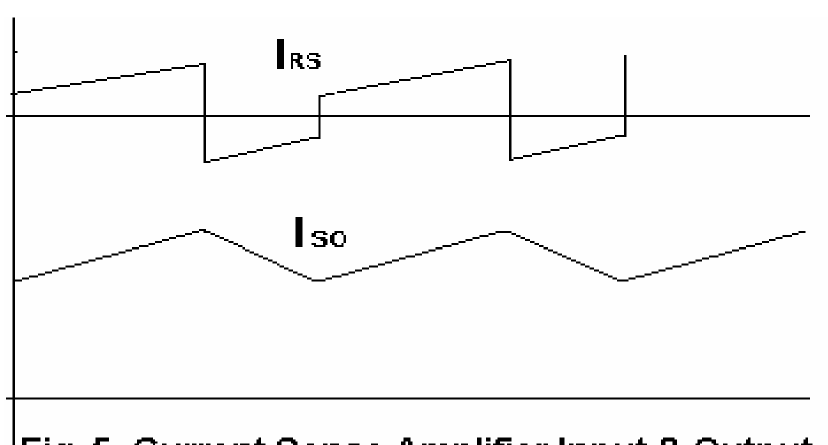


Fig. 5. Current Sense Amplifier Input & Output

Average Current Mode Control for BDC Motor Torque Controller

Figs. 6 and 9 show the implementation of a typical torque control loop. A voltage command proportional to the desired current is applied at Pin 17.

The current feedback signal is obtained by the internal bidirectional current sense amplifier.

Small signal compensation of the feedback control loop is provided by an internal error amplifier, EA1, and external RC components R4, R5, C2, & C3.

The error amplifier output, EA1o, is internally connected to one input of the PWM comparator, while the oscillator ramp across C_T is connected to the other input.

Since the torque is proportional to the average phase current, the torque is controlled via duty cycle control.

EA1 Transfer Function

Referring to Fig.6, Let

C_P be the equivalent capacitance of C1 and C2 in parallel.

R_P be the equivalent resistance of the R3 and R4 in parallel.

The transfer function of the error amplifier EA1 is

$$G_{EA1}(s) = \frac{V_{EA1}(s)}{V_{CA}(s)} = \frac{R_P(R_5C_3s+1)/R_1}{\left[R_PR_5C_PC_3s^2+(R_PC_3+R_PC_P+R_5C_3)s+1\right]} \qquad 5$$

or,

$$G_{EA1}(s) = \frac{K_1(s/\omega_{ZC}+1)}{(s/\omega_{PC1}+1)(s/\omega_{PC2}+1)} \qquad 6$$

Where

$$K_1 = \frac{R_P}{R_1} \qquad 7$$

$$\omega_{ZC} = \frac{1}{R_5C_3} \qquad 8$$

ω_{PC1} and ω_{PC2} can be calculated by equating the denominator of equation 5 to zero.

If $R_P >> R_5$, ω_{PC1} and ω_{PC2} can simplified by

$$\omega_{PC1} \approx \frac{1}{R_P(C_P+C_3)} \qquad 9$$

$$\omega_{PC2} \approx \frac{(R_p+R_5)}{R_PR_5C_P} \qquad 10$$

The error amplifier transfer function Bode plot is shown in Fig. 10. The EA1 gain at ω_{ZC} is

$$G_{EA1}(s=\omega_{ZC}) \approx \frac{R_5C_3}{R_1(C_p+C_3)} \qquad 11$$

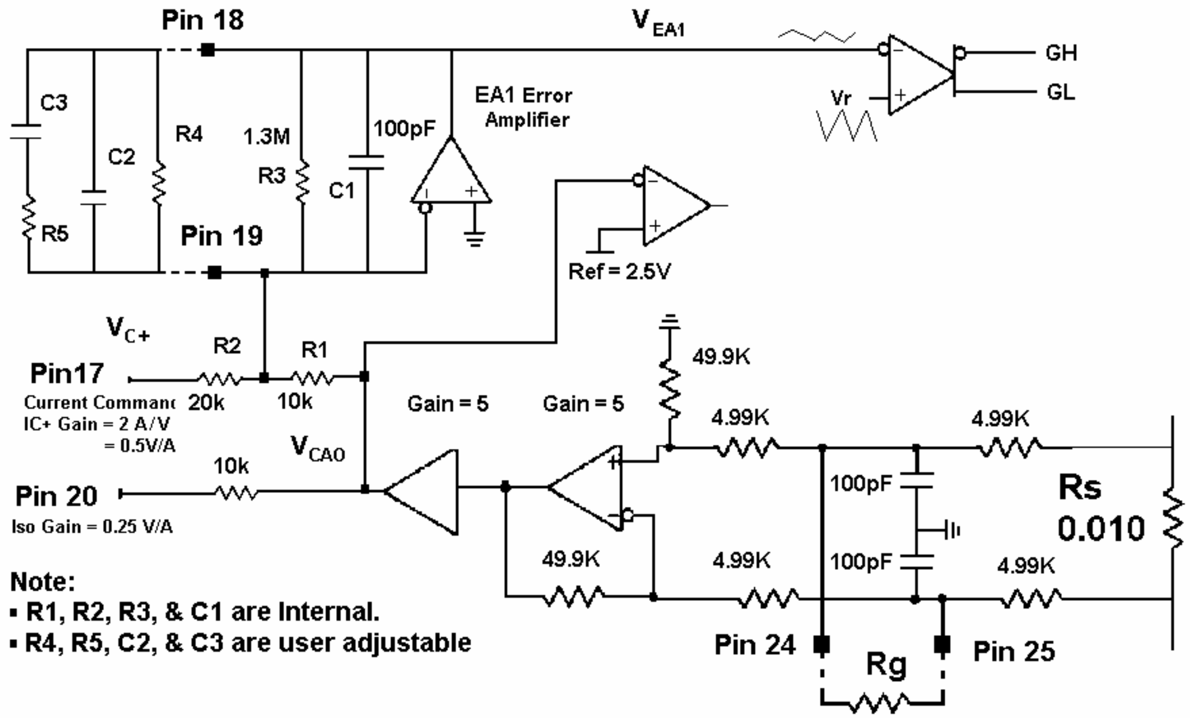
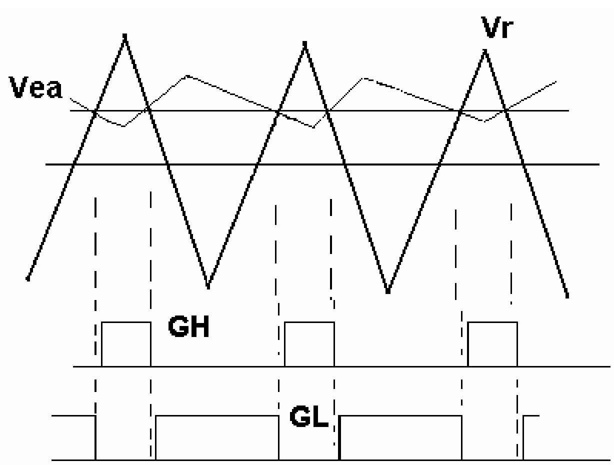


Fig. 6. Current Sense Amplifier & EA1 Compensation Closing Current Loop



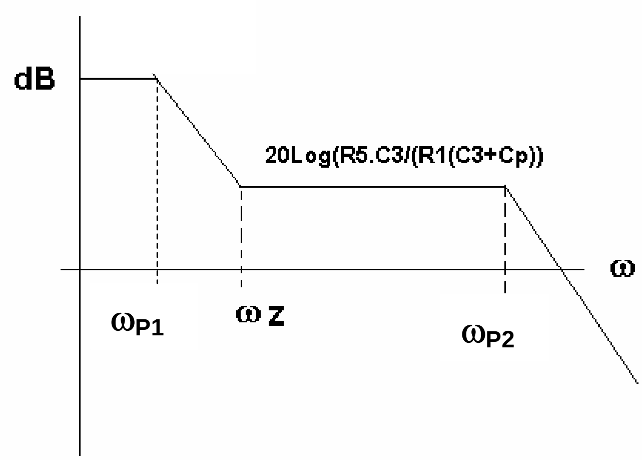


Fig.7. PWM Comparator Inputs & Outputs	Fig.8. Error Amplifier Transfer Function Bode Plot

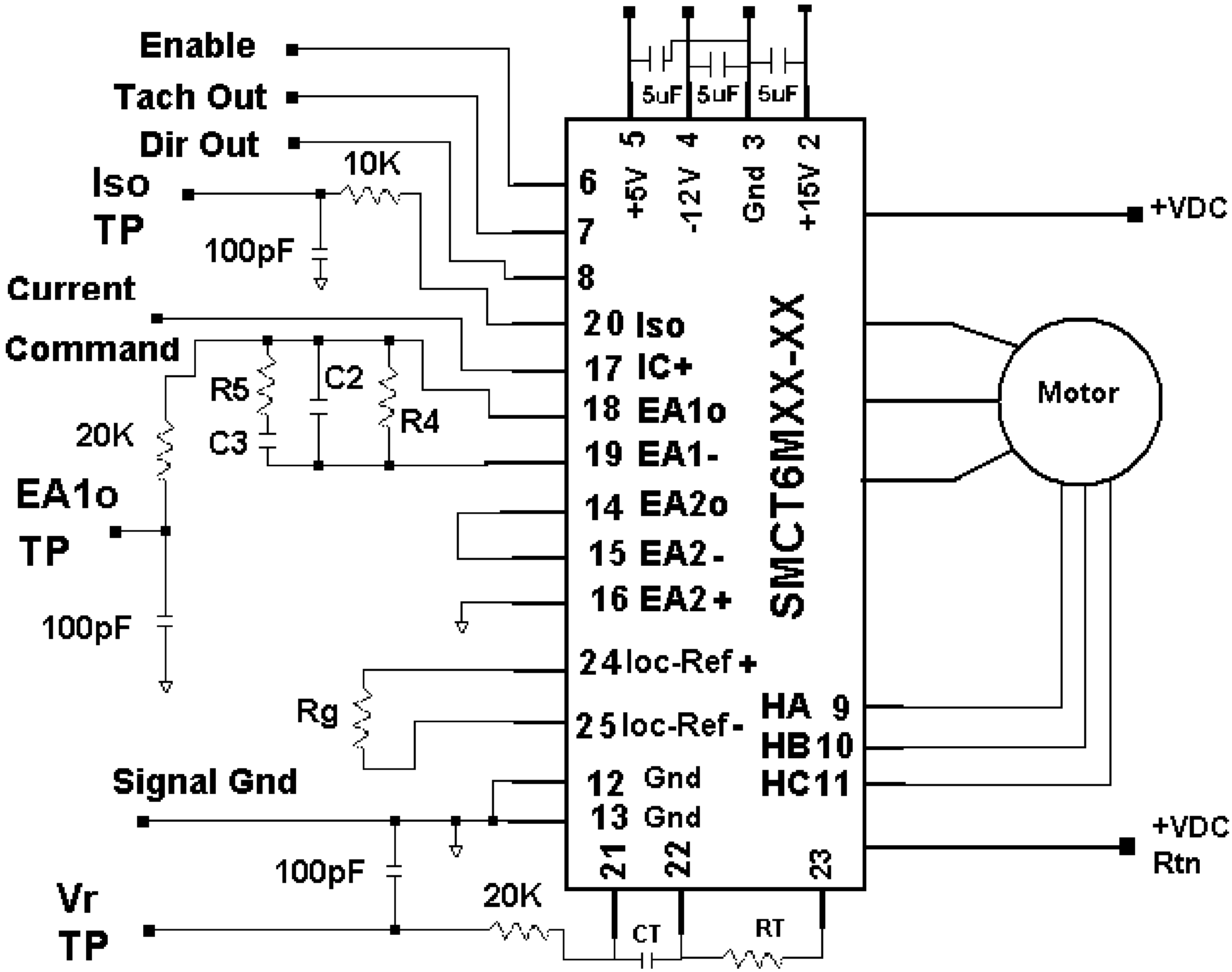


Fig.9 : Closed Loop Current (Torque) Control

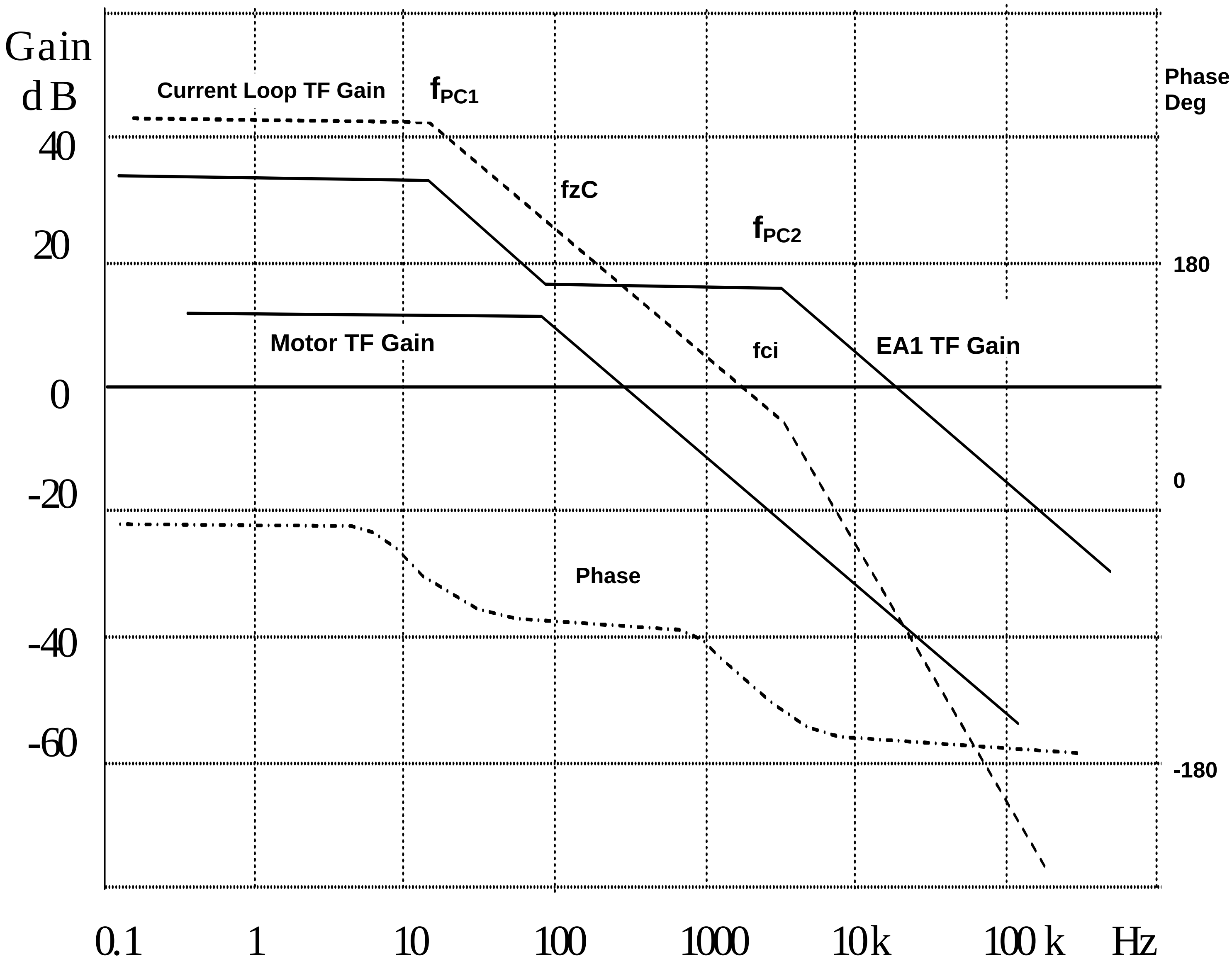


Fig. 10: Current Loop Open Loop Gain & Phase versus Frequency

Motor Power Circuit TF

In the circuit of Fig. 11:

The output of the current sense amplifier is V_{CA},
The output of the current loop error amplifier is V_{EA1}.
PWM duty cycle is D
Motor inductance is L_m, and resistance is R_m
Input DC bus voltage is V_{dc}

The motor current is determined by the PWM duty cycle, motor impedance, and input DC bus voltage.

$$I_m(s) = \frac{DV_{dc}}{sL_m + R_m}, D = \frac{V_{EA1}(s)}{V_S} \qquad 12$$

The current sense amplifier output is

$$V_{CA}(s) = 25K_C R_S \, \mathrm{I_m}(s) \qquad 13$$

Substituting 13 in 12

$$\frac{V_{CA}(s)}{25K_C R_S} = \frac{V_{dc}V_{EA1}(s)}{V_S(sL_m + R_m)} \qquad 14$$

The CA output to control gain G_{Cm} of the motor power circuit is

$$G_{Cm}(s) = \frac{V_{CA}(s)}{V_{EA1}(s)} = \frac{25K_C R_S V_{dc}}{V_S R_m(s/\omega_m + 1)} \qquad 15$$

where

$$\omega_m = \frac{R_m}{L_m} rad/\sec, \; f_m = \frac{R_m}{2\pi L_m} Hz \qquad 16$$

The DC gain of the current loop power section

$$G_{Cm}(s=0) = \frac{25K_C R_S V_{dc}}{V_S R_m} \qquad 17$$

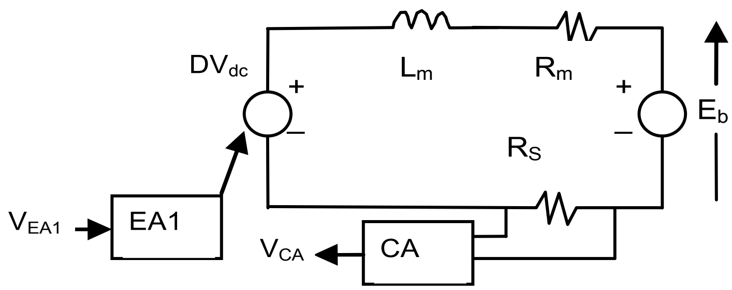


Fig. 11. Current Loop Small Signal Equivalent Circuit

Current Loop Compensation

All PWM circuits are prone to sub-harmonic oscillation if the modulation comparator's two input waveform slopes are inappropriately related. Average current feedback systems will exhibit similar behavior if the current amplifier gain is excessively high at the switching frequency.

A voltage proportional to motor current is generated by the current sense resistor and the current sense amplifier circuitry internal to the SMCT6MXX-XX. This waveform is amplified and inverted by the error amplifier and applied to the PWM comparator input.

To avoid sub-harmonic oscillation, the amplified motor current slope at one input of the PWM comparator must not exceed the oscillator ramp slope at the other comparator input. This criterion sets the maximum error amplifier gain at the switching frequency, and indirectly establishes the maximum current loop gain crossover frequency, limiting f_{Ci} to about $0.1f_S$.

A motor control system typically operates over a wide range of input voltages, and is usually powered from an unregulated supply. The operating conditions which cause the greatest motor current slope must be determined in order to determine the maximum current amplifier gain which will maintain stability.

The current sense amplifier input waveform is rectangular as shown in Fig.5. The maximum value of this waveform is 0.1V, the corresponding maximum output of 2.5V is the over-current limit threshold. The rectangular waveform is converted into a triangular waveform as shown in Fig. 7. The up slope or down slope of the resulting triangular waveform at the EA1 output must not exceed the oscillator ramp slope.

The motor current maximum down slope happens when the motor is spinning at maximum speed, and the command polarity is instantaneously reversed. In this case

At f_{ci}, the EA1 maximum output slope $= \dfrac{R_5 C_3}{R_1(C_p + C_3)} \dfrac{25 K_C R_S 2V_{dc}}{L_m}$ 18

The oscillator ramp slope is:

$$\text{Oscillator Ramp Slope } = \frac{4V_S}{T_S} = 4\, V_S\, f_S \qquad 19$$

Where: V_S, is the oscillator ramp peak voltage (internally set to +/-7.2 V peak-to-peak across C_T)
T_S, is the switching period
f_S, is the switching frequency (internally set to 18kHz).

The maximum EA1 integrator gain at 18 kHz is the gain at which the maximum EA1 output slope equals the oscillator ramp slope. Equating the slopes from 18 and 19 and solving for max G_{EA1} gain:

$$\frac{R_5 C_3}{R_1(C_p + C_3)} = \frac{4V_S\, f_S\, L_m}{25 K_C R_S\, 2V_{dc}} \qquad 20$$

Choose R_5 to be about one half the value given by 20 to accommodate for system variations and worst case conditions. Also, since C_3>>C_P, 20 can be simplified to

$$R_5 = \frac{V_S\, f_S\, L_m R_1}{25 K_C R_S V_{dc}} \qquad 21$$

Using R_1=10 K, f_S =18 KHz, V_S =7.2 V, R_S=0.01, V_{dc} =110 then maximum value for R_5 = 90K

The current loop error amplifier gain is described in equation 5, and the CA output to control gain of the motor power circuit is described in equation 15. The overall open loop gain of the current loop is the product of the current loop error amplifier gain and the output to control gain of the motor power circuit.

The EA1 gain in the flat portion of Fig 8 is described in equation 11. The loop gain crossover frequency, f_{Ci}, should be within the flat portion of the EA1 gain in order to achieve enough phase margin.

The overall open loop gain of the current loop is set equal to one to solve for the loop gain crossover frequency, f_{Ci}, as shown in equation 22.

$$\frac{25K_C R_S V_{dc}}{2\pi V_S L_m f_{C\,i}} \frac{R_5 C_3}{R_1 (C_p + C_3)} = \frac{25K_C R_S V_{dc}}{2\pi V_S L_m f_{C\,i}} \frac{2V_S f_S L_m}{25K_C R_S V_{dc}} = 1 \qquad 22$$

Solving for f_{Ci}, equation 23 will give the maximum value for f_{Ci}

$$f_{C\,i} = \frac{f_S}{2\pi} \qquad 23$$

The motor TF pole (f_{pm}) created by R_m, L_m can be compensated by placing the EA1 amplifier zero at about four times the motor pole frequency.

$$f_{Pm} = \frac{R_m}{2\pi L_m},\ f_{ZC} = \frac{1}{2\pi R_5 C_3} \quad \text{Hz} \qquad 24$$

$$C_3 = \frac{L_m 10^9}{4R_5 R_m} \quad \text{nF} \qquad 25$$

Where, L_m in Henry, R_5 and R_m in Ohms
Place a high frequency pole at about one third of the switching frequency f_{ci} by adding C_p

$$C_P = \frac{3\times 10^9}{2\pi R_5 f_S} \quad \text{nF} \qquad 26$$

- **From system parameters: use 21 to calculate R_5, 25 to calculate C_3, and 26 to calculate C_p.**
- **Use the closest standard values for R_5, C_3, C_P**
- **Use the above values to re-calculate f_{Ci} in 27**

$$f_{C\,i} = \frac{25K_C R_S V_{dc} R_5 C_3}{2\pi V_S L_m R_1 (C_p + C_3)} \qquad 27$$

- **Verify that phase margin (PM) is more than 50 degrees, using equation 27**

$$PM = 180 - \left[\tan^{-1}\left\{\frac{f_{C\,i}}{f_{Pm}}\right\} + \tan^{-1}\left\{\frac{f_{C\,i}}{f_{PC1}}\right\} + \tan^{-1}\left\{\frac{f_{C\,i}}{f_{PC2}}\right\} - \tan^{-1}\left\{\frac{f_{C\,i}}{f_{ZC}}\right\} \right] \quad \text{degree} \qquad 28$$

Design Example

In the following design example RBE-03010-A motor will be used. The motor parameters are:

R_m = 0.974 Ω, L_m = 1.9 mH
Kb = 0.0431 V/rpm, Kt = 0.304 Ib.ft/A = 0.412 N.m/A, Tm= 8.49 Nm for linear Kt
P = Number of poles = 12
V_{dc} = Supply voltage = 110 V
I_P = Peak motor current =10 A
J = 4.52E-4 Nm.sec^2, = 2.03 ft.Ib. sec^2
B = Viscous damping = 6.86E-5 Nm/rpm=5.06 ft.Ib/rpm

Since I_P=10A, R_S= 0.010 Ω, then from equation 4, K_C=1.
Let R_1=10K, V_S=7.2V, V_{dc} = 110 V, K_C =1, R_S = 0.01.

- **Use 21 to calculate R_5, 25 to calculate C_3, and 26 to calculate C_p.**
- **Use the closest standard values for R_5, and C_3, then use equation 22 to recalculate f_{Ci}.**
- **R_P is set internally to 1300 KΩ, use the characteristic equation 29 to calculate f_{PC1} and f_{PC2}**

$$R_P R_5 C_P C_3 s^2 + (R_P C_3 + R_P C_P + R_5 C_3)s + 1 = 0 \qquad 29$$

Table 1

	$f_{C\,i}$	C_P pF	R_5 KΩ	C_3 nF	f_{ZC} Hz	f_{PC1} Hz	f_{PC2} Hz	PM Deg
Using Exact values	2720	296	89.5	5.45	326	21.3	6410	
Using Standard values	**2770**	**200**	**90.0**	**5.0**	**354**	**22.12**	**9786**	**69**

Note:

- At high line, where the supply is 110 Volts DC, $f_{C\,i}$, is 2.77 kHz. The crossover frequency drops to 1.76 kHz at low line, where the supply is approximately 70 Volts DC. If greater bandwidth is required, the current amplifier gain must be increased, requiring a corresponding increase in switching frequency to satisfy equations 21, 25, 26 and 27.

- **The loop phase margin (PM) is related to closed loop damping, roughly,**

  Damping Ratio = PM/100
- **Closed loop speed and rise time are roughly proportional to gain cross over frequency (loop bandwidth).**

Closed Loop Current Loop Transfer Function

When the current loop is closed, the output voltage of the current sense amplifier (25.R_S.K_C.I_m)2/3 is equal to the current command voltage (VC+)/3 at frequencies below the crossover frequency. The closed loop current loop transconductance is simply:

$$G_{C\ i} = \frac{\Delta I_m}{\Delta V_{C+}} = \frac{1}{50 R_S K_C} \quad \text{A / V} \qquad 30$$

With R_S=0.010 Ω, K_C=1, G_{Ci} is internally set to 2 A / V

At the open loop crossover frequency, the transconductance rolls off and assumes a single pole characteristic. The input divider network attenuates current sense amplifier output voltages signal to provide compatibility with typical servo controller, as described in equation (30). The overall amplifier transconductance is set internally to 2 amps/volt, allowing full scale current (+/-10 amps) with a +/-5 volt input command.
The closed loop current loop TF is

$$G_{C\,i}(jf) = \frac{I_m(jf)}{V_{C+}(jf)} = \frac{1}{50 R_S K_C (1 + j\ f / f_{C\,i})} \quad \text{A / V} \qquad 31$$

Fig 12 shows the current loop Bode Plots.

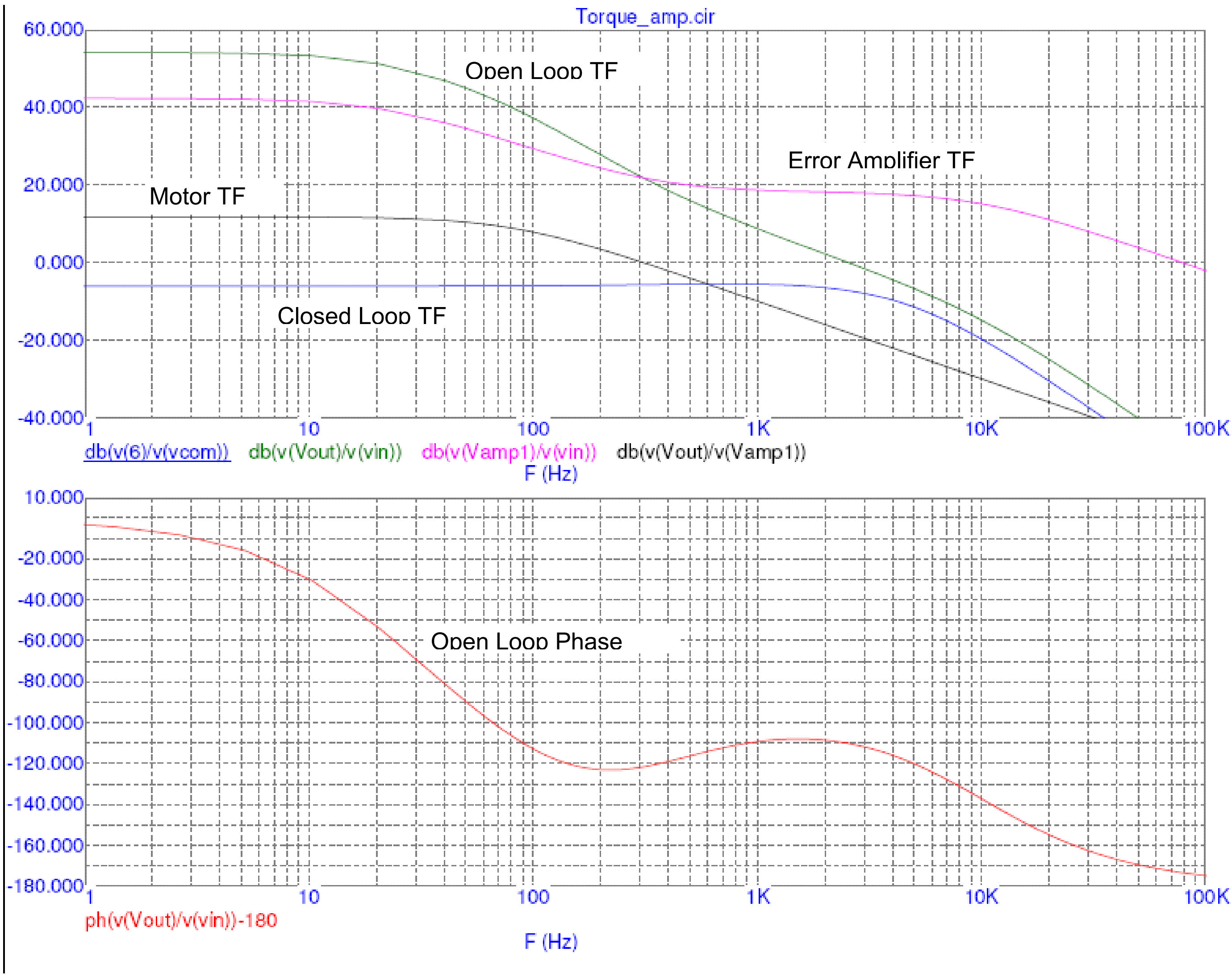


Fig.12 : Current Loop Bode Plots

Current Loop Design Summery

1- **First step,** from the system design requirements calculate the motor maximum peak current. The motor maximum peak current shall be about 20% higher than the maximum controlled current.

1-1 If the peak current is below 10A, then $K_C = 1$.
1-2 If the peak current I_P is between 10A to 20A, then calculate K_C from equation 4 assuming $R_S = 0.010\ \Omega$.
1-3 Use equation 1 to calculate Rg. Then connect Rg between Pins 24 and 25.

1-4 If the peak current is higher than 20A, then an additional current sense resistor shall be connected between Pins 27, 28 and Pins 29, 30. The recommended external sense resistor is 0.010 Ohm. The external resistor rating shall be 20A minimum. The total current sense resistor value will be $0.005\ \Omega$.
1-5 Calculate K_C from equation 4 assuming $R_S = 0.005\ \Omega$.
1-6 Use equation 1 to calculate Rg. Then connect Rg between Pins 24 and 25.

$$K_C = \frac{1}{10 R_S 1.2 I_P} \quad \text{A}$$

$$Rg = \frac{4.99 Kc}{1 - Kc}$$

2- **Second step,** use equation 21 to calculate R_5. f_S is internally set to 18KHz. If higher bandwidth is required, then f_S should be increased.

3- **Third step,** use equation 25 to calculate C_3.

4- **Fourth step,** use equation 26 to calculate C_P.

5- **Fifth step,** use the closest standard values for R_5, C_3, C_P in the following steps.

6- **Sixth step,** using equation 24, calculate EA1 zero f_{ZC}, and motor impedance pole.

7- **Seventh step,** calculate EA1 poles $\mathbf{f_{PC1}}$, $\mathbf{f_{PC2}}$ using equations 9 and 10. Then verify that $\mathbf{f_{PC2}}$ is more than $2\mathbf{f_{Ci}}$.

8- **Eighth step,** using equation 28, calculate current loop phase margin.

9- **Ninth step,** calculate the closed loop current loop TF, using equation 31.

$$G_{C\,i}(jf) = \frac{I_m(jf)}{V_{C+}(jf)} = \frac{1}{50 R_S K_C (1 + j\ f / f_{C\,i})} \quad \text{A / V}$$

10- **Tenth step,** Plot the current loop Bode Plots as shown in Fig. 12.

Current Loop Design Verification

Once the current loop compensation is completed, the system shall be tested at different operating conditions.

1- Initially, set DC bus voltage at less than 50% of nominal value. Start the motor by applying a command voltage at Pin 17. Monitor the phase output waveform and the output current waveform I_{SO} at Pin 20. The motor should run smoothly and the waveforms should be switching at 18KHz.
2- Change the command voltage from +ve to –ve. The motor will change direction of rotation. At zero command, the motor will stop while the output current will be a minimum at about zero average.
3- Use a square wave command voltage with frequency of 1Hz or less. Change to sinusoidal input command and monitor I_{SO} which should respond tracking the command input waveform.
4- Monitor the EA1 output at Pin 18, and the oscillator ramp voltage at Pin 21. Use two channels to monitor both waveforms. Verify that EA1 output slope is lower than the oscillator ramp slope. Lock the motor rotor and repeat this test with +/- commands.
5- If steps 1 to 4 are successful, increase the DC bus voltage to rated value and repeat tests.

Warning:

a- When running test 3, a voltage clamp should be used on the DC bus to protect against over-voltage due to motor regenerative action.
b- When monitoring Pins 18, 20, 21, use an RC, as shown in Fig. 9, to minimize noise injection by the oscilloscope probe.

Bipolar Command Test Example:

Fig. 13 shows a 100Hz sinusoidal current command and the corresponding motor current as measured at Pin 20.
Notice the motor current linear transition through zero current crossing, without zero-crossing distortion.

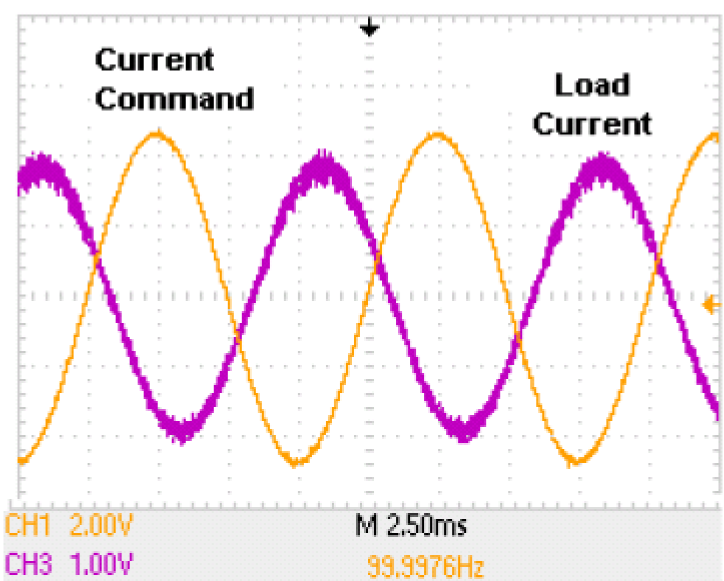


Fig.13 : Sinusoidal Current Command Response

Closed Loop Speed Control

Once the current loop is closed and tested, the speed loop can be closed, as shown in Figs. 15 and 16. An external Tachometer is needed to provide speed feedback.

A voltage command proportional to the desired speed is applied at speed command input. The speed feedback signal is obtained by an external Tachometer. The speed command shall be equal to the Tachometer output. The spare amplifier EA2 can be used to close the speed loop.

The Tachometer feedback signal shall be bipolar.

Small signal compensation of the speed control loop can be provided by an internal error amplifier, EA2.

Motor Mechanical Load Modeling

Industrial loads have different speed-torque characteristics. Fans, hoists, and compressors have the following speed-torque characteristics:

$$T_L = K\omega_m^2 \quad \text{for fans and pumps} \qquad 32$$

$$T_L = K_1 + K_2\omega_m \quad \text{for hoists} \qquad 33$$

$$T_L = K_1 + K_2\omega_m + K_3\omega_m^2 \quad \text{for compressors} \qquad 34$$

Where T_L is the torque in Nm, and ω_m is the angular speed in rad/sec.

Motor manufacturers usually specify a torque constant (K_T Nm/A) and a back emf constant (K_b V/rpm) for their motors. These constants have different values when the torque and speed are measured in English units, but they have the same numerical value when SI units are used. This becomes obvious when equating the motor internal input electro-magnetic power and the internal output mechanical power:

$$T_m\omega_m = E_b I_m \quad \text{watts} \qquad 35$$

$$\frac{T_m}{I_m} = K_T = \frac{E_b}{\omega_m} = K_b \qquad 36$$

where E_b is the motor internally generated armature voltage or bemf. The SI units for K_T and K_b are either Nm/A, or Vsec/rad.

The motor internal torque is related to motor current by the torque constant K_T. The motor total load is the sum of its internal load and the external load. The internal load is due motor inertia, friction, and viscous damping.

In the following analysis, assume the load is similar to a hoist load as described in equation 33.

$$T_m(s) = (J_m s + J_L s + B_V + B_L)\omega_m(s) + K_1 \qquad 37$$

where $\mathbf{J_m}$ is the motor inertia in **Nm.sec^2**, $\mathbf{B_V}$ is the motor viscous coefficient in **Nm.sec/rad**, $\mathbf{B_L}$ is the load constant in **Nm.sec/rad**, and $\mathbf{K_1}$ is the load fixed torque in **Nm**.

Substituting for T_m, and ω_m from 36 in 37

$$K_T I_m(s) = (J_m s + J_L s + B_V + B_L)\frac{E_b(s)}{K_T} + K_1 \qquad 38$$

$$I_m(s) = (J_m s + J_L s + B_V + B_L)\frac{E_b(s)}{K_T^2} + \frac{K_1}{K_T} \qquad 39$$

Equation 39 is analogous to the electric equation 40, for and RC circuit

$$I_m(s) = (C_m s + C_L s + \frac{1}{R_V} + \frac{1}{R_L})E_b(s) + I_1 \qquad 40$$

$$I_m(s) = \frac{1}{R}(RC\ s + 1)E_b(s) + I_1 \qquad 41$$

where

$$R = \frac{R_V R_L}{R_V + R_L}\ ,\ C = C_m + C_L \qquad 42$$

Analogies between 39 and 40 results in an RC equivalent circuit of the motor mechanical load, as shown in Fig. 13. Motor radian speed in rad/sec is analogous to voltage in V, motor torque in Nm is analogous to current in A, inertia in Nm.sec^2 is analogous to capacitance in F, and Viscous damping in Nm.sec/rad is analogous to resistance in Ω.
where:

$$C_m = \frac{J_m + J_L}{K_T^2} \quad \mathrm{F} \qquad 43$$

$$R_V = \frac{K_T^2}{B_V}\ ,\ \ R_L = \frac{K_T^2}{B_L} \quad \Omega \qquad 44$$

$$I_1 = \frac{K_1}{K_T} \quad \mathrm{A} \qquad 45$$

Now the motor and the load mechanical parameters are represented by an electrical circuit as shown in Fig.14 and equations 39 to 45.
Compensating 41 in 31, results in a small signal TF for the power section of the speed loop.

$$G_{Cm}(jf) = \frac{E_b(jf)}{V_{C+}(jf)} = \frac{R}{50R_S K_C(j\ f / f_{C\ i} + 1)(2\pi RC\ jf + 1)} \quad \mathrm{V/V} \qquad 46 \quad \text{use s}$$

The mechanical load pole is f_{LP}

$$f_{Lp} = \frac{1}{2\pi RC} \quad \mathrm{Hz} \qquad 47$$

The Bode plot of $G_{Cm}(jf)$ is shown in Fig. 16

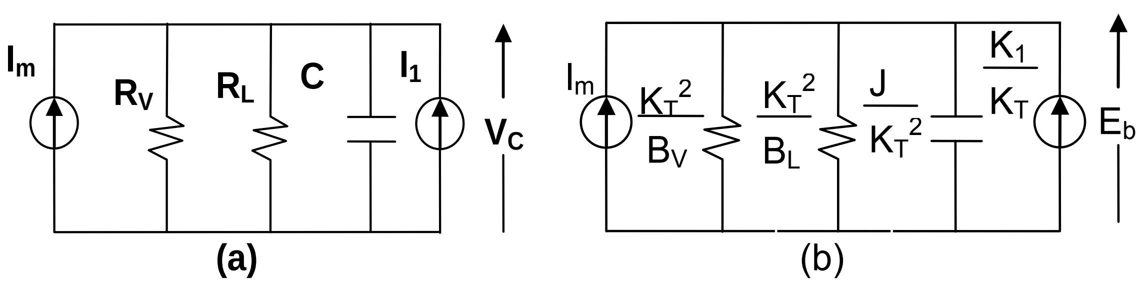


Fig. 14. Electric RC Circuit & Motor Analogous Mechanical Equivalent Circuit

EA Transfer Function

Referring to Fig.15, Let

C_{PV} be the equivalent capacitance of C_5 and C_6 in parallel.

The transfer function of the EA2 is

$$G_{EA2}(s) = \frac{V_{C+}(s)}{E_{tm}(s)} = \frac{R_9(R_7C_4s+1)/R_6}{\left[R_9R_7C_{PV}C_4s^2 + (R_9C_4 + R_9C_{PV} + R_7C_4)s + 1\right]} \quad 48$$

or,

$$G_{EA2}(s) = \frac{K_2(s/\omega_{ZV}+1)}{(s/\omega_{PV1}+1)(s/\omega_{PV2}+1)} \quad 49$$

Where

$$K_2 = \frac{R_9}{R_6} \quad 50$$

$$\omega_{ZV} = \frac{1}{R_7C_4} \quad 51$$

ω_{PV1} and ω_{PV2} can be calculated by equating the denominator of equation 5 to zero, or simplified by

$$\omega_{PV1} \approx \frac{1}{R_9(C_{PV}+C_4)} \quad 52$$

$$\omega_{PV2} \approx \frac{(R_9+R_7)}{R_9R_7C_{PV}} \quad 53$$

The voltage loop error amplifier EA2 transfer function Bode plot is shown in Fig. 17.
If $R_9 >> R_7$, the EA2 gain at ω_{ZV} is

$$G_{EA2}(s=\omega_{ZV}) \approx \frac{R_7C_4}{R_6(C_{PV}+C_4)} \quad 54$$

More accurately, assuming $\omega_{P2} >> \omega_{ZV}$, ω_{P1}

$$G_{EA2}(s=\omega_{ZV}) \approx \frac{R_9}{R_6}\frac{\omega_{P1}}{\omega_{ZV}} = \frac{R_9}{R_6}\frac{f_{PV1}}{f_{ZV}} \quad 55$$

Closing the Speed (Voltage) Loop

The internal error amplifier EA2 can be used to close the speed loop as shown in Figs. 15, and 16. An external tachometer provides a voltage proportional to speed. The tachometer output will be used to close the speed loop.

Tachometer TF

Tachometer output dc voltage E_{tm} is related to motor speed n_m rpm by a constant gain K_{tm}

$$\frac{E_{tm}(jf)}{n_m(jf)} = K_{tm} \qquad \text{V / rpm} \qquad 56$$

But

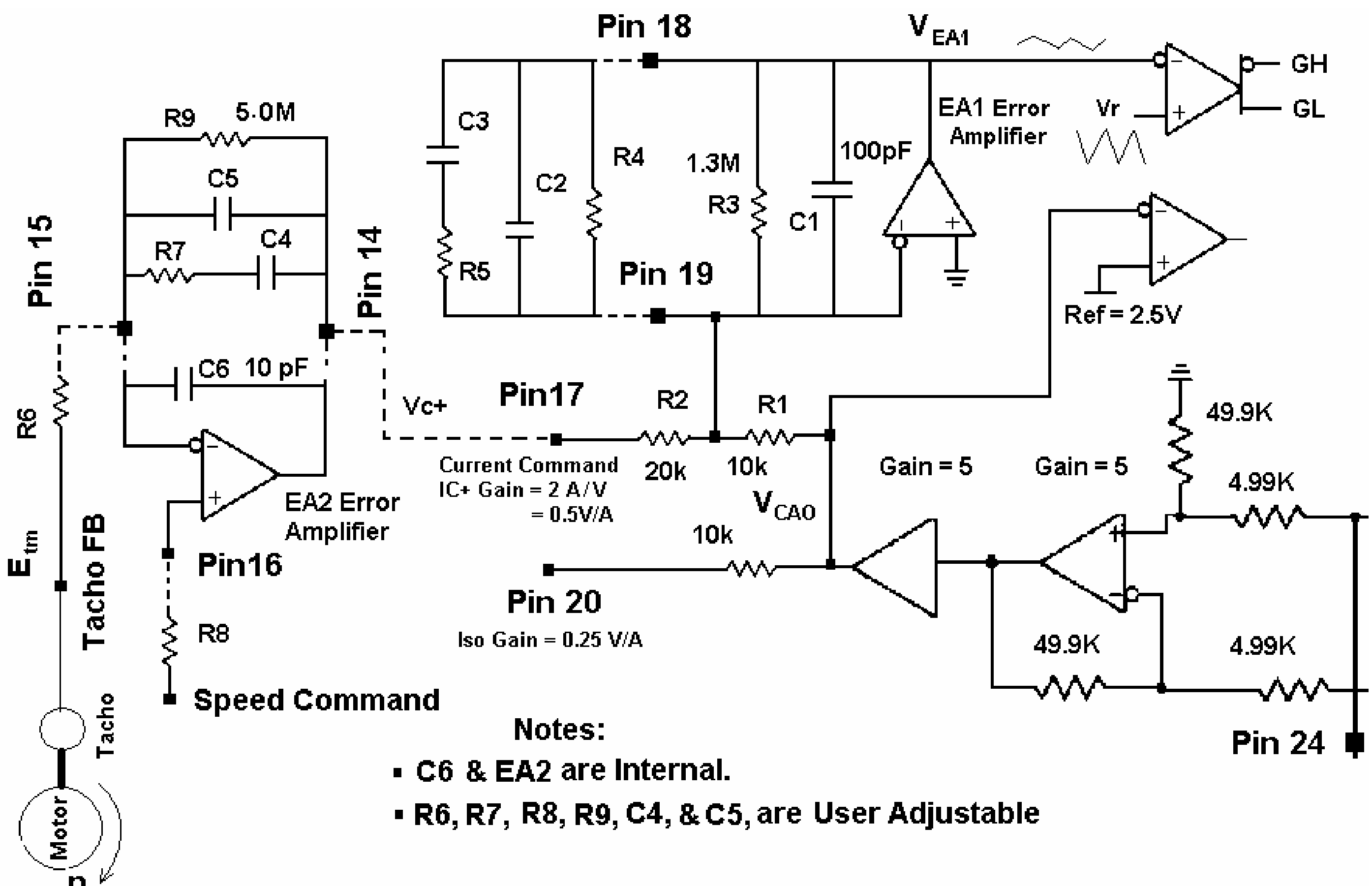
$$\frac{E_b(jf)}{n_m(jf)} = \frac{\pi K_T}{30} \qquad \text{V / rpm} \qquad 57$$

Substituting 57 in 56

$$G_{tm}(jf) = \frac{E_{tm}(jf)}{E_b(jf)} = \frac{30K_{tm}}{\pi K_T} \qquad \text{V / V} \qquad 58$$

From 58 and 46, the small signal TF of the motor mechanical power circuit including the tachometer is

$$G_{Cm}(jf)G_{tm}(jf) = \frac{E_{tm}(jf)}{V_{C+}(jf)} = \frac{30K_{tm}R}{50\pi R_S K_T K_C (j\ f / f_{LP} + 1)(j\ f / f_{C\ i} + 1)} \qquad \text{V / V} \qquad 59$$



Notes:
- C6 & EA2 are Internal.
- R6, R7, R8, R9, C4, & C5, are User Adjustable

Fig. 15 . Closing Voltage Loop (Speed Loop) Using EA2

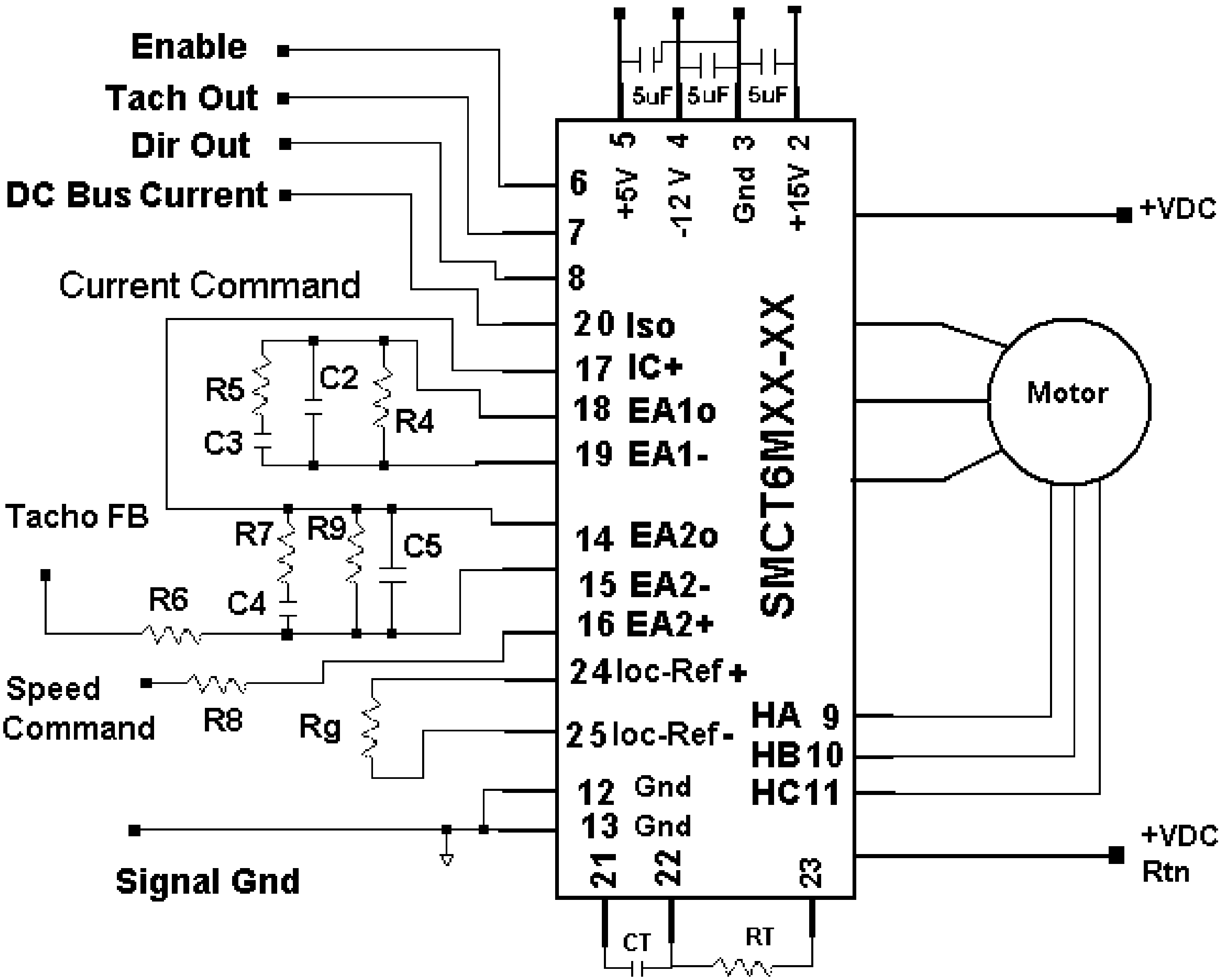


Fig. 16: Closed Loop Speed & Current Control

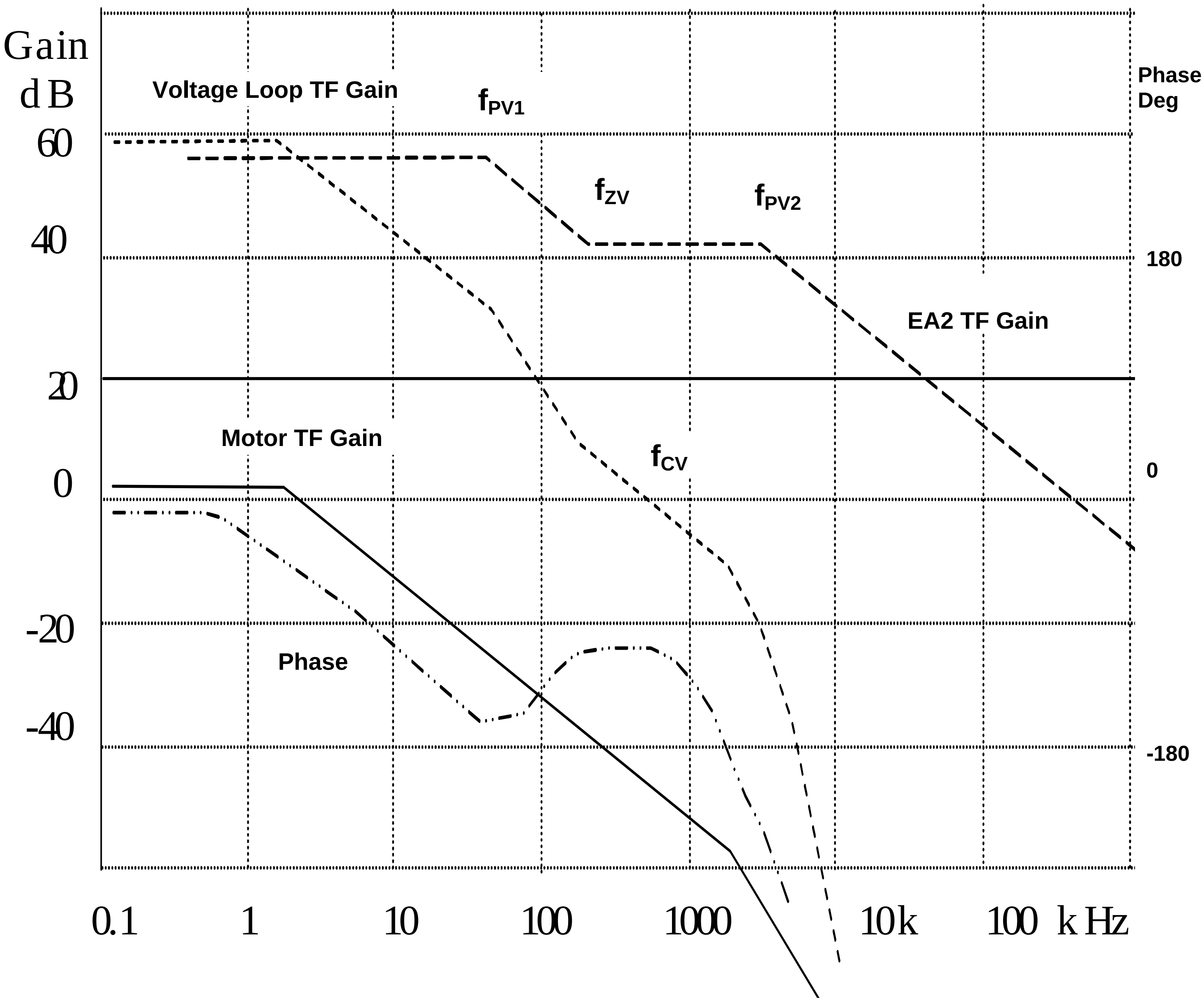


Fig. 17: Voltage Loop Open Loop Gain & Phase versus Frequency

Voltage Loop Compensation

Speed loop error amplifier EA2 has a low frequency pole, and the motor mechanical TF G_{Cm} also has a low frequency pole. It is desired to have the speed loop cross over frequency, f_{CV}, to be in the flat portion of EA2 Bode plot, so that the overall speed loop has -20db slope at f_{CV}.

Design Rules and Steps:

- **Speed loop cross over frequency must be below the current loop cross over frequency.**
- **Typically f_{CV} is less than $0.5f_{Ci}$.**

$$f_{CV} = 0.3 f_{C\,i} \qquad 60$$

- The speed loop error amplifier gain is described in equation 48, and the motor mechanical power circuit gain is described in equation 59. The overall open loop gain of the speed loop is the product of the speed loop error amplifier gain and motor mechanical power circuit gain. The EA2 gain in the flat portion of Fig 17 is described by equations 54 and 55. The loop gain crossover frequency, f_{CV}, should be within the flat portion of the EA2 gain in order to achieve enough phase margin.

The overall open loop gain of the speed loop is set equal to one to solve for the loop gain crossover frequency, f_{CV}, as shown in equation 61.

$$\frac{R_7 C_4}{R_6 (C_{PV} + C_4)} \frac{30 K_{tm}}{50\pi R_S K_T K_C 2\pi C \ f_{CV}} = 1 \qquad 61$$

But $C_4 >> C_{PV}$, substituting 60 in 61

$$\frac{R_7}{R_6} = \frac{\pi^2 R_S K_T K_C C \ f_{C\,i}}{K_{tm}} \qquad 62$$

Assume R_6 = 10K, then Calculate R_7 and use the closest standard value in the following steps.

- **Place EA2 zero f_{ZV} at $0.1f_{CV}$**

$$f_{ZV} = 0.03 f_{C\,i} = \frac{1}{2\pi R_7 C_4} \quad \text{Hz} \qquad 63$$

$$C_4 = \frac{10^{12}}{0.06\pi R_7 f_{C\,i}} \quad \text{pF} \qquad 64$$

- **Place EA2 pole f_{PV2} at above $2f_{CV}$, at least one decade above f_{ZV} to insure adequate loop phase margin of at least 40 degrees.**

$$f_{PV2} \approx \frac{(R_9 + R_7)}{2\pi R_9 R_7 C_{PV}} = 20\, f_{ZV} = 2\, f_{CV} = 0.6\, f_{C\,i} \qquad 65$$

Or,

$$C_{PV} = \frac{10^{12}(R_9 + R_7)}{1.2\pi R_9 R_7\, f_{C\,i}} = C_5 + C_6 \qquad \text{pF} \qquad 66$$

But C_6 =10pF

$$C_5 = C_{PV} - 10 \qquad \text{pF} \qquad 67$$

- **Use standard values for R_6, R_7, C_4, and C_5. Use R_9= 5.0M. Also, keep in mind the internal values for C_6=10pF. Once all the compensation RC component values are determined for EA2. Calculate the exact values for f_{ZV} using 63, and solve the characteristic equation 68 to calculate f_{PV1} and f_{PV2}, then calculate f_{CV} using 69,**

$$R_9 R_7 C_{PV} C_4 s^2 + (R_9 C_4 + R_9 C_{PV} + R_7 C_4)s + 1 = 0 \qquad 68$$

$$f_{CV} = \frac{30 K_{tm} R}{50\pi R_S K_T K_C} \; \frac{R_9}{R_6} \frac{f_{PV1}}{f_{ZV}}\, f_{LP} \qquad 69$$

- **Verify that phase margin (PM) is more than 40 degrees, using equation 70**

$$PM = 180 - \left[\tan^{-1}\left\{\frac{f_{CV}}{f_{LP}}\right\} + \tan^{-1}\left\{\frac{f_{CV}}{f_{C\,i}}\right\} + \tan^{-1}\left\{\frac{f_{CV}}{f_{PV1}}\right\} + \tan^{-1}\left\{\frac{f_{CV}}{f_{PV2}}\right\} - \tan^{-1}\left\{\frac{f_{CV}}{f_{ZV}}\right\} \right] \quad \text{degree} \qquad 70$$

Speed Loop Design Summery

1- **First step,** Assume R_6 = 10k, R_9 =5.0 M.
2- **Second step,** use equation 62 to calculate R_7 /R_6 .
3- **Third step,** use equation 64 to calculate C_4 .
4- **Fourth step,** use equation 66 to calculate C_{PV} .
5- **Fifth step,** use equation 68 to calculate f_{PV1}, f_{PV2}.
6- **Sixth step,** use equation 63 to calculate f_{ZV}.
7- **Seventh step,** use equation 69 to calculate f_{CV}.
8- **Eighth step,** use equation 70 to calculate speed loop phase margin.
9- **Ninth step,** Plot Bode plots.

Design Example

In the following design example RBE-03010-A motor will be used. The motor parameters are:

R_m = 0.974 Ω, L_m = 1.9 mH
K_b = 0.0431 V/rpm, **K_T = 0.412 N.m/A**= 0.304 lb.ft/A, max T_m = 8.49 Nm for linear K_T
P = Number of poles = 12
V_{dc} = Supply voltage = 110 V
I_P = Peak motor current =10 A
J_m = 4.52E-4 Nm.sec^2, = 2.03 ft.lb. sec^2
B_V = Viscous damping = 6.86E-5 Nm/rpm =5.06E-5 ft.lb/rpm
= 60(6.86E-5)/2π = **6.5508E-4 Nm.sec/rad**

- **Let the motor load be 628 watts (3 Nm) at rated speed of 2000 rpm (209.44 rad/sec), and load torque be expressed as 0.01432 ω_m. Also, let J_L= 3J_m.**

Use equations 35 to 44 to calculate motor and load parameters

$$C_m = \frac{4.52E-4}{0.412^2} = 0.00266 \quad \text{F}$$

$$C = C_m + C_L = 0.01064 \quad \text{F}$$

$$R_V = \frac{0.412^2}{6.5508E-4} = 259.1 \quad \Omega$$

$$R_L = \frac{0.412^2}{0.01432} = 11.854 \quad \Omega$$

$$R = \frac{R_V R_L}{R_V + R_L} = 11.335 \quad \Omega$$

$$E_b = 0.412 \times 209.44 = 86.29 \quad \text{V}$$

- **Notice that motor total internal mechanical power is E_b^2/R = 657 watts, and Im= 7.61 A.**

Table 2
Corresponding Speed Loop Compensation Parameters to Table 1

	f_{Ci} Hz	f_{CV} Hz	C_{PV} pF	R_7 MΩ	C_4 pF	f_{ZV} Hz	f_{PV1} Hz	f_{PV2} Hz	PM Deg
Using Exact values	**2770**	452	42.0	3.99	480	83	35	1708	
Using Standard values		**453**	**40**	**4.0**	**450**	**88**	**38**	**1825**	**61**

- **Plot the speed loop Bode Plots as shown in Fig. 18.**

Notes:

- R_6 =10K, R_9 = 5.1M.
- f_{LP}=1.32 Hz
- C_{PV} =C_5 + 10 pF
- At high line, where the supply is 110 Volts DC, $f_{C\,i}$, is 2.77 kHz. The current loop crossover frequency drops to 1.76 kHz at low line, where the supply is approximately 70 Volts DC.
- Notice that the speed loop crossover frequency is independent of DC bus voltage. **At low line voltage $f_{C\,i}$ moves to the left closer to f_{CV}, reducing speed loop phase margin to 55 degrees. Further drop in supply voltage may cause instability in the speed loop. The system should be tested for this condition.**
- **So far, the tachometer feedback gain is assumed to be constant. If the tachometer has a transfer function, it should be added to equation 59.**

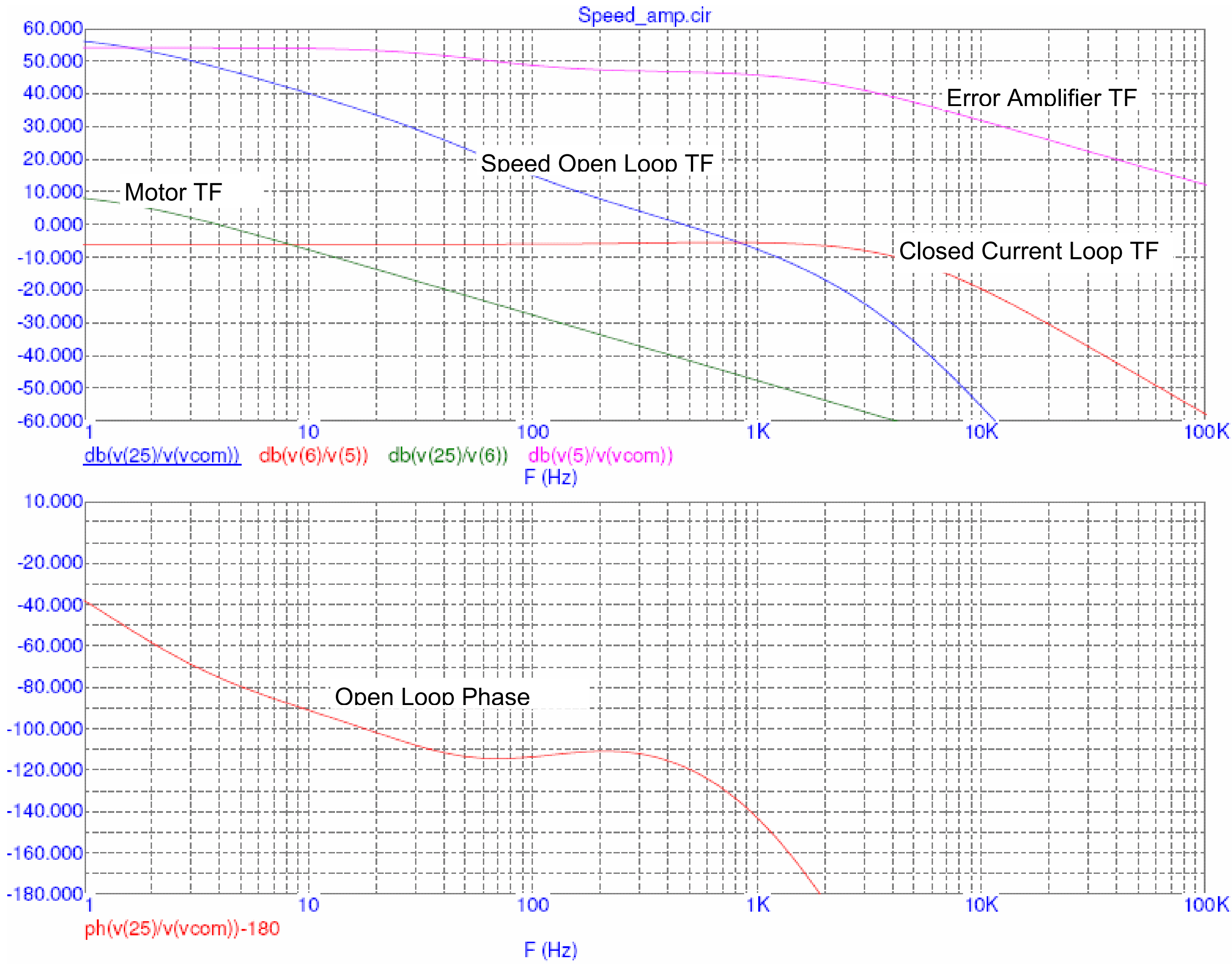


Fig.18 : Speed Loop Bode Plots

DC Bus Filtering

To minimize the circuit parasitic inductance effect on the power stage, the layout of Fig. 19 is suggested. C1, C2, C3, C4, C5 and C6 are 0.1μF to 0.5μF ceramic capacitors, connected across each leg of the three-phase bridge. Also, a bulk polarized capacitor C7 or a film capacitor, with adequately voltage-rated and low ESR, should be connected across the DC bus. The capacitor value depends on the ESR of the capacitor and the allowable DC bus voltage repel.

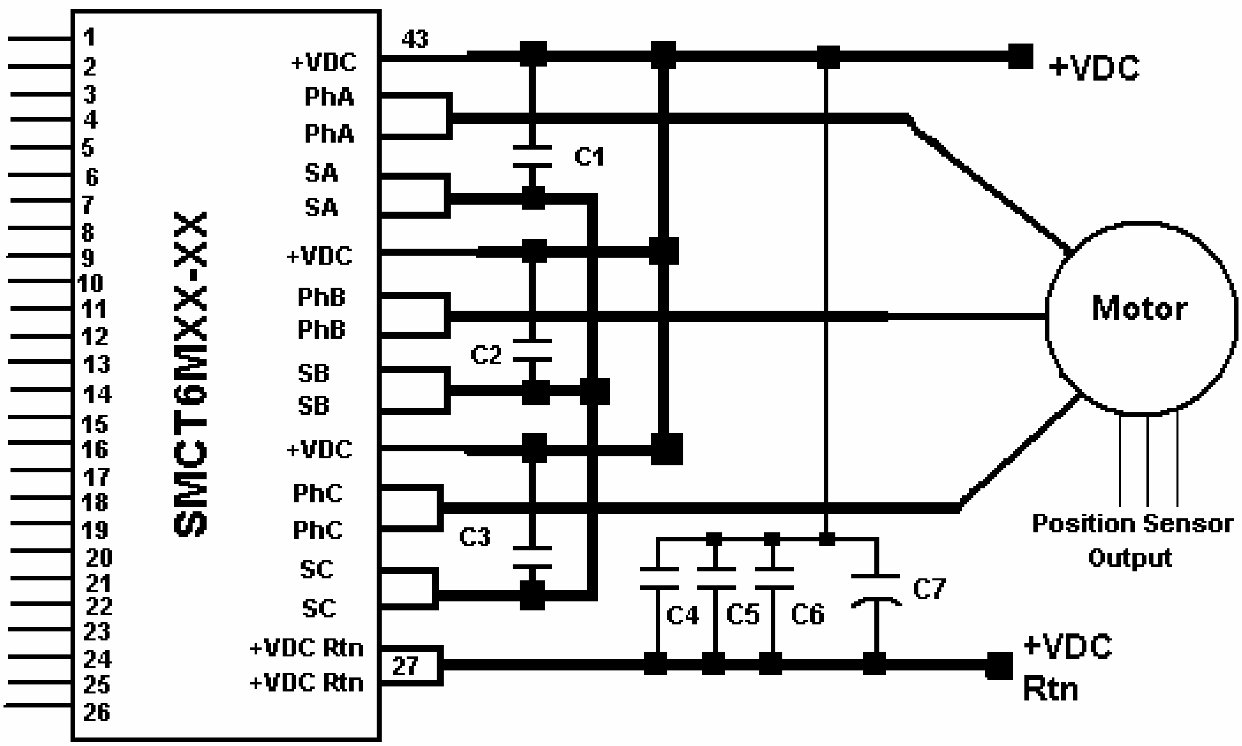


Fig. 19: DC Bus Bypass Capacitors

Complementary PWM Switching

The current command and the current sense amplifier can be configured in a closed loop mode as shown in Fig. 6. The output of the error amplifier EA1o is internally connected to one input of the PWM comparator, while the other input is connected to the oscillator ramp.

Fig. 20 shows the oscillator ramp, corresponding gate drive signals, and motor winding current for both standard 4-quadrant switching and complementary 4-quadrant switching for Fig. 23.

In the standard 4-quadrant switching S1 and S4 are modulated while S2 and S4 are off. When S1 and S4 are on, the motor current ramps up, and when S1 and S4 are off, the motor current freewheels in the body diodes of S2 and S3 down to zero. The motor average current is always positive. There is no way to make the motor average current zero. This method of control fails to provide linear transition through zero current crossing.

In the complementary 4-quadrant switching S1 and S4 are modulated and S2 and S4 are modulated in a complementary fashion to S1 and S4. When S1 and S4 are on, the motor current ramps up. When S1 and S4 are off, S2 and S3 are on and the motor current reverses polarity in each PWM cycle resulting in a zero average motor current. This method of control provides linear transition through zero current crossing as shown in Fig. 13.

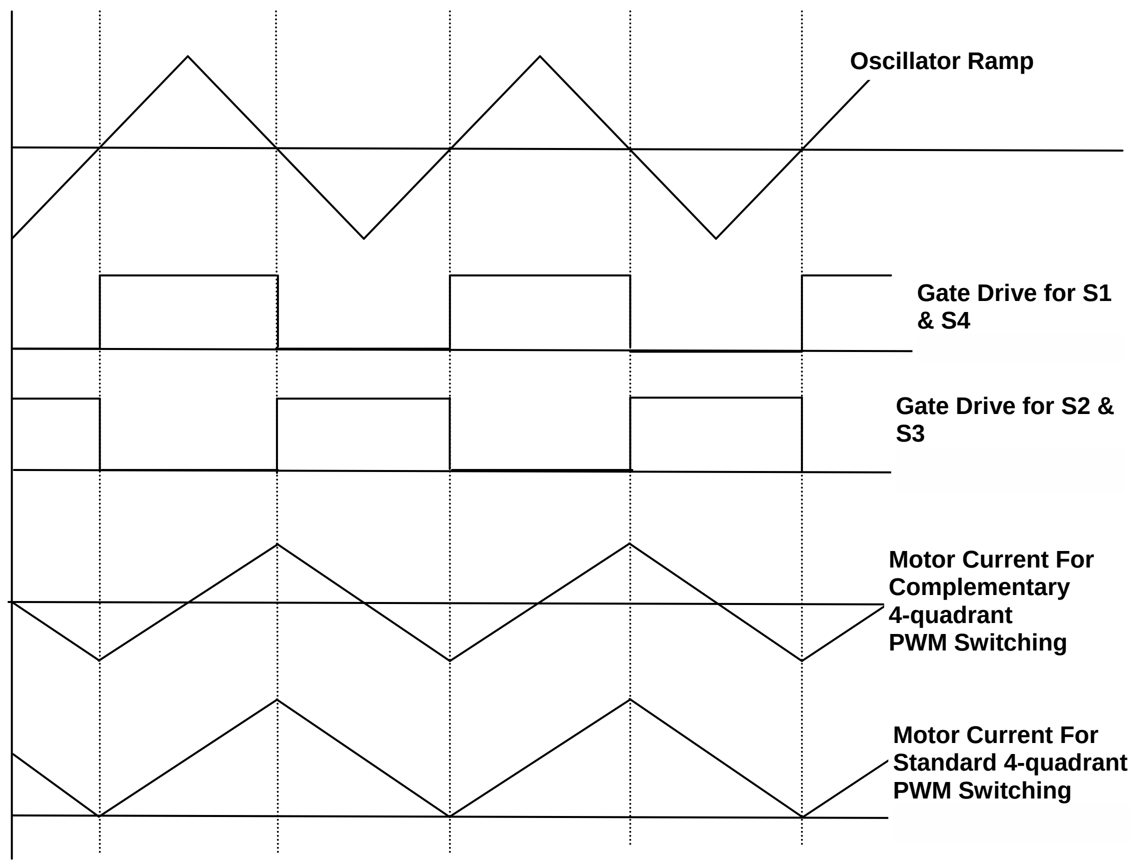


Fig. 20: Complementary PWM Switching of SMCT6MXX-XX, vs. Standard PWM Switching

Standard PWM Switching

Fig. 21 illustrates the four possible quadrants of operation for a BDC motor. Two-quadrant mode refers to a motor operating in quadrants I and III. With a two-quadrant BDC motor, friction is the only force to decelerate the load.

Four-quadrant control provides controlled operation in all quadrants, including II and IV, where torque and rotation are of opposite directions.

Two-quadrant mode modulates only the high-side devices of the output power stage. The current paths within the output stage during the PWM on and off times are illustrated in Fig. 22. During the on time, both switches S1 and S4 are on, the current flows through both switches and the motor winding. During the PWM cycle off time, the upper switch S1 is shut off, and the motor current circulates through the lower switch S4 and D2. The motor is assumed to be operated in quadrants I or III.

In four-quadrant mode, both upper and lower switches are modulated. Motor current always decays during off time, eliminating any uncontrolled circulating current. In addition, the current always flows through the current sense resistor. Fig. 23 illustrates the current paths during a PWM cycle.

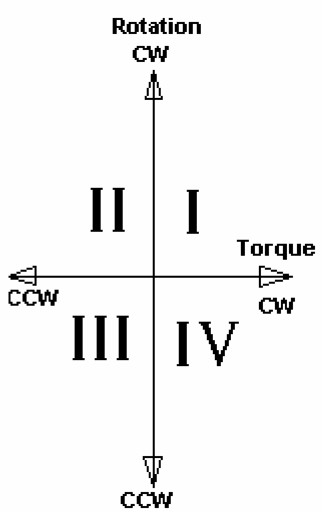


Fig.21. Four Quadrants of Operation

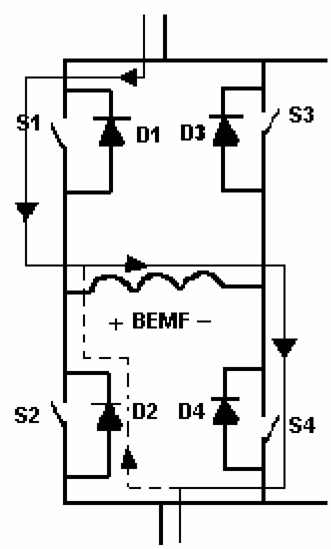


Fig.22 . Two-Quadrant Forward

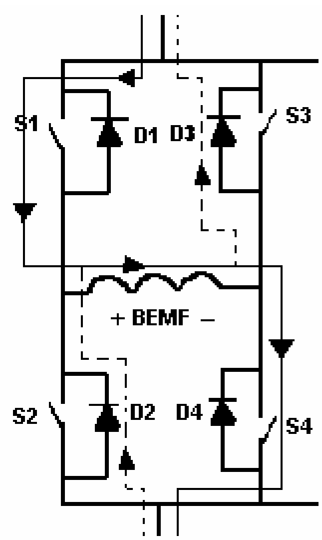


Fig. 23 . Four-Quadrant Forward

Operation and Advantages

SMCT6MXX-XX uses complementary four-quadrant PWM switching technique. This complementary PWM switching produces a 50% duty cycle in response to a zero current command at Pin 17. The average motor current at 50% duty cycle is near zero, motor magnetizing current only. The motor winding current and direction of rotation are controlled by the same command. A positive polarity command will produce a duty cycle higher than 50%, resulting in a clockwise rotation. A negative polarity command will produce a duty cycle less than 50%, resulting in a counter clockwise rotation. The motor current magnitude is a follower of the current command value. The motor speed depends on the current command and the load.

The average motor current and direction of rotation can be controlled though zero crossing in a linear fashion.

Fig. 6 illustrates a closed loop current control. The use of average current control simplifies the control loop by eliminating the pole created by the motor winding inductance.

Major Advantages of Complementary PWM Switching

The major advantages of complementary 4-quadrant mode of operation are:

1- Provides holding torque at zero current command.
2- Motor direction of rotation is defined by the command polarity.
3- Linear motor phase current control from negative to positive through zero crossing.
4- No dead band around zero.
5- Can produce high current loop bandwidth.
6- Average current control eliminates the pole created by the motor winding inductance.

120° Rotor Position Sensing

HA, HB, and HC are designed to accept rotor position information from hall sensors positioned 120° apart. Motors with 60° position sensing may be used if one or two of the hall-effect sensor signals is inverted prior to connection to the hall-effect inputs. **HA, HB, and HC** inputs are internally pulled up, Zener clamped to 5.0V, and filtered.

60° Rotor Position Sensing

SMCT6MXX-XX is designed to operate with 120° position sensing encoding. In this format, the three position sensor signals are never simultaneously high or low. Internal protection logic disables all three Phase Outputs when the Hall-Effect Inputs are set to an illegal condition (i.e., all logic low or all logic high).

The output of HA is in phase with motor back EMF voltage VAC, HB is in phase VCB, and HC is in phase with VBA as shown in Fig. 24.

Motors whose sensors provide 60° encoding can be converted to 120° using the circuit shown in Fig. 25.

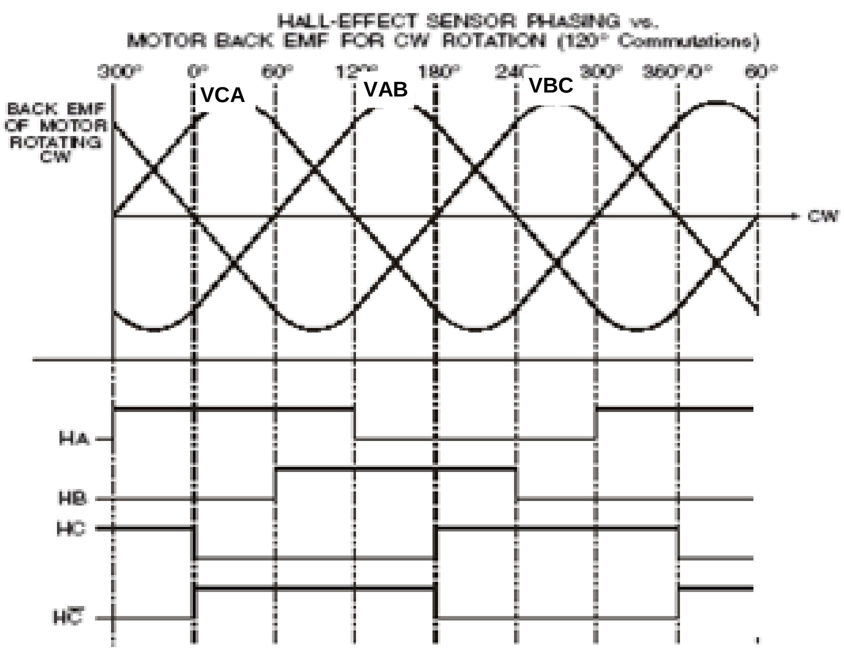


Fig. 24: Hall Signal Phasing

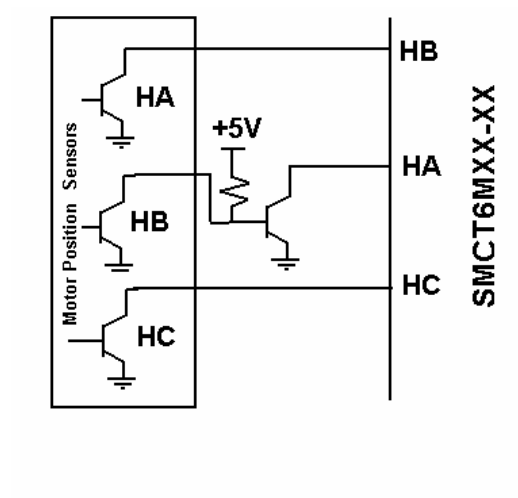


Fig. 25. Converting Hall Sensors' Position from 60° to 120°

ORDERING INFORMATION

SELECTOR GUIDE TABLE

Part Number	Operating DC Bus Supply Voltage	Peak DC Bus Voltage	RMS Output Current	Peak Output Current	Rds(on) I_D =40A	Hermetic?
SMCT6M40-10-YY	60	100	40	60	5	Yes
SMCT6M40-10-1-YY	60	100	40	60	5	No
SMCT6M40-25-YY	150	250	30	40	35	Yes
SMCT6M40-25-1-YY	150	250	30	40	35	No

PART NUMBERING SYSTEM

SMCT6MXX-XX-1-YYZ where YY is the sense resistor value and Z is the lead bend option as required:

For instance, part number SMCT6MXX-XX-1-30B has a 30mOhm resistor, and option B lead bend.

Typical current limit sense resistor values are 10, 30, mOhms. Contact the factory for other options.

Cleaning Process:
Suggested precaution following cleaning procedure:
If the non-hermetic parts are to be cleaned in an aqueous based cleaning solution, it is recommended that the parts be baked immediately after cleaning. This is to remove any moisture that may have permeated into the device during the cleaning process. For aqueous based solutions, the recommended process is to bake for at least 2 hours at 125^{o}C.
Do not use solvents based cleaners.

Soldering Procedure:
Recommended soldering procedure
Signal pins 1-26: 210^{o}C for 10 seconds max
Power pins 27-43: 260^{o}C for 10 seconds max. Pre-warm module to 125 ^{o}C to aid in power pins soldering

DISCLAIMER:

1- The information given herein, including the specifications and dimensions, is subject to change without prior notice to improve product characteristics. Before ordering, purchasers are advised to contact the Sensitron Semiconductor sales department for the latest version of the datasheet(s).
2- In cases where extremely high reliability is required (such as use in nuclear power control, aerospace and aviation, traffic equipment, medical equipment , and safety equipment) , safety should be ensured by using semiconductor devices that feature assured safety or by means of users' fail-safe precautions or other arrangement .
3- In no event shall Sensitron Semiconductor be liable for any damages that may result from an accident or any other cause during operation of the user's units according to the datasheet(s). Sensitron Semiconductor assumes no responsibility for any intellectual property claims or any other problems that may result from applications of information, products or circuits described in the datasheets.
4- In no event shall Sensitron Semiconductor be liable for any failure in a semiconductor device or any secondary damage resulting from use at a value exceeding the absolute maximum rating.
5- No license is granted by the datasheet(s) under any patents or other rights of any third party or Sensitron Semiconductor.
6- The datasheet(s) may not be reproduced or duplicated, in any form, in whole or part, without the expressed written permission of Sensitron Semiconductor.
7- The products (technologies) described in the datasheet(s) are not to be provided to any party whose purpose in their application will hinder maintenance of international peace and safety nor are they to be applied to that purpose by their direct purchasers or any third party. When exporting these products (technologies), the necessary procedures are to be taken in accordance with related laws and regulations.