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TECHNICAL DATA  
DATA SHEET 5044, PRELIMINARY RELEASE

## **TORQUE CONTROLLER BRUSHLESS DC MOTOR DRIVER MODULE 600V/70A, 1200V/60A**

### **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
- 70A peak Phase Current with 300V 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
- Desaturation protection for all six IGBTs.
- Case temperature sense output.
- Over-temperature shutdown.
- Package size - 3.59" x 1.55" x 0.80"
- Total Weight 5.0 OZ.

### **APPLICATIONS:**

- Servo positioning systems
- Actuation systems
- Hoists

### **DESCRIPTION:**

SMCT6GXXX-XXX 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. SMCT6GXXX-XXX is best used as a four-quadrant torque controller for controlling servo systems.

The SMCT6GXXX-XXX 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 SMCT6GXXX-XXX 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.

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**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° 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<sub>C</sub>=25 °C) unless otherwise noted

Characteristic	Maximum
Motor DC Bus Supply Voltage SMCT6G070-060-1 SMCT6G060-120-1	350V 700V
Motor Peak Voltage SMCT6G070-060-1 SMCT6G060-120-1	600V 1200V
RMS Output Current SMCT6G070-060-1 SMCT6G060-120-1	60 A 40A
Peak Output Current SMCT6G070-060-1 SMCT6G060-120-1	70 A 60A
+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
Operating & Storage Junction Temperature	-55°C to +150°C
IGBT Thermal Resistance R <sub>thjC</sub> Diode Thermal Resistance R <sub>thjC</sub>	0.45 °C/W 0.80 °C/W

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**ABSOLUTE MAXIMUM RATINGS (continued)**

(T<sub>C</sub>=25 °C) unless otherwise noted

Characteristic	Maximum
Pin-to-Case Voltage Isolation, at room conditions	1500V DC
Lead Soldering Temperature, 10 seconds maximum, 0.125" from case	300°C

**RECOMMENDED OPERATING CONDITIONS**

(T<sub>C</sub>=25 °C) unless otherwise noted

Characteristic	Maximum
Motor DC Bus Supply Voltage SMCT6G070-060-1 SMCT6G060-120-1	300V 600V
RMS Output Current, T <sub>C</sub> =80 °C SMCT6G070-060-1 SMCT6G060-120-1	50 A 30A
Peak Output Current, T <sub>C</sub> =80 °C SMCT6G070-060-1 SMCT6G060-120-1	60 A 40A
+15V Supply Voltage	+ 14 V +/-5%

PARAMETER SYMBOL CONDITIONS <sup>(1)</sup>	MIN.	TYP.	MAX.	UNITS
<b>Power Output Section</b>				
<b>Collector-Emitter Leakage Current</b> ICES SMCT6G070-060-1 at VCE=480V SMCT6G060-120-1 at VCE=960V			400	uA
<b>Collector-to-Emitter Saturation Voltage</b> VCEsat, at VCC=15V SMCT6G070-060-1 IC=30A IC=70A IC=90A SMCT6G060-120-1 IC=40A		1.15 1.50 1.70 1.9	1.30 1.80 2.00 2.3	V
<b>Diode Forward Voltage</b> SMCT6G070-060-1 IC=30A IC=70A IC=90A SMCT6G060-120-1 IC=40A		1.18 1.50 1.62 1.8	1.35 1.80 1.95 2.3	V

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**RECOMMENDED OPERATING CONDITIONS (continued)**

(T<sub>C</sub>=25 °C) unless otherwise noted

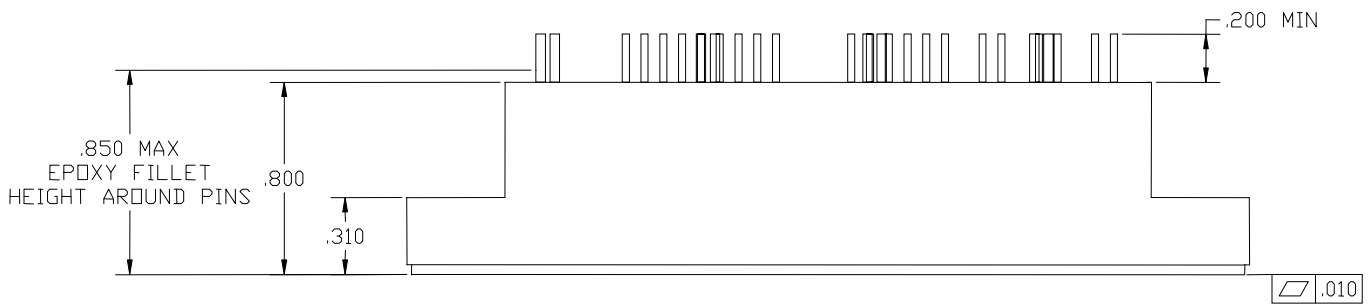
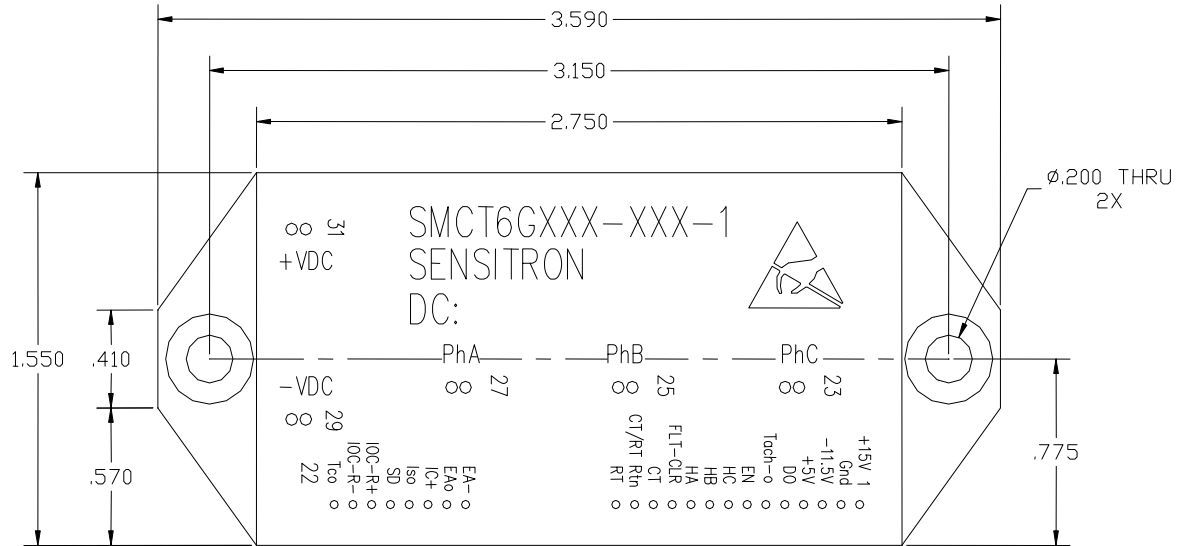
PARAMETER SYMBOL CONDITIONS <sup>(1)</sup>	MIN.	TYP.	MAX.	UNITS
<b>Control Section</b>				
Input Supply Current I <sub>cc</sub> at +14V supply without any external load on -11.5V or +5V			80	mA
-11.5V Output Range	-11.0	11.5	-12.0	V
+5V Output Range	4.7		5.3	V
<b>Error Amplifier EA</b>				
Output Offset Voltage	-20	-	20	mV
Amplifier Output Voltage Range	-11	-	11.5	V
<b>Current-Sense Amplifier Section</b>				
Amplifier Voltage Gain	0.058	0.06	0.062	V/A
Out DC Offset	-10	0	+10	mV
Current Limit Set Point Voltage	+/-2.3	+/-2.5	+/-2.7	V
Corresponding Current Limit with R <sub>s</sub> =0.002 Ohm	-	+/-42.0	-	A
<b>Logic Input Section</b>				
HA, HB, HC, and EN High Level Input Voltage Threshold	2.0	-	-	V
Low Level Input Voltage Threshold	-	-	1.0	V
<b>Tachometer</b>				
Tachometer Output High Level V <sub>oh</sub>	4.5	4.8	5.2	V
Tachometer Output Low Level V <sub>ol</sub>	-	-	200	mV
Tachometer Frequency Ratio To Hall Input Frequency	3	3	3	-
<b>Oscillator Section</b>				
Oscillator Frequency f <sub>s</sub>	13	15.0	18.0	kHz

**NOTES:**

- All parameters specified for T<sub>a</sub> = 25°C, V<sub>cc</sub> = +14Vdc, and all Phase Outputs unloaded. All negative currents shown are sourced by (flow from) the Pin under test.
- Pulse Test: Pulse Width < 300 μSec, Duty Cycle < 2%.

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**Package Drawing Top View**  
**(All dimensions are in inches, tolerance is +/- 0.010")**



**Fig. 2: Mechanical Outline**



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**PIN OUT**

PIN NUMBER	NAME	DESCRIPTION
1	+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. <b>The input current requirement is 80mA without any external loads on Pins 4 and 5.</b> Recommended input range is 14V min, 15.0V max.
2	Signal Gnd	Return for +15V supply, and -11.5V, +5V outputs Reference ground for all control signals of the device. All bypass capacitors and loop compensation components must be connected as close as possible to Pin 2. 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 29 and 30.
3	-11.5V Output	<b>-11.5V Output.</b> 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.
4	+5V Output	<b>+5V Output.</b> . 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.
<b>+15V supply should be an isolated power supply</b>		
5	Direction Output	A logic output. High corresponds to motor CW rotation. Low corresponds to motor CCW rotation.
6	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 $f_t = \frac{P \cdot n}{40} \quad \text{Hz}$ Where P is the number of poles, n is the motor speed in rpm.
7	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.

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**PIN OUT (continued)**

8	HC Input	Hall input of Phase C
9	HB Input	Hall input of Phase B
10	HA Input	Hall input of Phase A
11	Flt-Clr	Fault clear input. It can be used to reset a latching fault condition, due to desaturation protection. Pin 21 an active high input. It is internally pulled down by 2.0K Ohms. A latching fault due to desaturation can be cleared by pulling this input high to +5V by 200-500 Ohms, or to +15V by 3-5K Ohms. <b>It is recommended to activate fault clear input for about 300 usec at startup.</b>
12	C <sub>T</sub>	PWM oscillator programming capacitor in nF. Insert a capacitor between this Pin and Pin13 to reduce switching frequency. Use a high quality capacitor for best results. The voltage on C <sub>T</sub> is symmetrical about the signal ground. The peak voltage is internally set to +/-7.2V. The voltage across C <sub>T</sub> is internally applied to the PWM comparator.
13	R <sub>T</sub> /C <sub>T</sub> - Rtn	Return Pin for the PWM oscillator programming capacitor at Pin 12, and oscillator programming resistor at Pin 14.
14	R <sub>T</sub>	PWM oscillator programming resistor in K ohms. Insert a resistor between this Pin and Pin13 to increase switching frequency. $f_s = \frac{200(12 + R_T)}{12R_T(C_T + 1)} \text{ KHz}$ Where R <sub>T</sub> is in K Ohm, and C <sub>T</sub> is in nF
Refer to Fig.9 for R <sub>T</sub> , C <sub>T</sub> connections		
15	EA1- Input	Error amplifier inverting input. This input is available for current loop PI loop calibration.
16	EA1 Output	Output of the error amplifier and is internally connected to the PWM comparator.
The error amplifier non-inverting input, EA1+, is internally connected to signal ground.		
17	IC+	Current Command input. The input command gain is 0.12 V/A.



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**PIN OUT (continued)**

18	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 shall be about 20% higher than the target peak motor current. The gain of Iso is internally set to 0.060 V/A.
19	SD	It is an active low, dual function input/output pin. It is internally pulled high to +15V by 15K ohm. As a low input it shuts down all IGBTs regardless of the Hin and Lin signals. SD is internally activated by the over-temperature shutdown, or desaturation protection SD can be used to shutdown all IGBTs by an external command. An open collector switch shall be used to pull down SD externally. SD can be used as a fault condition output. Low output at SD indicates a latching fault situation.
20	loc-Ref +	Over-current limit reference adjustment.
21	loc-Ref -	Over-current limit adjustment. Connect a resistor Rg KOhms between Pins 20 and 21 to decrease the current amplifier gain and increase peak current limit. The current amplifier gain attenuation due to Rg will be $Kc = \frac{Rg}{Rg + 1.0}$ The output signal gain at Pin 18 will be 0.06*K <sub>C</sub> V/A. The command input gain at Pin17 will be 8.33/K <sub>C</sub> A/V
22	Tco	Analog output of case temperature sensor. The sensor output gain is 0.010 V/°C, with zero DC offset. This sensor can measure only positive °C. The internal impedance of this output is 4.99K ohm The internal block diagram of the temperature sensor is shown in Fig. 19.
23, 24	Phase A Output	Phase A terminals. Both terminals shall be used.
25, 26	Phase B Output	Phase B terminals. Both terminals shall be used.
27, 28	Phase C Output	Phase C terminals. Both terminals shall be used.
29, 30	+VDC Return	Motor supply DC bus return.
31, 32	+VDC	These pins are the motor input power supply positive terminal. Both Pins shall be used. +VDC bus should be bypassed to +VDC Rtn with adequately voltage-rated low ESR capacitor.
Case	NC	Not connected

## Application Report

By  
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### 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.

SMCT6GXXX-XXX 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

$\omega_m$  = 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

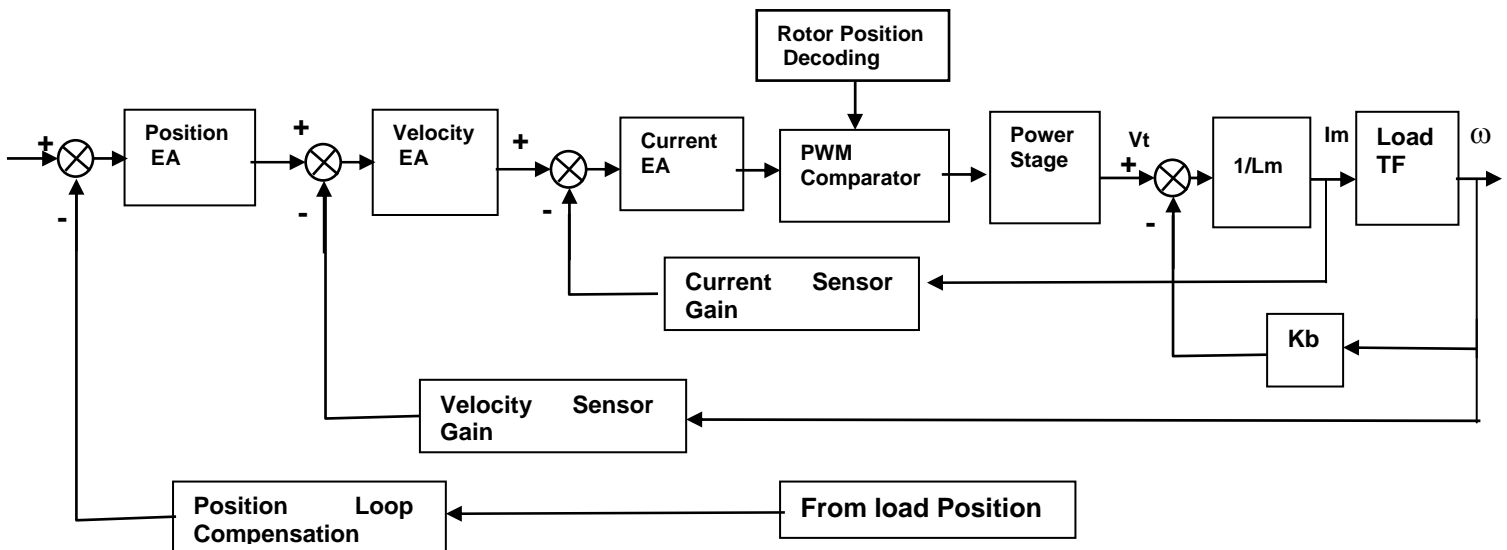
$L_m$  = Motor Inductance

$n_m$  = 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**

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**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  $R_g$  selected to accommodate the current command signal range. Connect a resistor  $R_g$  K Ohms between Pins 20 and 21 to decrease the current amplifier gain and increase peak current limit. The current amplifier gain attenuation  $K_c$  will be

$$K_c = \frac{R_g}{R_g + 1.0} \quad 1$$

The internal current sense resistor  $R_s$  is 0.002 Ohm. This resistor is rated up to 70A maximum. The output signal gain at Pin 18 will be

$$G_o = 30R_s K_c \quad \text{V/A} \quad 2$$

$G_o$  is internally set to 0.060 V/A .

The command input gain at Pin17 will be

$$G_c = \frac{1}{60R_s K_c} \quad \text{A/V} \quad 3$$

$G_c$  is internally set to 8.33 A/V

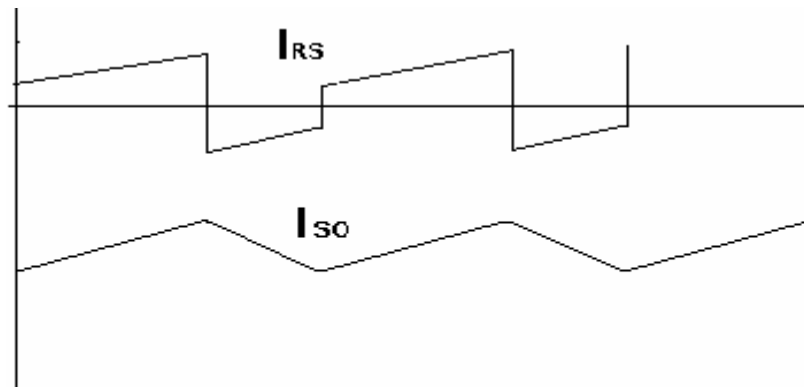
The peak current limit  $I_p$  is

$$I_p = \frac{2.5}{30R_s K_c} \quad \text{A} \quad 4$$

$I_p$  is internally set to 42 A with  $R_g$  open ( $K_c=1$ ).

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<sub>T</sub> 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<sub>p</sub> be the equivalent capacitance of C1 and C2 in parallel.

R<sub>p</sub> be the equivalent resistance of the R3 and R4 in parallel.

The transfer function of the error amplifier EA1 is

$$G_{EA1}(s) = \frac{V_{EA1}(s)}{V_{CA}(s)} = \frac{R_p(R_5C_3s + 1)/R_1}{\left[ R_pR_5C_pC_3s^2 + (R_pC_3 + R_pC_p + R_5C_3)s + 1 \right]} \quad 5$$

or,

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

Where

$$K_1 = \frac{R_p}{R_1} \quad 7$$

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

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

If  $R_p \gg R_5$ ,  $\omega_{PC1}$  and  $\omega_{PC2}$  can be simplified by

$$\omega_{PC1} \approx \frac{1}{R_p(C_p + C_3)} \quad 9$$

$$\omega_{PC2} \approx \frac{(R_p + R_5)}{R_pR_5C_p} \quad 10$$

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

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

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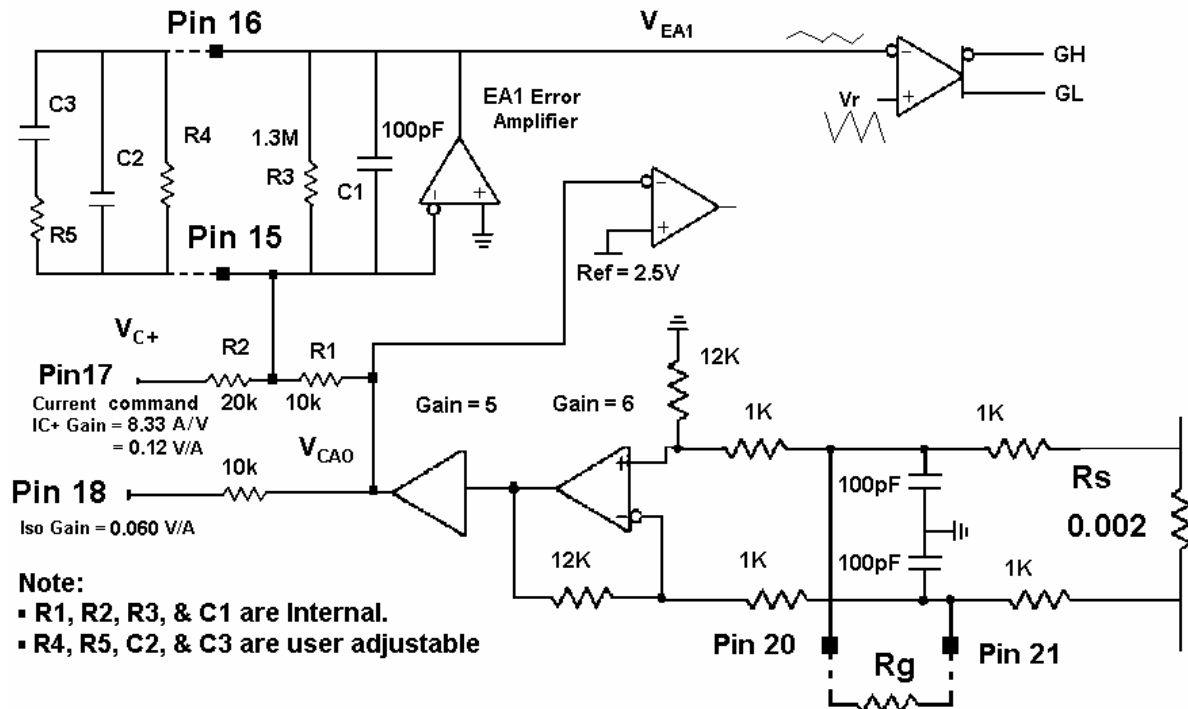


Fig. 6. Current Sense Amplifier & EA1 Compensation Closing Current Loop

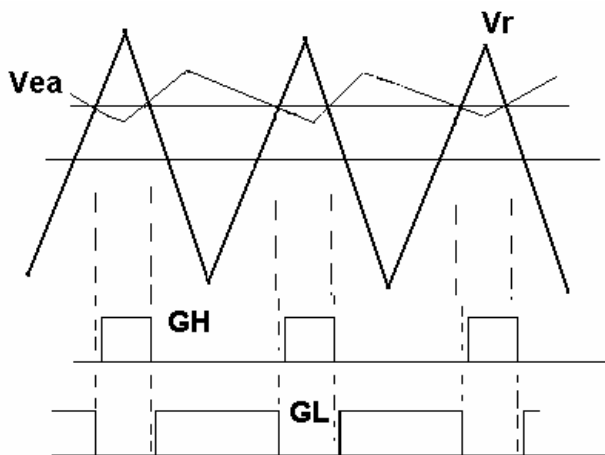


Fig.7. PWM Comparator Inputs & Outputs

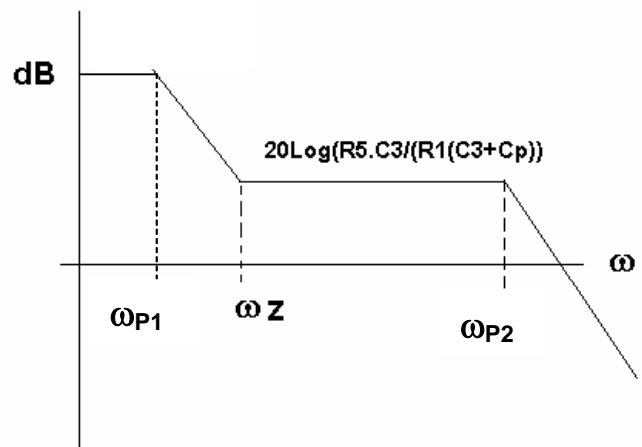
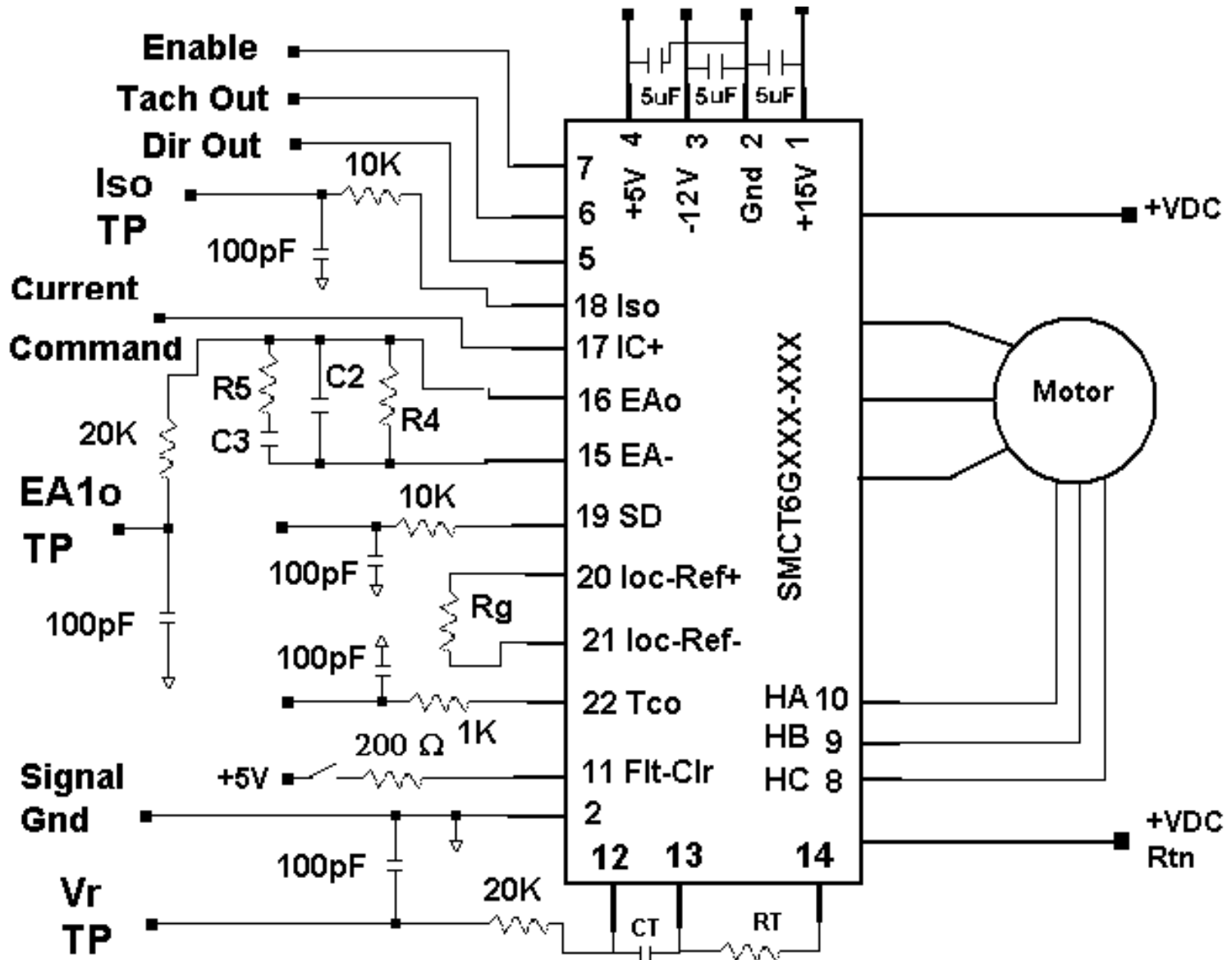


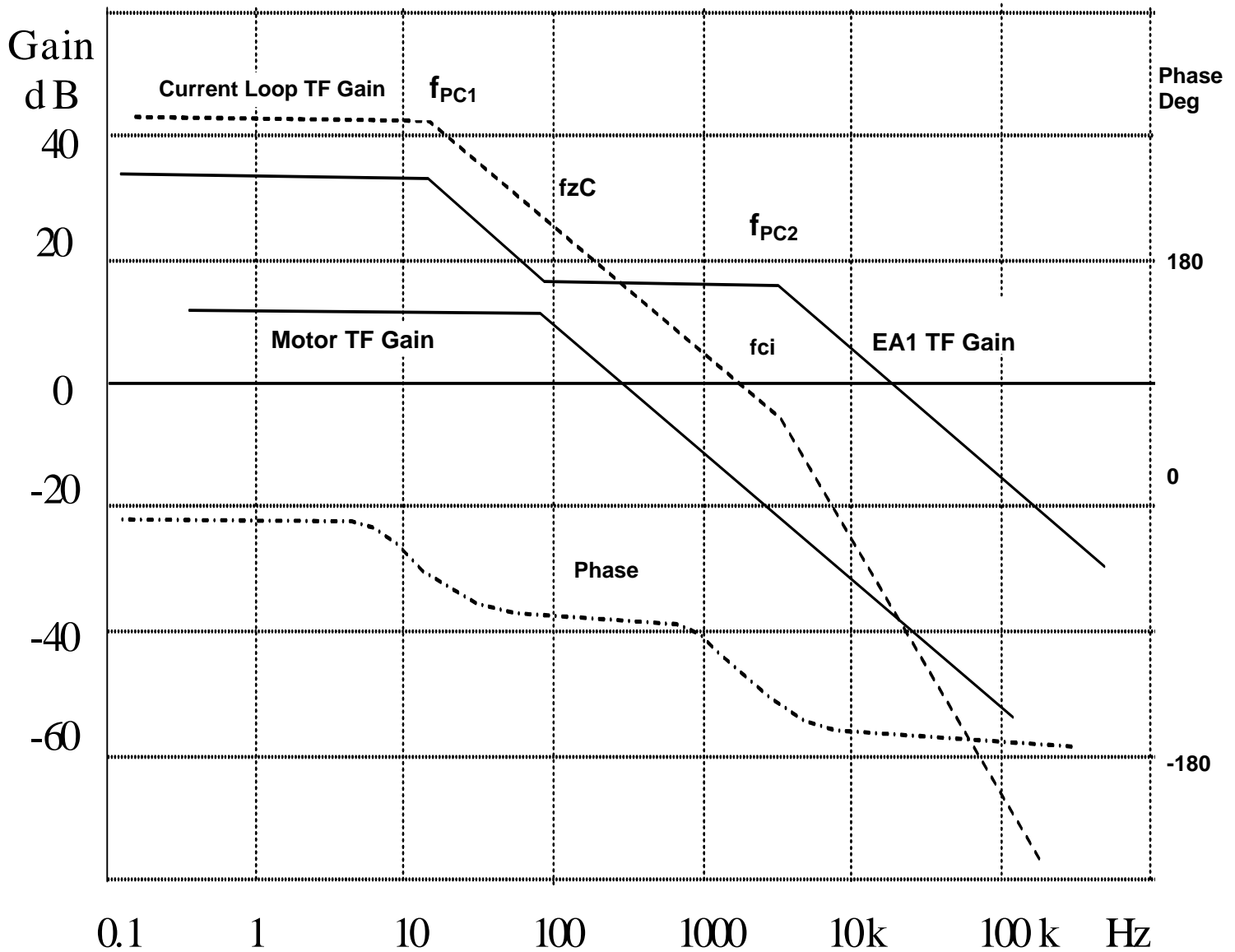
Fig.8. Error Amplifier Transfer Function Bode Plot

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**Fig.9 : Closed Loop Current (Torque) Control**

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

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**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.

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

The current sense amplifier output is

$$V_{CA}(s) = 30K_C R_S I_m(s) \quad 13$$

Substituting 13 in 12

$$\frac{V_{CA}(s)}{30K_C R_S} = \frac{V_{dc} V_{EA1}(s)}{V_S (sL_m + R_m)} \quad 14$$

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

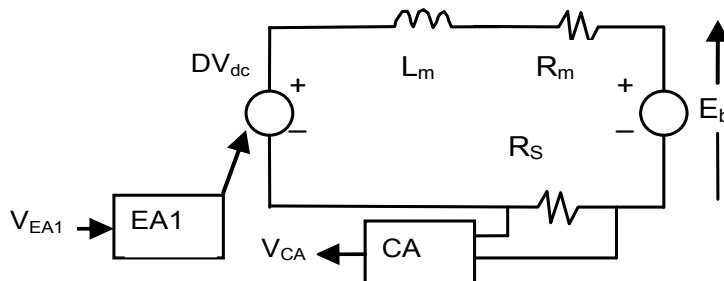
$$G_{Cm}(s) = \frac{V_{CA}(s)}{V_{EA1}(s)} = \frac{30K_C R_S V_{dc}}{V_S R_m (s / \omega_m + 1)} \quad 15$$

where

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

The DC gain of the current loop power section

$$G_{Cm}(s=0) = \frac{30K_C R_S V_{dc}}{V_S R_m} \quad 17$$



**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 SMCT6GXXX-XXX. 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 over-current limit threshold is internally set to 2.5V. 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. In this case

$$\text{At } f_{ci}, \text{ the EA1 maximum output slope} = \frac{R_5 C_3}{R_1 (C_p + C_3)} \frac{30 K_C R_S 2V_{dc}}{L_m} \quad 18$$

The oscillator ramp slope is:

$$\text{Oscillator Ramp Slope} = \frac{4V_s}{T_s} = 4 V_s f_s \quad 19$$

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 15 kHz).

The maximum EA1 integrator gain at 15 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:

$$\frac{R_5 C_3}{R_1 (C_p + C_3)} = \frac{4V_s f_s L_m}{30 K_C R_S 2V_{dc}} \quad 20$$

**Choose  $R_5$  to be about one half the value given by 20 to accommodate for system variations and worst case conditions.** Also, since  $C_3 \gg C_p$ , 20 can be simplified to

$$R_5 = \frac{V_s f_s L_m R_1}{30 K_C R_S V_{dc}} \quad 21$$

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Using  $R_1=10\text{ K}$ ,  $f_s =15\text{ KHz}$ ,  $V_s =7.2\text{ V}$ ,  $R_s=0.002$ ,  $V_{dc} =110$  then maximum value for  $R_5 = 90\text{K}$

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.

$$\frac{30K_C R_S V_{dc}}{2\pi V_S L_m f_{Ci}} \frac{R_5 C_3}{R_1 (C_p + C_3)} = \frac{30K_C R_S V_{dc}}{2\pi V_S L_m f_{Ci}} \frac{2V_S f_s L_m}{30K_C R_S V_{dc}} = 1 \quad 22$$

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

$$f_{Ci} = \frac{f_s}{2\pi} \quad 23$$

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.

$$f_{pm} = \frac{R_m}{2\pi L_m}, \quad f_{zc} = \frac{1}{2\pi R_5 C_3} \quad \text{Hz} \quad 24$$

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

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  $C_p$

$$C_p = \frac{3 \times 10^9}{2\pi R_5 f_s} \quad \text{nF} \quad 26$$

- 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

$$f_{Ci} = \frac{30K_C R_S V_{dc} R_5 C_3}{2\pi V_S L_m R_1 (C_p + C_3)} \quad 27$$

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

$$PM = 180 - \left[ \tan^{-1} \left\{ \frac{f_{Ci}}{f_{pm}} \right\} + \tan^{-1} \left\{ \frac{f_{Ci}}{f_{PC1}} \right\} + \tan^{-1} \left\{ \frac{f_{Ci}}{f_{PC2}} \right\} - \tan^{-1} \left\{ \frac{f_{Ci}}{f_{zc}} \right\} \right] \quad \text{degree} \quad 28$$

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**Design Example**

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

- $R_m = 0.974 \Omega$ ,  $L_m = 1.9 \text{ mH}$
- $K_b = 0.0431 \text{ V/rpm}$ ,  $K_t = 0.304 \text{ lb.ft/A} = 0.412 \text{ N.m/A}$ ,  $T_m = 8.49 \text{ Nm}$  for linear  $K_t$
- $P = \text{Number of poles} = 12$
- $V_{dc} = \text{Supply voltage} = 270 \text{ V}$
- $I_p = \text{Peak motor current} = 30 \text{ A}$
- $J = 4.52\text{E-}4 \text{ Nm.sec}^2$ ,  $= 2.03 \text{ ft.lb. sec}^2$
- $B = \text{Viscous damping} = 6.86\text{E-}5 \text{ Nm/rpm} = 5.06 \text{ ft.lb/rpm}$

Since  $I_p = 30\text{A}$ ,  $R_s = 0.002 \Omega$ , then from equation 4,  $K_C = 1$ .

Let  $R_1 = 10\text{K}$ ,  $V_s = 7.2\text{V}$ ,  $V_{dc} = 270 \text{ V}$ ,  $K_C = 1$ ,  $R_s = 0.002$ .

- 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 K $\Omega$ , use the characteristic equation 29 to calculate  $f_{PC1}$  and  $f_{PC2}$

$$R_p R_5 C_p C_3 s^2 + (R_p C_3 + R_p C_p + R_5 C_3) s + 1 = 0 \quad 29$$

**Table 1**

	$f_{ci}$	$C_p$ pF	$R_5$ K $\Omega$	$C_3$ nF	$f_{zC}$ Hz	$f_{PC1}$ Hz	$f_{PC2}$ Hz	PM Deg
Using Exact values	2240	209	152	3.2	326	36	5580	
Using Standard values	2210	200	150	3.0	354	35	6236	64

**Note:**

- At high line, where the supply is 270 Volts DC,  $f_{ci}$  is 2.21 kHz. The crossover frequency drops to 1.23 kHz at low line, where the supply is approximately 150 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,

$$\text{Damping Ratio} = \text{PM}/100$$

- 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  $(30 \cdot R_s \cdot K_C \cdot I_m)^{2/3}$  is equal to the current command voltage  $(V_{C+})/3$  at frequencies below the crossover frequency. The closed loop current loop transconductance is simply:

$$G_{Ci} = \frac{\Delta I_m}{\Delta V_{C+}} = \frac{1}{60 R_s K_C} \text{ A / V} \quad 30$$

With  $R_s = 0.002 \Omega$ ,  $K_C = 1$ ,  $G_{Ci}$  is internally set to 2 A / V

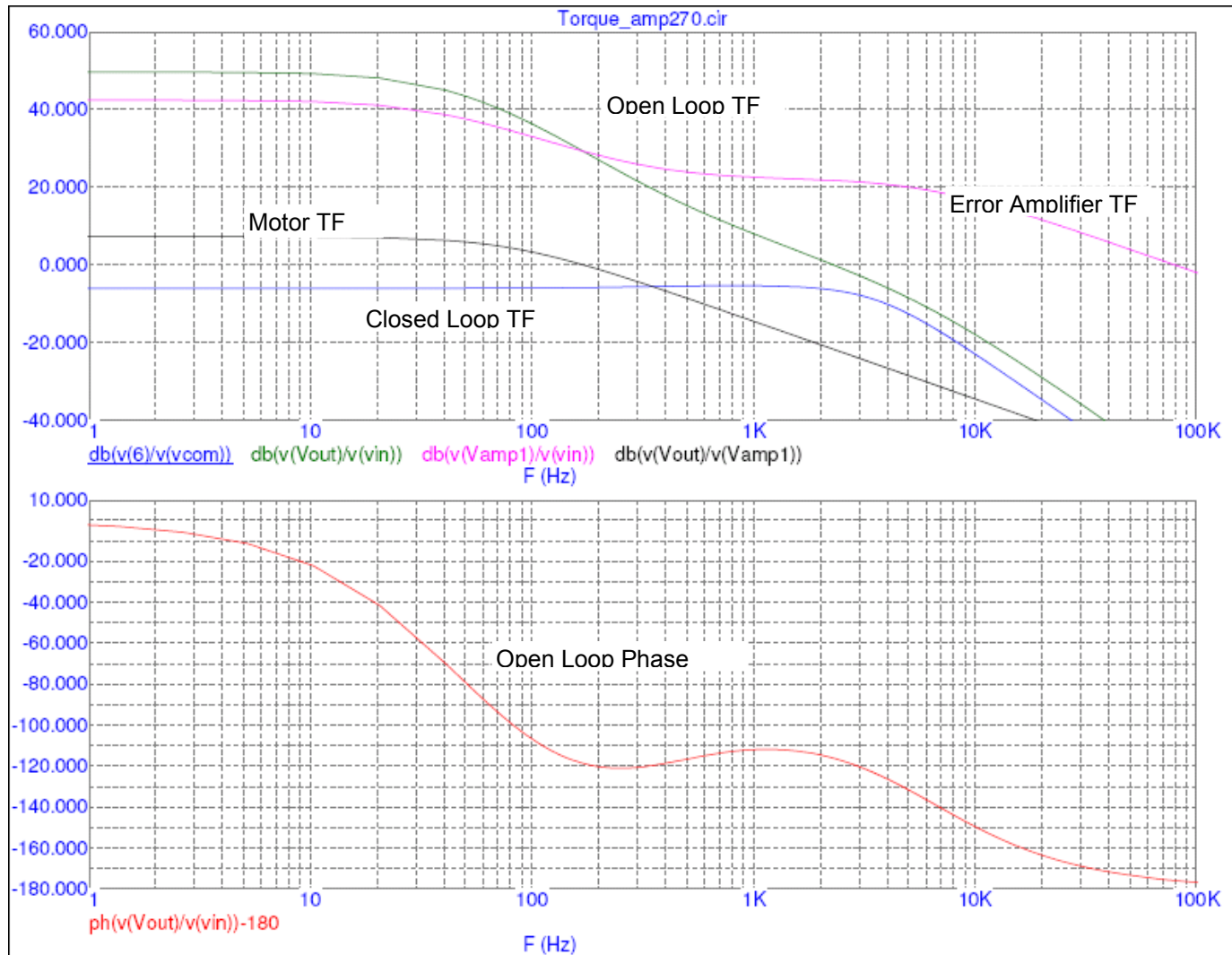
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At the open loop crossover frequency, the transconductance rolls off and assumes a single pole characteristic. The input divider network attenuates current sense amplifier output voltages signal to provide compatibility with typical servo controller, as described in equation (30). The overall amplifier transconductance is set internally to 8.33 amps/volt, allowing full scale current (+/-42 amps) with a +/-5 volt input command.

The closed loop current loop TF is

$$G_{c_i}(jf) = \frac{I_m(jf)}{V_{c+}(jf)} = \frac{1}{60R_s K_C (1 + j f / f_{c_i})} \quad \text{A / V} \quad 31$$

Fig 12 shows the current loop Bode Plots.



**Fig.12 : Current Loop Bode Plots**

## Current Loop Design Summary

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 35A, then  $K_C = 1$ .

1-2 If the peak current  $I_P$  is above 35A, then calculate  $K_C$  from equation 4 assuming  $R_S = 0.002 \Omega$ .

$$K_C = \frac{2.5}{30R_S 1.2I_P} = \frac{34.7}{I_P}$$

1-3 Use equation 1 to calculate  $R_g$ . Then connect  $R_g$  between Pins 20 and 21.

$$R_g = \frac{K_C}{1 - K_C}$$

2- **Second step**, use equation 21 to calculate  $R_5$ .  $f_s$  is internally set to 15KHz. 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  $f_{PC1}$ ,  $f_{PC2}$  using equations 9 and 10. Then verify that  $f_{PC2}$  is more than  $2f_{Ci}$ .

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

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

$$G_{C_i}(jf) = \frac{I_m(jf)}{V_{C_+}(jf)} = \frac{1}{60R_S K_C (1 + j f / f_{C_i})} \quad A / V$$

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 18. The motor should run smoothly and the waveforms should be switching at 15KHz.
- 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 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 16, and the oscillator ramp voltage at Pin 12. 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.

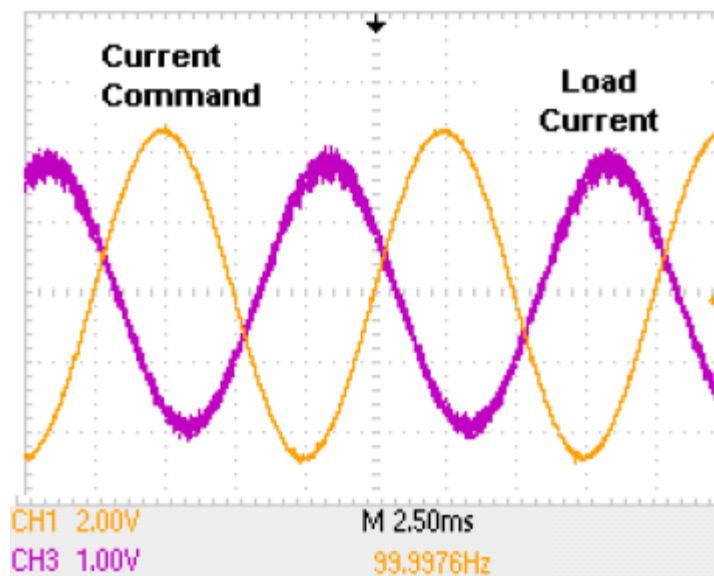
### Warning:

- 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 12, 15, 18, 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 18.

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 Fig. 15. 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. An external 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 the error amplifier, EA2.

## Motor Mechanical Load Modeling

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

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

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

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

Where  $T_L$  is the torque in Nm, and  $\omega_m$  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:

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

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

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 it's 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.**

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

where  $J_m$  is the motor inertia in **Nm.sec<sup>2</sup>**,  $B_V$  is the motor viscous coefficient in **Nm.sec/rad**,  $B_L$  is the load constant in **Nm.sec/rad**, and  $K_1$  is the load fixed torque in **Nm**.

Substituting for  $T_m$ , and  $\omega_m$  from 36 in 37

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

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

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

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

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

where

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

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<sup>2</sup> is analogous to capacitance in F, and Viscous damping in Nm.sec/rad is analogous to resistance in Ω.

where:

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

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

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

Now the motor and the load mechanical parameters are represented by an electrical circuit as shown in Fig.14 and equations 39 to 45.

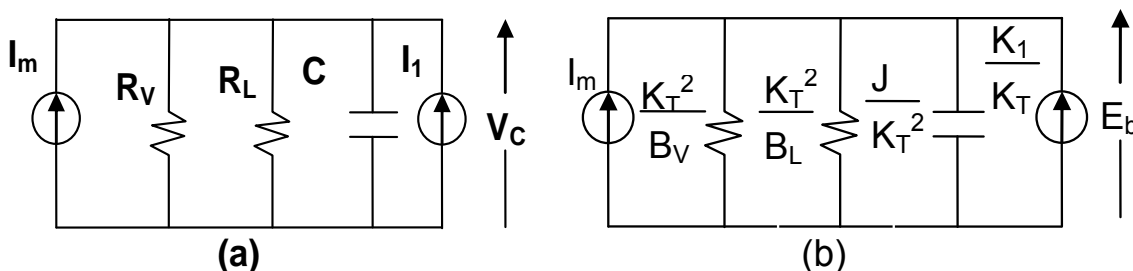
Compensating 41 in 31, results in a small signal TF for the power section of the speed loop.

$$G_{C_m}(jf) = \frac{E_b(jf)}{V_{c+}(jf)} = \frac{R}{60R_s K_c (j f / f_{c_i} + 1)(2\pi RC jf + 1)} \quad \text{V / V} \quad 46$$

The mechanical load pole is  $f_{LP}$

$$f_{LP} = \frac{1}{2\pi RC} \quad \text{Hz} \quad 47$$

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



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



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## 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

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

or,

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

Where

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

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

$\omega_{PV1}$  and  $\omega_{PV2}$  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. 16.

If  $R_9 \gg R_7$ , the EA2 gain at  $\omega_{ZV}$  is

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

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

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

## Closing the Speed (Voltage) Loop

An external error amplifier EA2 can be used to close the speed loop as shown in Figs. 15. 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  $K_{tm}$

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

But

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

Substituting 57 in 56

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

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

$$G_{Cm}(jf)G_m(jf) = \frac{E_{tm}(jf)}{V_{C+}(jf)} = \frac{30K_{tm}R}{60\pi R_s K_T K_C (j f / f_{LP} + 1)(j f / f_{C_i} + 1)} \quad \text{V / V} \quad 59$$

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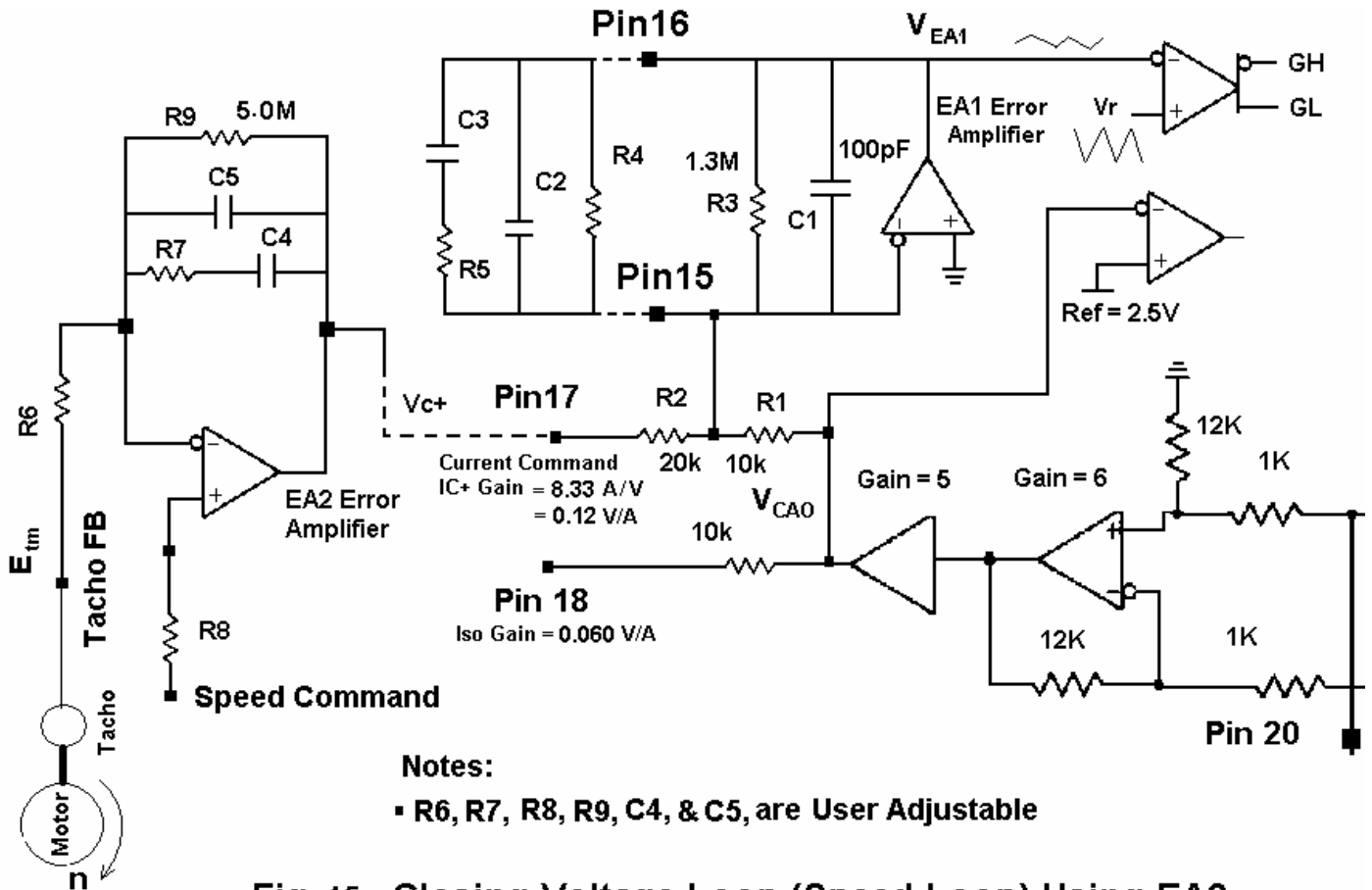
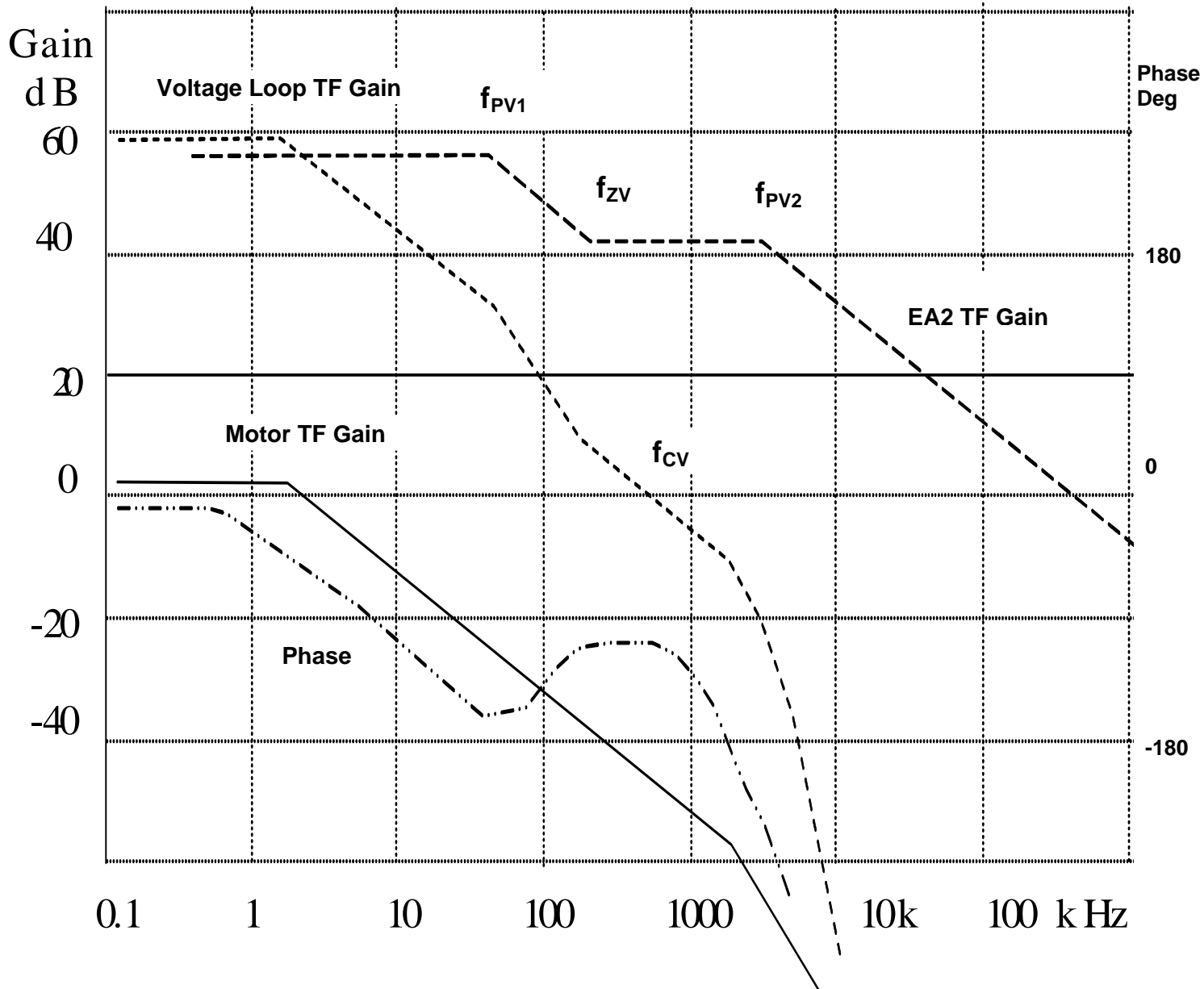


Fig. 15 . Closing Voltage Loop (Speed Loop) Using EA2

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**Fig. 16: 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  $0.5f_{ci}$ .**

$$f_{cv} = 0.3f_{ci} \quad 60$$

- 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 16 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.**

$$\frac{R_7 C_4}{R_6 (C_{pv} + C_4)} \frac{30 K_m}{60 \pi R_s K_T K_C 2 \pi C f_{cv}} = 1 \quad 61$$

**But  $C_4 \gg C_{pv}$ , substituting 60 in 61**

$$\frac{R_7}{R_6} = \frac{1.2 \pi^2 R_s K_T K_C C f_{ci}}{K_m} \quad 62$$

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

- **Place EA2 zero  $f_{zv}$  at  $0.1f_{cv}$**

$$f_{zv} = 0.03 f_{ci} = \frac{1}{2 \pi R_7 C_4} \quad \text{Hz} \quad 63$$

$$C_4 = \frac{10^{12}}{0.06 \pi R_7 f_{ci}} \quad \text{pF} \quad 64$$

- **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.**

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$$f_{PV2} \approx \frac{(R_9 + R_7)}{2\pi R_9 R_7 C_{PV}} = 20 f_{ZV} = 2 f_{CV} = 0.6 f_{Ci} \quad 65$$

Or,

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

But  $C_6 = 10\text{pF}$

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

- Use standard values for  $R_6$ ,  $R_7$ ,  $C_4$ , and  $C_5$ . Use  $R_9 = 5.0\text{M}$ . Also, keep in mind the internal values for  $C_6 = 10\text{pF}$ . 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,

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

$$f_{CV} = \frac{30 K_m R}{60\pi R_s K_T K_C} \frac{R_9}{R_6} \frac{f_{PV1}}{f_{ZV}} f_{LP} \quad 69$$

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

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

## Speed Loop Design Summery

- 1- **First step**, Assume  $R_6 = 10\text{k}$ ,  $R_9 = 5.0\text{M}$ .
- 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.

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**Design Example**

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

- $R_m = 0.974 \Omega$ ,  $L_m = 1.9 \text{ mH}$
- $K_b = 0.0431 \text{ V/rpm}$ ,  $K_T = 0.412 \text{ N.m/A} = 0.304 \text{ lb.ft/A}$ , max  $T_m = 8.49 \text{ Nm}$  for linear  $K_T$
- $P = \text{Number of poles} = 12$
- $V_{dc} = \text{Supply voltage} = 270 \text{ V}$
- $I_p = \text{Peak motor current} = 10 \text{ A}$
- $J_m = 4.52E-4 \text{ Nm.sec}^2$ , = 2.03 ft.lb. sec<sup>2</sup>
- $B_v = \text{Viscous damping} = 6.86E-5 \text{ Nm/rpm} = 5.06E-5 \text{ ft.lb/rpm}$
- =  $60(6.86E-5)/2\pi = 6.5508E-4 \text{ Nm.sec/rad}$
- **Let Tachometer gain be  $K_{tm} = 0.003 \text{ V/rpm}$ .**

- **Let the motor load be 900 watts (3 Nm) at rated speed of 2866 rpm (300 rad/sec), and load torque be expressed as  $0.010 \omega_m$ . Also, let  $J_L = 2J_m$ .**

Use equations 35 to 44 to calculate motor and load parameters

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

$$C = C_m + C_L = 0.00798 \quad \text{F}$$

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

$$R_L = \frac{0.412^2}{0.01432} = 16.97 \quad \Omega$$

$$R = \frac{R_v R_L}{R_v + R_L} = 15.93 \quad \Omega$$

$$E_b = 0.412 \times 209.44 = 123.6 \quad \text{V}$$

- **Notice that motor total internal mechanical power is  $E_b^2/R = 959 \text{ watts}$ , and  $I_m = 7.76 \text{ A}$ .**

**Table 2**  
**Corresponding Speed Loop Compensation Parameters to Table 1**

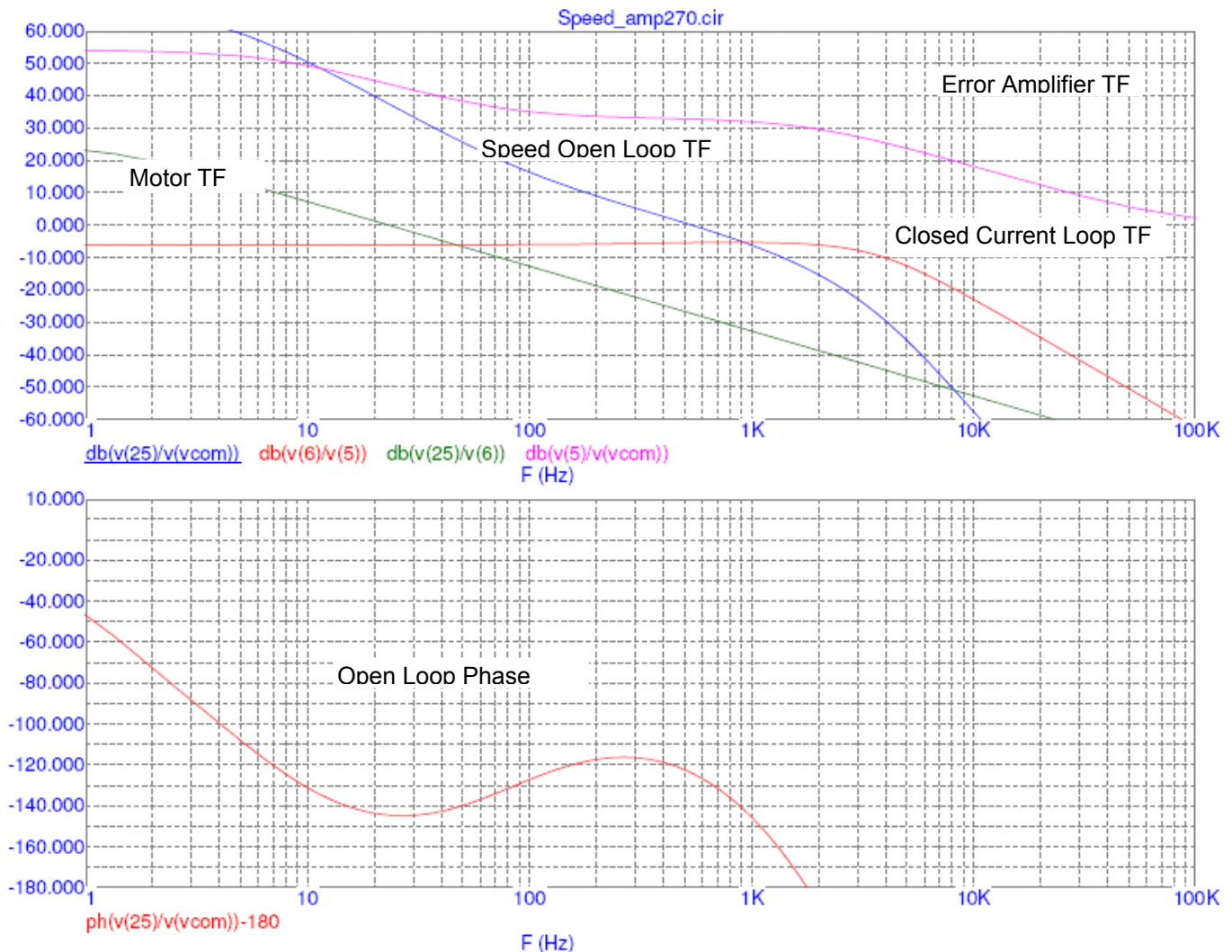
	$f_{Ci}$ Hz	$f_{CV}$ Hz	$C_{PV}$ pF	$R_7$ K $\Omega$	$C_4$ nF	$f_{zV}$ Hz	$f_{PV1}$ Hz	$f_{PV2}$ Hz	PM Deg
Using Exact values	2210	569	233	573	4.2	66	6.5	1385	
Using Standard values		504	200	500	4	79	6.9	1823	54

- **Plot the speed loop Bode Plots as shown in Fig. 17.**

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**Notes:**

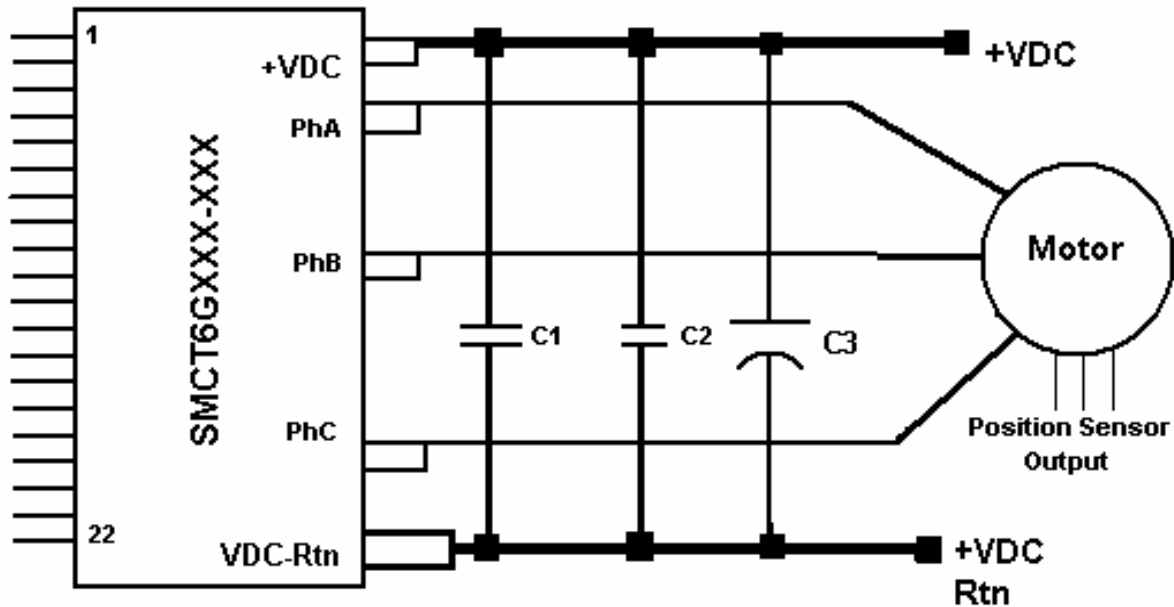
- $R_6 = 10K$ ,  $R_9 = 5.1M$ .
- $f_{LP} = 1.25$  Hz
- $C_{PV} = C_5 + 10$  pF
- At high line, where the supply is 270 Volts DC,  $f_{Ci}$  is 2.21 kHz. The current loop crossover frequency drops to 1.23 kHz at low line, where the supply is approximately 150 Volts DC.
- Notice that the speed loop crossover frequency is independent of DC bus voltage. **At low line voltage  $f_{Ci}$  moves to the left closer to  $f_{CV}$ , reducing speed loop phase margin to 47 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.17 : Speed Loop Bode Plots**



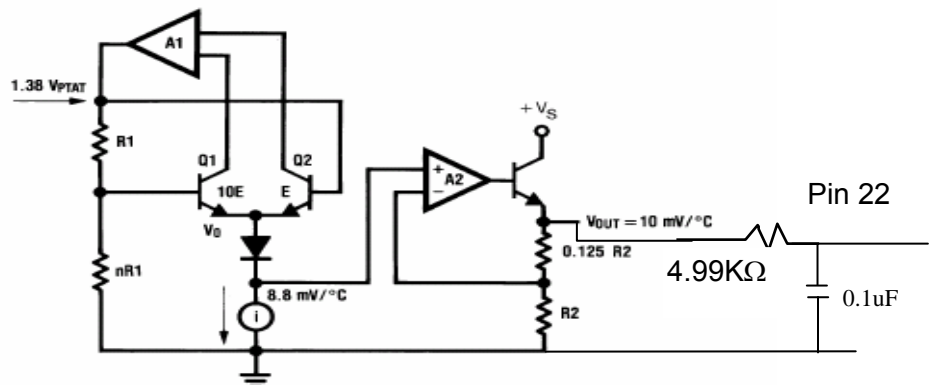
### DC Bus Filtering



**Fig. 18 . DC Bus Bypass Capacitors**

To minimize the circuit parasitic inductance effect on the power stage, the layout of Fig. 18 is suggested. C1 and C2 are 0.5 $\mu$ F ceramic capacitors. Also, a bulk polarized capacitor C3 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 ripples.

### Temperature Sensor Output:



**Fig. 19 Temperature Sensor Internal Block Diagram**

For both negative and positive temperature measurement capability, Contact the Factory.

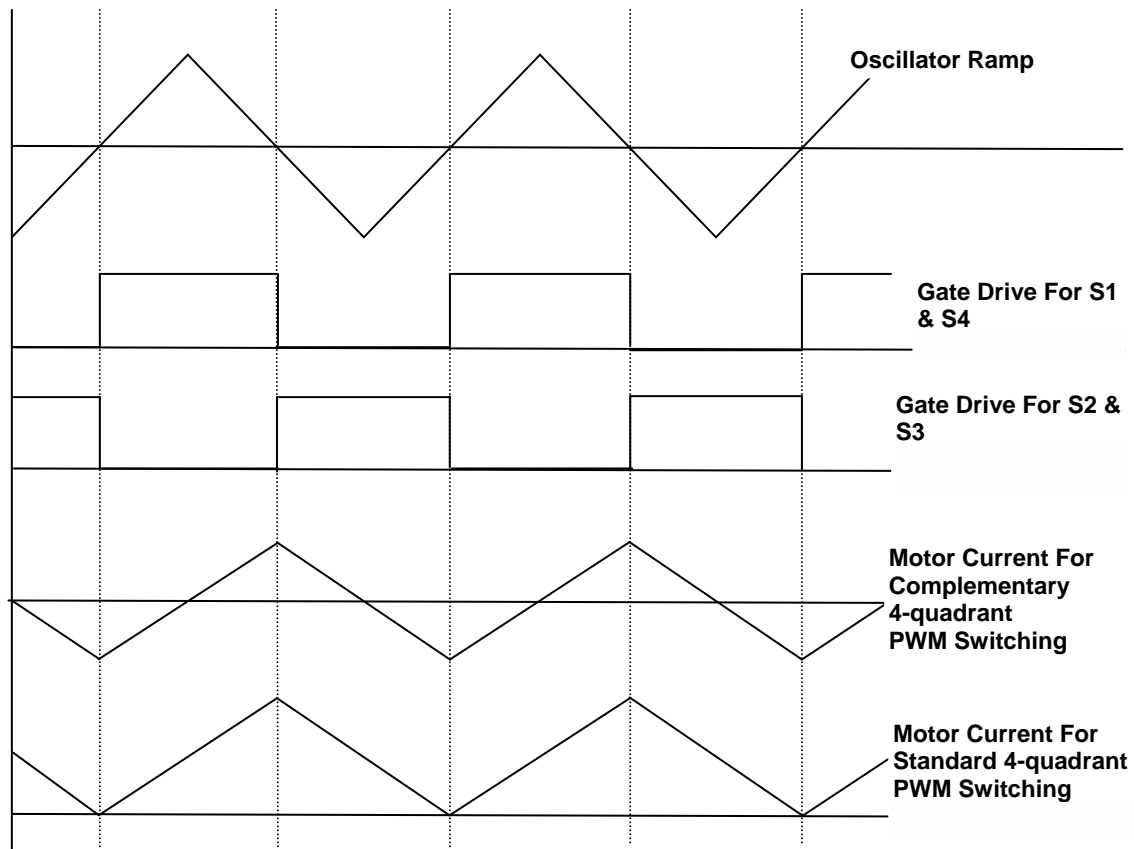
### Complementary PWM Switching

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. 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 SMCT6GXXX-XXX, vs Standard PWM Switching**

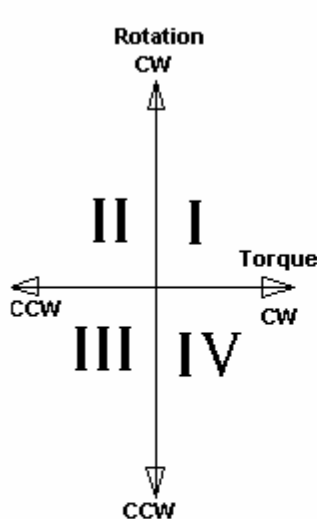
## Standard PWM Switching

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.

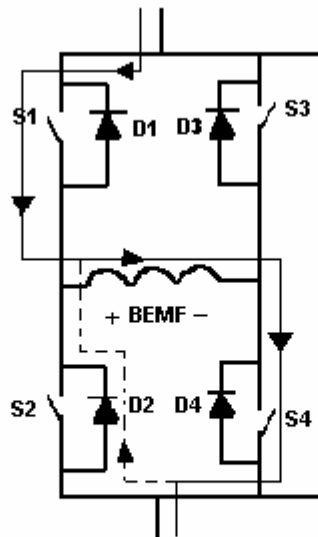
Four-quadrant control provides controlled operation in all quadrants, including II and IV, where torque and rotation are of opposite directions.

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.

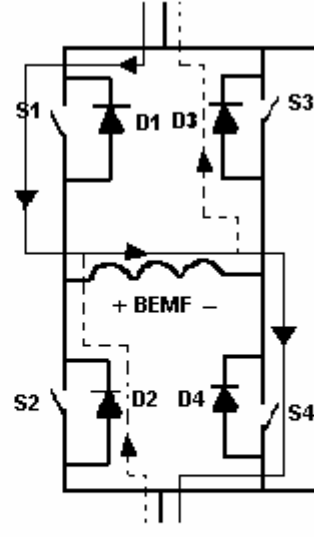
In four-quadrant mode, both upper and lower switches are modulated. Motor current always decays during off time, eliminating any uncontrolled circulating current. In addition, the current always flows through the current sense resistor. Fig. 23 illustrates the current paths during a PWM cycle.



**Fig. 21. Four Quadrants of Operation**



**Fig. 22. Two-Quadrant Forward**



**Fig. 23. Four-Quadrant Forward**

## Operation and Advantages

SMCT6GXXX-XXX 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 through 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.

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**120° Rotor Position Sensing**

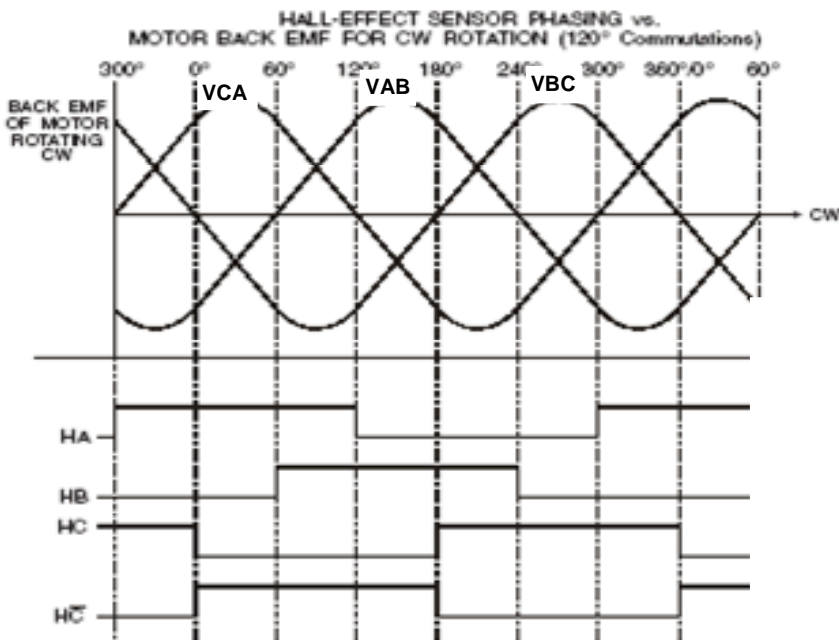
**HA, HB, and HC** are designed to accept rotor position information from hall sensors positioned 120° apart. Motors with 60° position sensing may be used if one or two of the hall-effect sensor signals is inverted prior to connection to the hall-effect inputs. **HA, HB, and HC** inputs are internally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

**60° Rotor Position Sensing**

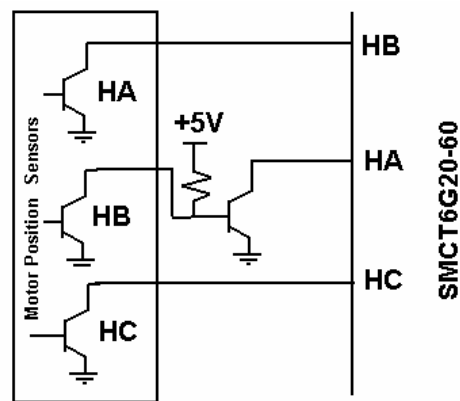
SMCT6GXXX-XXX is designed to operate with 120° position sensing encoding. In this format, the three position sensor signals are never simultaneously high or low. Internal protection logic disables all three Phase Outputs when the Hall-Effect Inputs are set to an illegal condition (i.e., all logic low 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°**

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## ORDERING INFORMATION

### SELECTOR GUIDE TABLE

Part Number	Operating DC Bus Supply Voltage	Peak DC Bus Voltage	RMS Output Current	Peak Output Current
SMCT6G070-060	300	600	60	70
SMCT6G060-120	600	1200	40	60

### PART NUMBERING SYSTEM

SMCT6GXXX-XXX-YY where YY is the sense resistor value as required:

For instance, part number SMCT6GXXX-XXX-30 has a 30mOhm resistor.

Typical current limit sense resistor values are 2, 10, 30, mOhms. Contact the factory for other options.

#### Cleaning Process:

Suggested precaution following cleaning procedure:

If the 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°C. Do not use solvents based cleaners.

#### Soldering Procedure:

Recommended soldering procedure

Signal pins 1-22: 210°C for 10 seconds max

Power pins 23-31: 260°C for 10 seconds max. Pre-warm module to 125°C to aid in power pins soldering

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