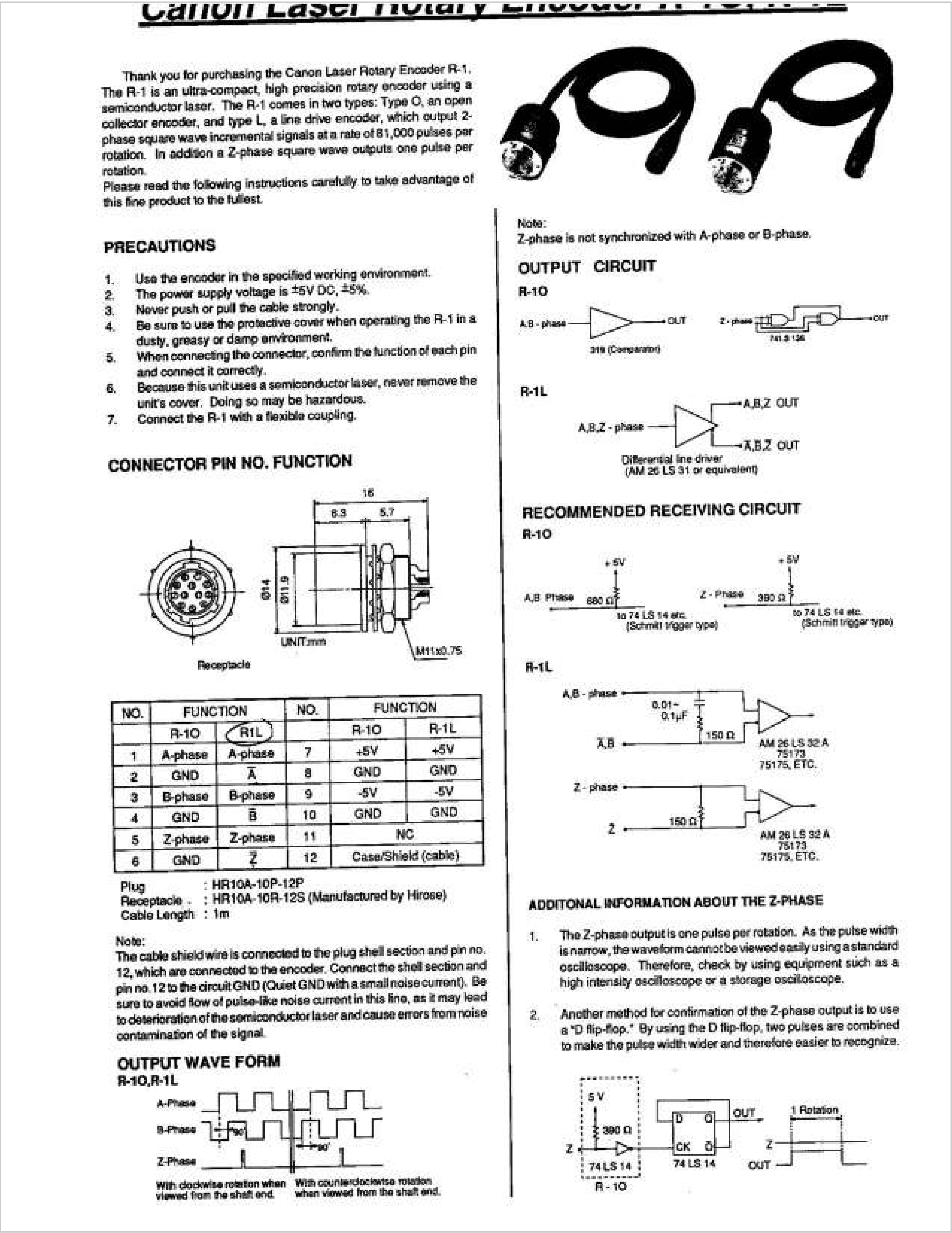
Sunday, April 27, 2008
4:52 PM



Canon Laser Rotary Encoder

Thank you for purchasing the Canon Laser Rotary Encoder R-1. The R-1 is an ultra-compact, high precision rotary encoder using a semiconductor laser. The R-1 comes in two types: Type O, an open collector encoder, and type L, a line drive encoder, which output 2-phase square wave incremental signals at a rate of 81,000 pulses per rotation. In addition a Z-phase square wave outputs one pulse per rotation.
Please read the following instructions carefully to take advantage of this fine product to the fullest.

PRECAUTIONS

1. Use the encoder in the specified working environment.
2. The power supply voltage is ±5V DC, ±5%.
3. Never push or pull the cable strongly.
4. Be sure to use the protective cover when operating the R-1 in a dusty, greasy or damp environment.
5. When connecting the connector, confirm the function of each pin and connect it correctly.
6. Because this unit uses a semiconductor laser, never remove the unit's cover. Doing so may be hazardous.
7. Connect the R-1 with a flexible coupling.

CONNECTOR PIN NO. FUNCTION

NO.	FUNCTION		NO.	FUNCTION	
	R-1O	R1L		R-1O	R-1L
1	A-phase	A-phase	7	+5V	+5V
2	GND	$\overline{A}$	8	GND	GND
3	B-phase	B-phase	9	-5V	-5V
4	GND	$\overline{B}$	10	GND	GND
5	Z-phase	Z-phase	11	NC	
6	GND	$\overline{Z}$	12	Case/Shield (cable)	

Plug : HR10A-10P-12P
Receptacle : HR10A-10R-12S (Manufactured by Hirose)
Cable Length : 1m

Note:
The cable shield wire is connected to the plug shell section and pin no. 12, which are connected to the encoder. Connect the shell section and pin no. 12 to the circuit GND (Quiet GND with a small noise current). Be sure to avoid flow of pulse-like noise current in this line, as it may lead to deterioration of the semiconductor laser and cause errors from noise contamination of the signal.

OUTPUT WAVE FORM

R-1O,R-1L

With clockwise rotation when viewed from the shaft end.

With counterclockwise rotation when viewed from the shaft end.

Note:
Z-phase is not synchronized with A-phase or B-phase.

OUTPUT CIRCUIT

R-1O

R-1L

Differential line driver
(AM 26 LS 31 or equivalent)

RECOMMENDED RECEIVING CIRCUIT

R-1O

R-1L

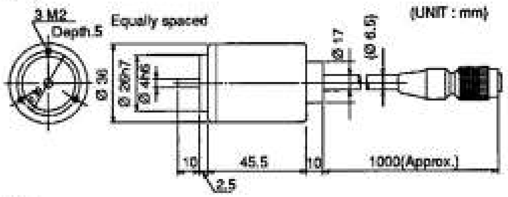
ADDITONAL INFORMATION ABOUT THE Z-PHASE

1. The Z-phase output is one pulse per rotation. As the pulse width is narrow, the waveform cannot be viewed easily using a standard oscilloscope. Therefore, check by using equipment such as a high intensity oscilloscope or a storage oscilloscope.
2. Another method for confirmation of the Z-phase output is to use a "D flip-flop." By using the D flip-flop, two pulses are combined to make the pulse width wider and therefore easier to recognize.

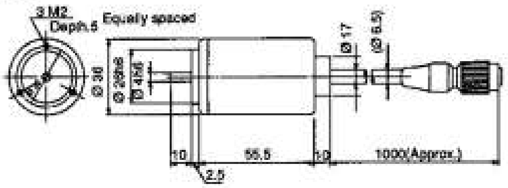
receiving IC needs a wider pulse width increase the width using a one-shot multi-vibrator (74 LS221 etc.).

OUTLINE DIMENSIONS

R-1O



R-1L



SPECIFICATIONS

1. Unit pulse no./rotation:
   81,000 (without electric division)
2. Angel/pulse:
   16arc-sec. (Without electric division)
3. Max. response: 500 kHz (6rps)
4. Output signal:
   81,000 pulses per rotation; 2-phase square wave incremental signals, plus a Z-phase square wave signal.
   a. A-phase and B-phase:
      R-1O: Open collector output
         I_{OL} = 8mA Max. ($V_{OL} \leq 0.4$)
         Recommended load resistance: 680Ω (5V)
         Pull-up voltage: 15V Max.
      R-1L: Line driver output (Balanced)
         Load current: ±20mA Max.
      Difference between A-phase and B-phase: 90° ±10° (Both R-1O and R-1L)
   b. Z-phase: 1 pulse/rotation:
      R-1O: Open collector output
         I_{OL} = 16mA Max. ($V_{OL} \leq 0.4V$)
         Recommended load resistance: 390Ω (5V)
         Pull-up voltage: 5.25V Max.
         Pulse width: 100 ~ 250 nsec.
      R1L: Line driver output (Balanced)
         Load current: ±20mA Max.
         Pulse width: 100 ~ 250 nsec.
5. Outer diameter: 36 mm
6. Overall length: R-1O: 48 mm; R-1L: 58mm
7. Weight: R-1O: 80 g; R-1L: 95 g
8. Max permissible rotating speed: 5,000 rpm
9. Starting torque: 9 g-cm Max
10. Rotar's intertial moment (GD^2): 8g-cm^2
11. Power supply (voltage): ±5V DC, ±5%
12. Power supply (current):
    R-1O: +5V ... 200mA Max. (Recommended load)
    -5V ... 100mA Max.;
    R-1L: +5V ... 250mA Max. (No load)
    -5V ... 100mA.Max.

Semiconductor laser
Wave length: 780nm,
Max.output: 5mW

14. Working environment:
    Operating temperature: 0°C – 50°C
    Storage temperature: -30°C – 80°C
    Humidity: 90% RH Max. (No condensation)
    Vibration: 10G 500 Hz or less
    Impact: 30G 11msec. or less
15. Max. permissible shaft load:
    Radial 0.4 kg
    Thrust 1.0 kg

PRECAUTIONS FOR INCORPORATING THE R-1O, R-1L INTO EQUIPMENT

When connecting the R-1O, R-1L to the drive shaft of equipment, shaft misalignment, shaft deflection, or thrust fluctuation of the drive shaft causes excess force on the bearing causing reduced accuracy and serviceability of the encoder, and damage to the encoder.
Make sure the work load for each model is within the specified permissible load when connecting the encoder to equipment.

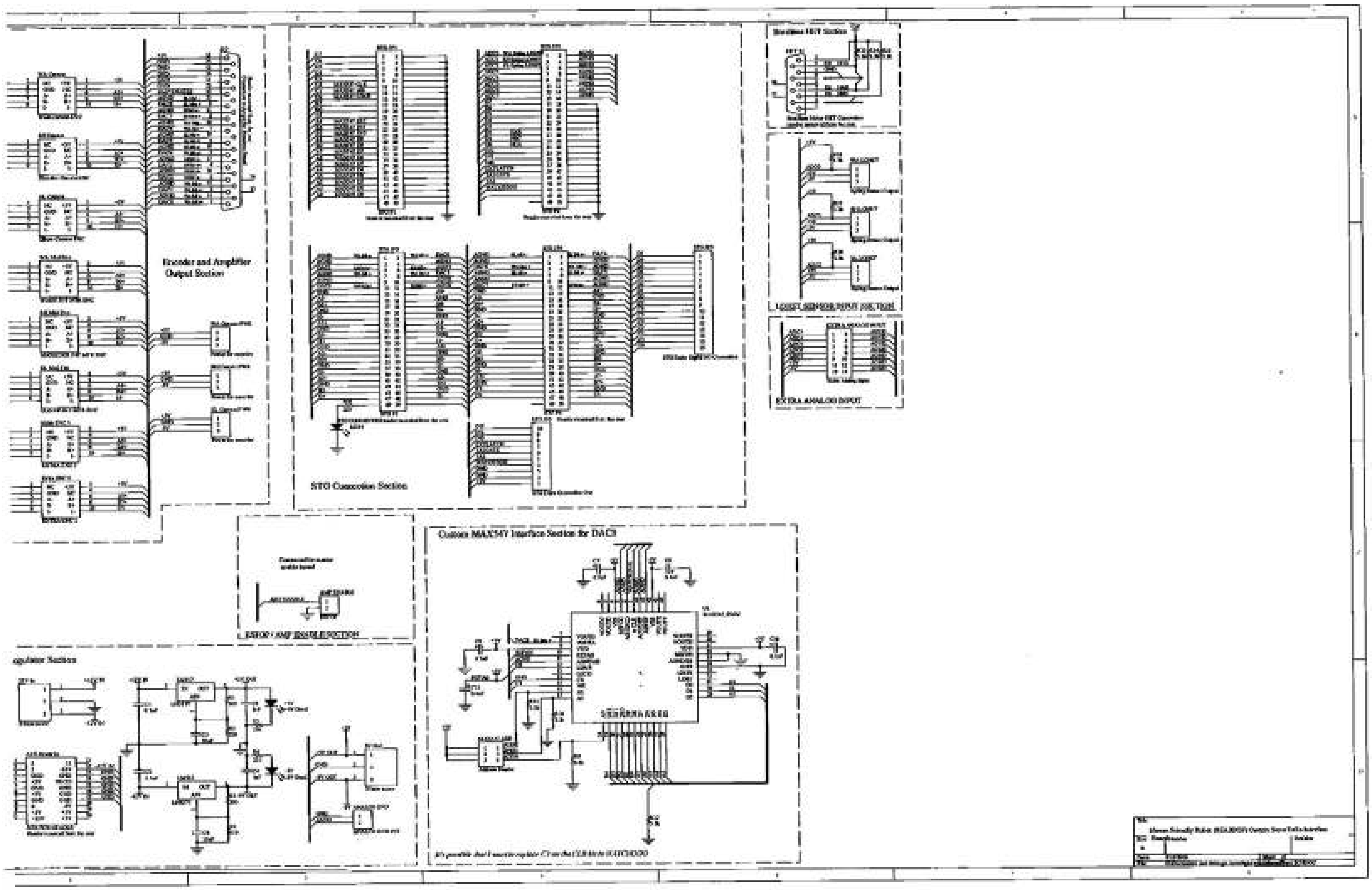
Permissible load	
Radial	0.4kg
Thrust	1.0kg

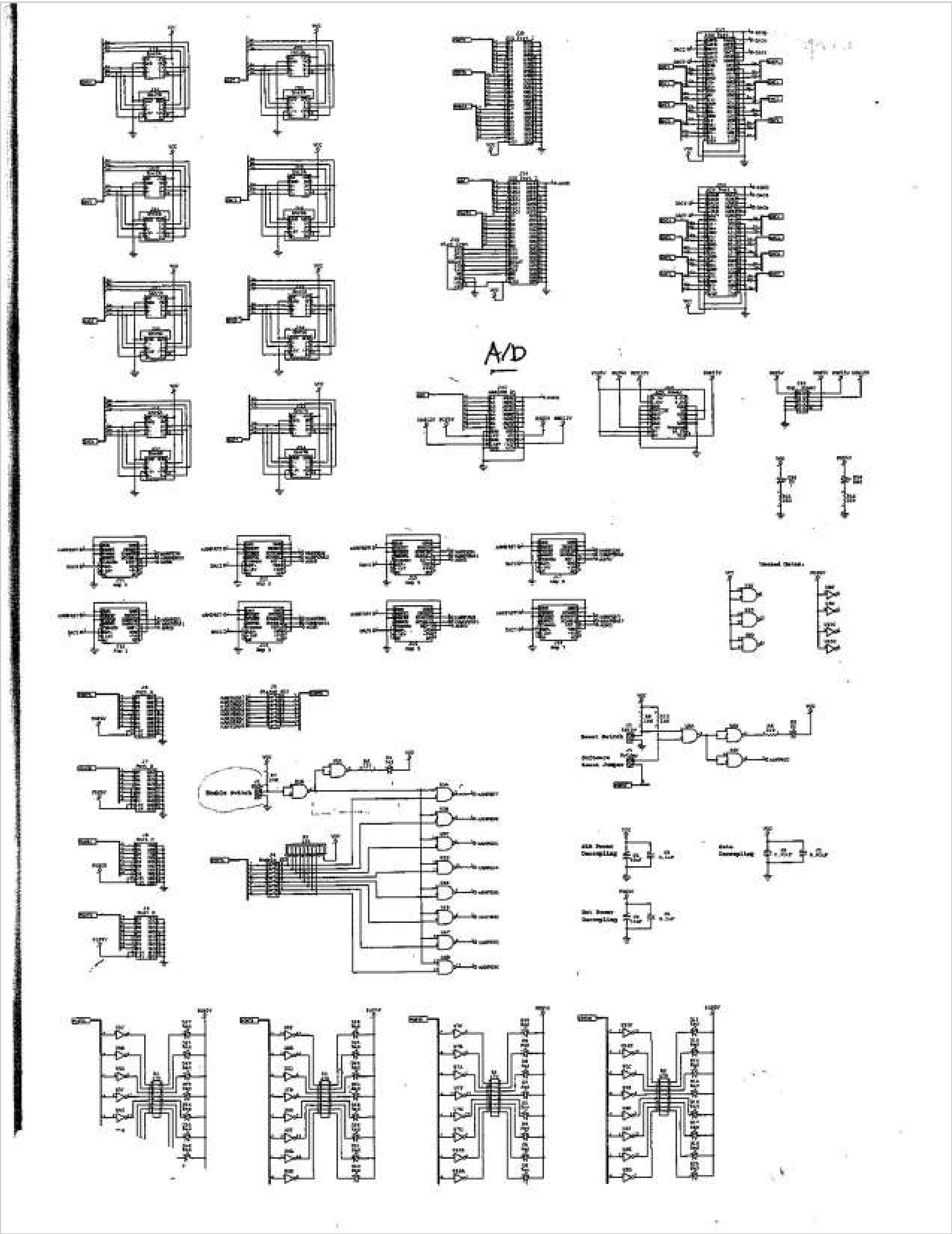
When the encoder is rigidly connected to equipment, the shaft misalignment should be 2μm or less and the thrust fluctuation should be 1μm or less.
When precise shaft alignment is impossible, it is recommended to absorb the shaft misalignment, shaft deflection, or thrust fluctuation by using a flexible coupling.
A flexible coupling with enough torsional rigidity should be selected to obtain precise transmission. When attaching the flexible coupling to the encoder shaft, be very careful not to apply any unnecessary load to the encoder shaft.
The flexible coupling can be attached to the static encoder shaft even if the load caused by the deflection and inclination of the drive shaft exceeds the maximum setting. In this case, however, trouble may occur because undesired stress can be imposed on the encoder shaft when it rotates.

For further details concerning the R-1O, R-1L, please direct inquiries to the following in writing or by telephone.

Canon

CANON U.S.A., INC.
Components Division
One Canon Plaza, Lake Success, N.Y. 11042-9979 U.S.A.
Telephone 516-488-6700 Facsimile 516-354-1114
Telex96-1333
CANON EUROPA N. V.
Industrial Products Division
Unit 3, Brent Trading Centre, North Circular Road, Neasden.
London NW10 OJF, United Kingdom
Telephone (81) 451-4511 Facsimile (81) 459-0331
Telex 295776
CANON AUSTRALIA PTY LTD
Optical Products Division
I Thomas Holt Drive, North Ryde NSW 2113
Telephone (02) 887-0166 Facsimile (02) 887-4484
Telex AA23762
CANON INC.
7-1, Nishi-Shinjuku 2-Chome, Shinjuku-ku, Tokyo 163, Japan.
Mailing address: P.O. Box 5050. Dai-ichi Seimei Building, Tokyo 163, Japan.


Encoder and Amplifier Output Section
STG Connection Section
LIMIT SENSOR INPUT SECTION
EXTRA ANALOG INPUT
ESTOP / AMP ENABLE SECTION
egulator Section


A/D
Enable Switch
Reset Switch
Software Reset Jumper
Unused Gates
Data Decoupling

2-May-05

DB25 connection to amplifier breakout board

I put all the amplifier (brushed and brushless) signals on the one DB25 connector to reduce cable clutter going out of the custom STG interface board.
This however requires the use of an amplifier breakout board but it will be local / close to the amplifiers and thus clutter is minimum.

- *The DB25 straight PC-mount connector is mounted from the rear and thus the PCB footprint is special as the pin alignment is flipped (mirrored).*

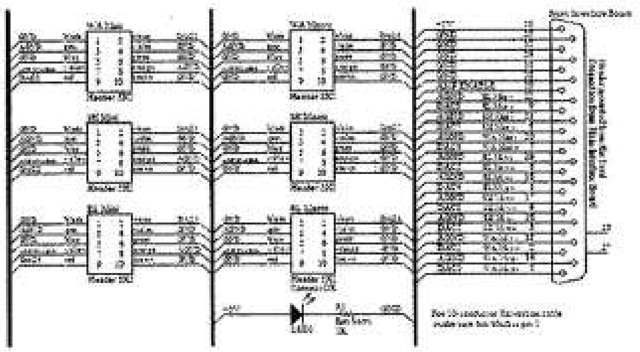
Pin #	Amplifier	Function
1	WA JOINT	u (+)
2	WA JOINT	v (+)
3	SH JOINT	u (+)
4	SH JOINT	v (+)
5	EL JOINT	u (+)
6	EL JOINT	v (+)
7	WA SEA	ref (+)
8	SH SEA	ref (+)
9	EL SEA	ref (+)
10	ENABLE	Amplifier enable (LOW = enable)
11	GND	
12	GND	
13	5V	
14	WA JOINT	u (-)
15	WA JOINT	v (-)
16	SH JOINT	u (-)
17	SH JOINT	v (-)
18	EL JOINT	u (-)
19	EL JOINT	v (-)
20	WA SEA	ref (-)
21	SH SEA	ref (-)
22	EL SEA	ref (-)
23	GND	
24	GND	
25	GND	

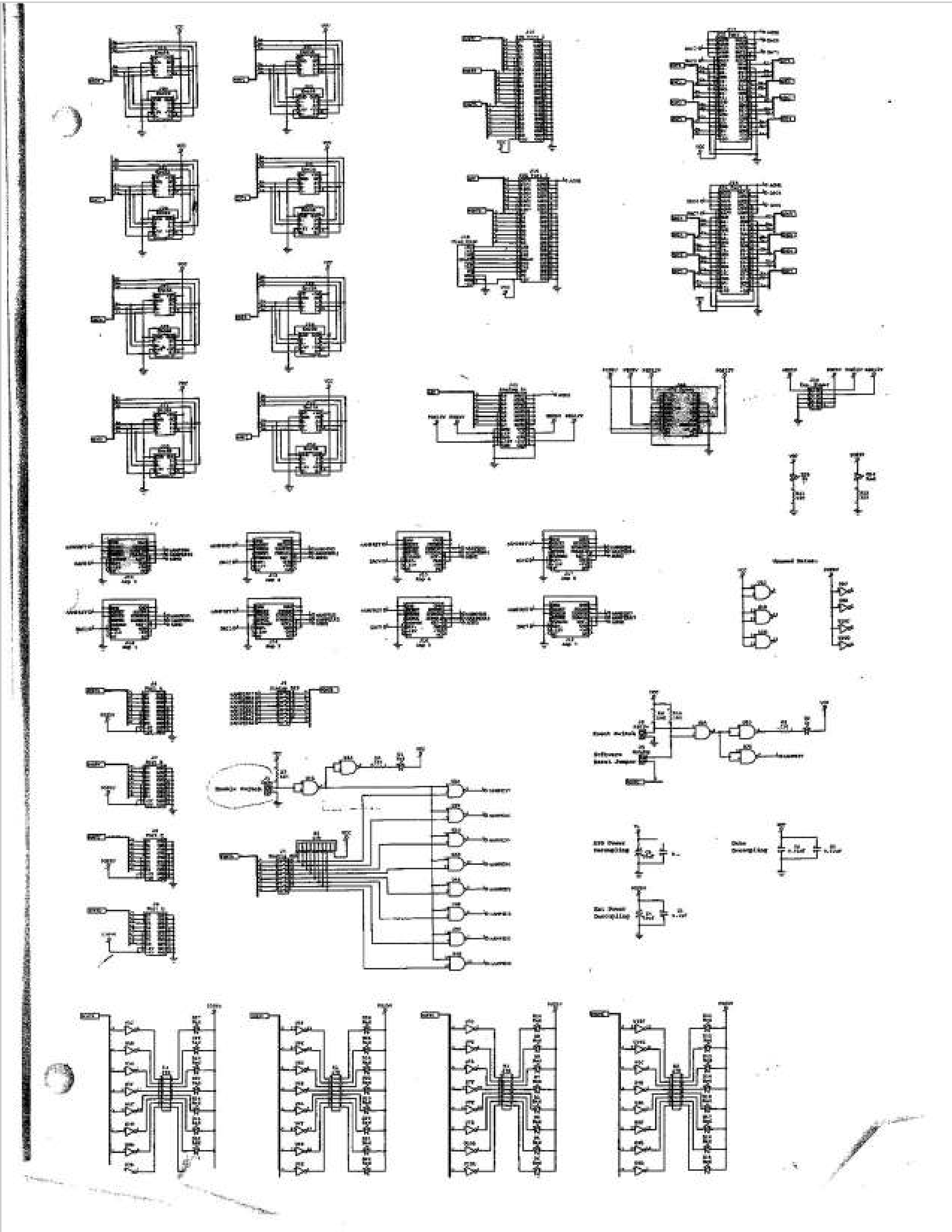
Amplifier Breakout Board Wiring Schematics
August 15 2005
Mini Connections

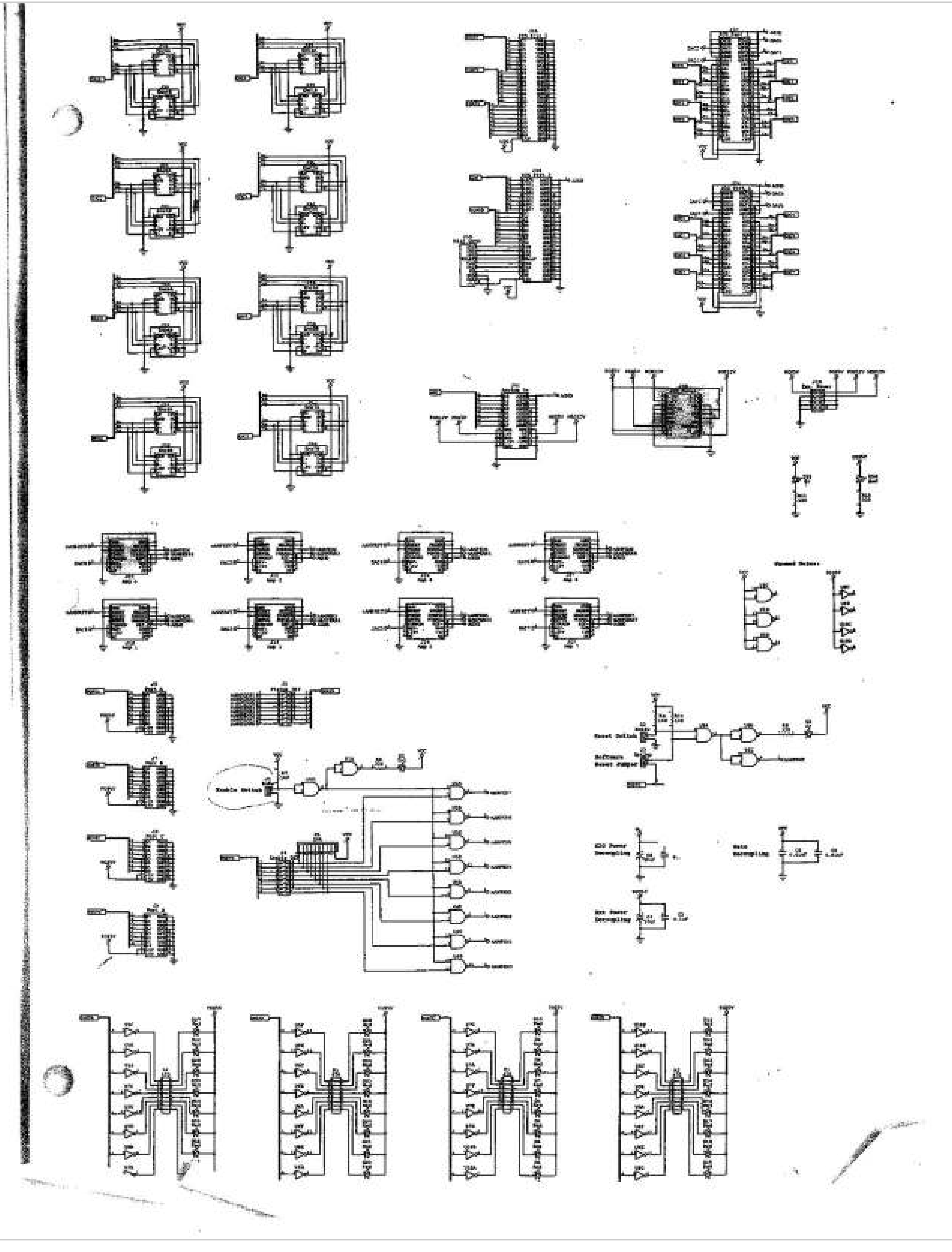
JP Pin#	Description	Color	Amp Header Pin#
1	GND	Black	13 (Signal GND)
2	V(+)	White	3
3	V(-)	Gray	15
4	GND	Violey	12 (GND)
5	GND	Blue	21 (MOTEMP)
6	GND	Green	22 (/RESET)
7	EN	Yellow	6 (EN)
8	U(-)	Orange	14
9	U(+)	Red	2
10	GND	Brown	16 (GND)

Macro Connections

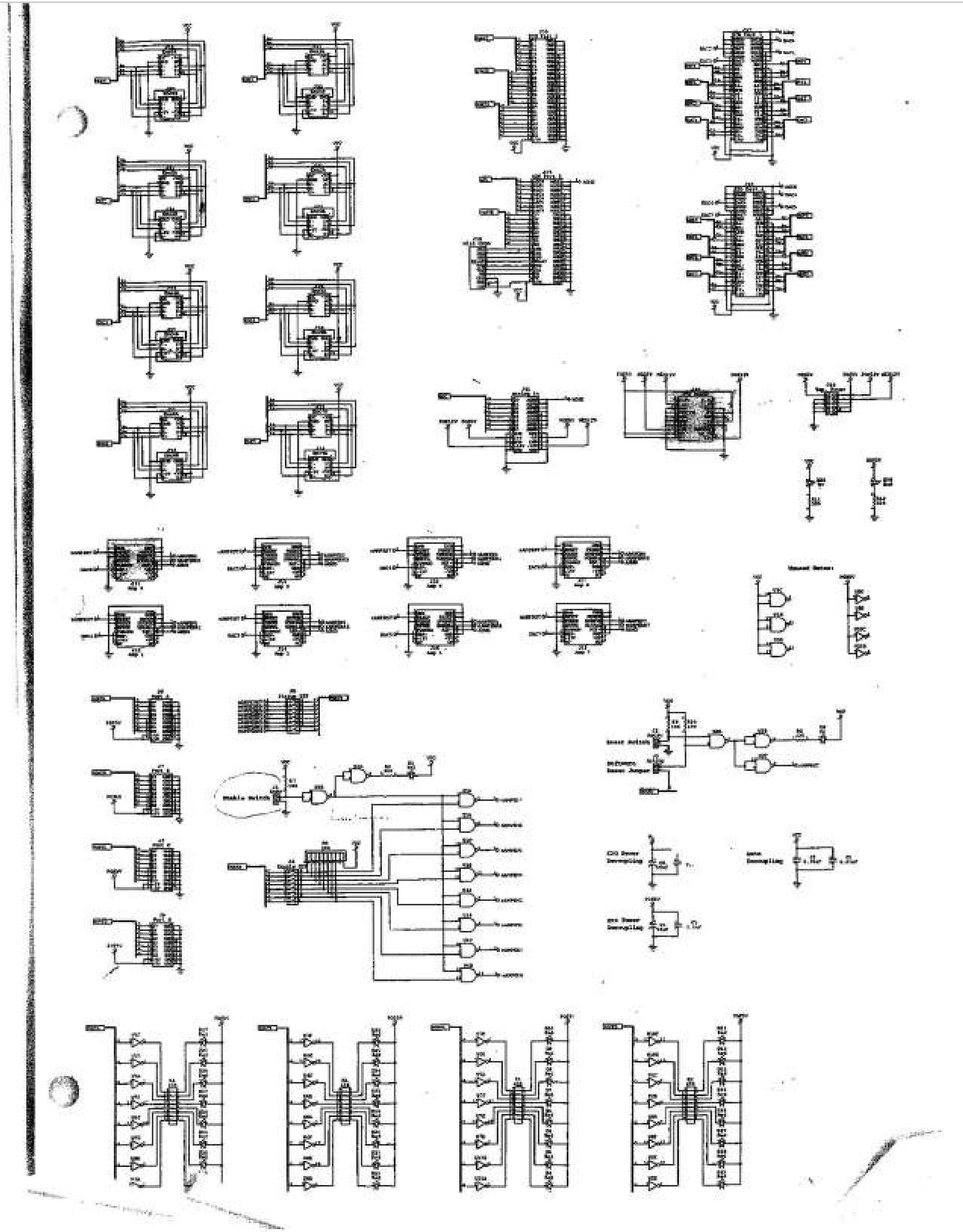
JP Pin#	Description	Color	Amp Header Pin#
1	GND	Black	2 (Signal GND)
2	DAC(+)	White	4
3	DAC(-)	Gray	5
4	GND	Violey	
5	GND	Blue	
6	GND	Green	10 (GND)
7	EN	Yellow	11 (EN)
8	GND	Orange	12 (/POSEN)
9	GND	Red	13 (/NEGEN)
10	GND	Brown	15 (RESET)



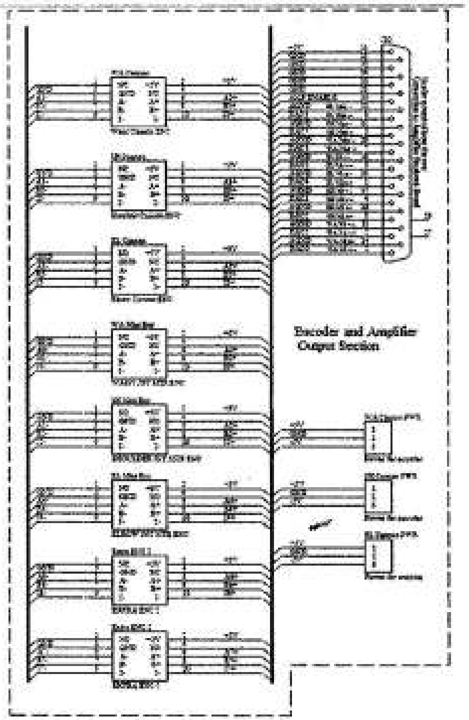

Reset Switch
Software
Reset Jumper
Power Decoupling
Decoupling
Power Decoupling



Reset Switch
Software
Reset Jumper
Enable Switch
Power
Decoupling
Main
Decoupling
Power
Decoupling


Unused Gates:
Enable Switch
Reset Switch
Reset Jumper
CPU Power Decoupling
Decoupling

Encoder Section Schematics



NEW

Cannon Hirose Connector Pinout

cannon pin	function
1	A+
2	A-
3	B+
4	B-
5	Z+
6	Z-
7	+5V
8	GND
9	-5V
10	GND
11	GND
12	shield

Canon custom shielded cable

alpha color	function	cannon pin
red	+5V	7
white	-5V	9
black	GND	8
green	shield	12
orange	GND	10
blue	A+	1
brown	A-	2
yellow	B+	3
gray	B-	4
pink	Z+	5
sand	Z-	6

Encoder header @ Custom STG Board

stg pin	ribbon color
PWR1	
PWR3	
PWR2	
3	red
6	cream
5	orange
8	cream
7	yellow
10	cream
9	green

Cannon Original Cable ==> Alpha Wire Cable ==> Ribbon Cable ==> IDC10 Header

Mini encoders + Macro encoders are direct plug in

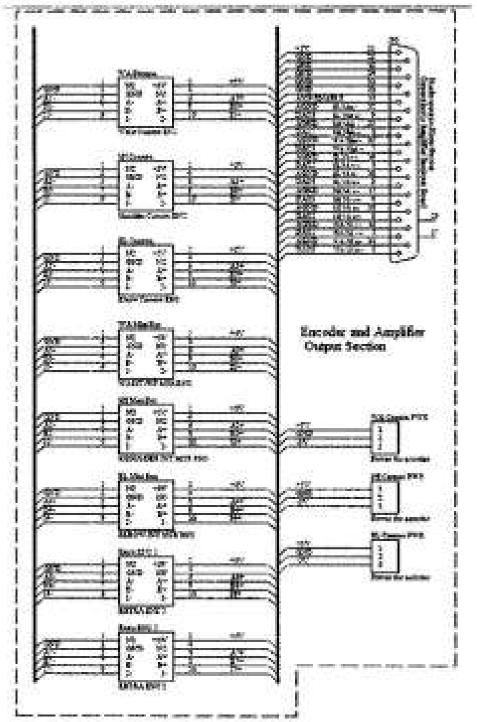
1	A+	blue
2	A-	brown
3	B+	yellow
4	B-	gray
5	Z+	pink
6	Z-	sand
7	+5V	red
8	GND	white
9	-5V	black
10	GND	
11	GND	
12	SHIELD	green

5 orange → brown
6 cream → blue
7 yellow → gray
8 cream → yellow
9 green → sand
10 cream → pink

+5V GND -5V
ORG Red cream

white → GND
black → -5V

Encoder Section Schematics



NEW

Cannon Hirose Connector Pinout		**Canon custom shielded cable**			**Encoder header @ Custom STG Board**	
cannon pin	function	alpha color	function	cannon pin	sig pin	ribbon color
1	A+	red	+5V	7	PWR1	
2	A-	white	-5V	9	PWR3	
3	B+	black	GND	8	PWR2	
4	B-					
5	Z+	green	shield	12		
6	Z-	orange	GND	10	3	red
7	+5V	blue	A+	1	6	cream
8	GND	brown	A-	2	5	orange
9	-5V	yellow	B+	3	8	cream
10	GND	gray	B-	4	7	yellow
11	GND	pink	Z+	5	10	cream
12	shield	sand	Z-	6	9	green

Cannon Original Cable ==> Alpha Wire Cable ==> Ribbon Cable ==> IDC10 Header

Mini encoders + Macro encoders are direct plug in

Waist and shoulder

Peak cont. current 3 A

H_4 280 Ω ✓

H_6 280 Ω

H_{13} [illegible] 330 kΩ

Transconductance (A/V) 1.0

H_3 29.4 k ✓ 30k

H_5 29.4 k ✓ 30k

I^2T limit (200)

H_{14} 150 k

$$\text{count} \cdot 0.0024 \frac{V}{\text{count}} \cdot 0.07 \frac{A}{V}$$

$$9.9\,A \cdot \frac{1}{0.07} \frac{V}{A} \cdot \frac{1}{0.0024} \frac{\text{count}}{V}$$

Elbow

Peak current cont. 3.5 A

H_4 390 Ω

H_6 390 Ω

H_{13} 200 kΩ

Transconductance 1.0

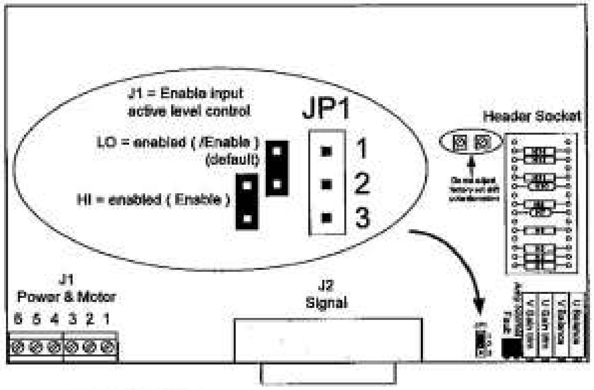
H_3, H_5 30k

I^2T limit 200

H_{14} 150k

MODEL 7225X1
LINE-ISOLATED AC BRUSHLESS SERVO AMPLIFIER
WITH +/-10V ANALOG U-V INPUTS

PC BOARD LAYOUT



HEADER SOCKET COMPONENTS

Part	Value	Remarks
H15	N/a	No function
H14	60.4kΩ	I^2T Current Limit select
H13	0Ω<short>	I^2T Threshold Current select
H12	<out>	Ch. V Current Error Amp hi-frequency roll off
H11	30.1kΩ	Ch. V Current Error Amp proportional gain
H10	100nF	Ch. V Current Error Amp integrator
H9	<out>	Ch. U Current Error Amp hi-frequency roll off
H8	30.1kΩ	Ch. U Current Error Amp proportional gain
H7	100nF	Ch. V Current Error Amp integrator
H6	<out>	Ch. V Continuous Current Limit
H5	75kΩ	Ch. V Transconductance
H4	<out>	Ch. U Continuous Current Limit
H3	75kΩ	Ch. U Transconductance
H2	1.5MΩ	Ch. U Balance Range select
H1	1.5MΩ	Ch. V Balance Range select

HEADER SOCKET COMPONENT SELECTION

LOAD INDUCTANCE

L (mH)	H8, H11 @ 80V	H8, H11 @ 160V	H7, H10
0.4	16.5k	11k	33nF
1	***32.4k***	***18.2k***	***33nF***
3	86.6k	42.4k	33nF
10	249k	124k	33nF
30	750k	392k	33nF

Note: Table values apply with components H9 & H12 not installed. Values in ***bold and italic*** are factory installed.

CURRENT LIMITS

A micro controller uses an I^2T algorithm to monitor to protect against overload conditions. The I^2T overload protection for each channel operates independent of the other. The algorithm detects when the current is any phase exceeds the continuous current limit level set by the header component H13. The I^2T algorithm tracks the energy of the overload (A^2 sec) and when the I^2T limit is reached, the output current is limited to a level set by H4 and H6. The following tables or equations can be used to select header component values to obtain the desired over-current protection setting.

Cont. Current (A)	H4 & H6 (Ohm_)	H13 (Ohm)
10	*<out>*	***0 Ohms (short)***
8	2.5k	16k
6	825	49k
4	383	150k
2	150	<out>

I^2T Limit (A^2sec)	H14 (Ohm)
1250	***0 (short)***
800	16k
450	49k
200	150k
50	<out>

$$H13 = 47.5\text{k ohms} * \frac{(10 - I_{cont})}{(I_{cont} \ -2)}$$

$$H14 = 47.5\text{k ohms} * \frac{\left(6.25 - \sqrt{\frac{I^2T_{limit}}{32}}\ \right)}{\left(\sqrt{\frac{I^2T_{limit}}{32}} - 1.25\right)}$$

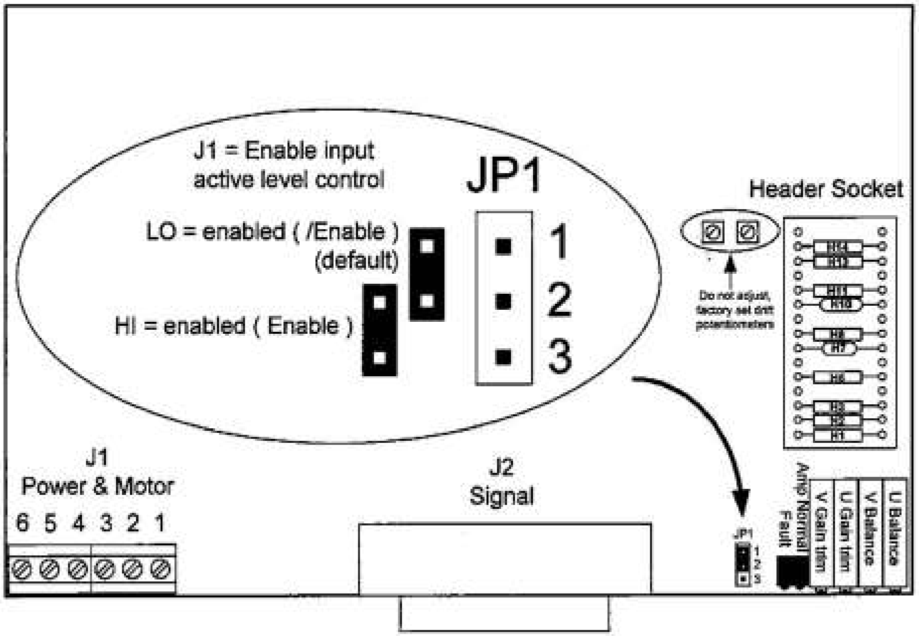
Example: The I^2T set point applies only to the energy delivered to the load over and above the continuous rating of the load. The amplifier's microchip is informed of the continuous current rating of the load via header resistor H13. The I^2T set point is set via header resistor H14. Using a 0 Ohm value for H14 gives an I^2T set point of 1250 A^2*S. If a 0 ohm value is also used for H13, the continuous current setting is set to 10A. This means for a 25 Arms current on either phase U,V, or W, the I^2*T protection will activate (current is forced to continuous limit as set by H4,H6 after a time $T = 1250\ A^2{*}S/(25^2-10^2) = 2.4$ seconds.

BALANCE RANGE AND TRANSCONDUCTANCE SETTINGS

Header components H1 & H2 control the offset range. Default value is 1.5Mohm that gives a range of +/-350mA. The ratio between output current, and the reference voltage at the input is the *transconductance* of the amplifier. It is measured in Amps/Volt, and is controlled by components H3 & H5. The chart below gives some common settings.

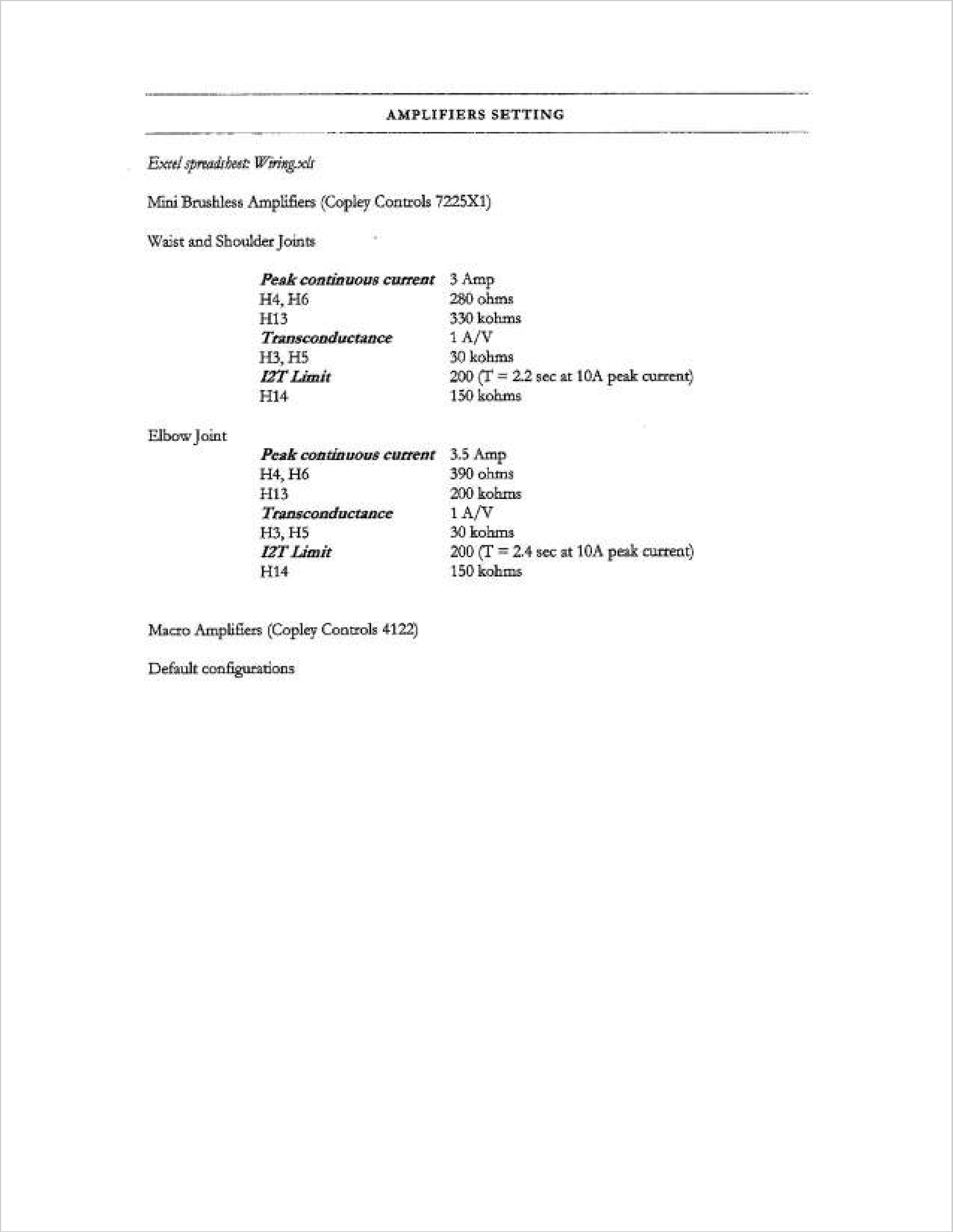
Gain (A/V)	H3 & H5
2.5	102k
2.0	***75.0k***
1.5	59k
1	29.4k
0.5	14.3k

PC BOARD LAYOUT



HEADER SOCKET COMPONENTS

Part	Value	Remarks
H15	N/a	No function
H14	86.6kΩ	I^2T Current Limit select
H13	0Ω<short>	I^2T Threshold Current select
H12	<out>	Ch. V Current Error Amp hi-frequency roll off
H11	30.1kΩ	Ch. V Current Error Amp proportional gain
H10	100nF	Ch. V Current Error Amp integrator
H9	<out>	Ch. U Current Error Amp hi-frequency roll off
H8	30.1kΩ	Ch. U Current Error Amp proportional gain
H7	100nF	Ch. V Current Error Amp integrator
H6	<out>	Ch. V Continuous Current Limit
H5	75kΩ	Ch. V Transconductance
H4	<out>	Ch. U Continuous Current Limit
H3	75kΩ	Ch. U Transconductance
H2	1.5MΩ	Ch. U Balance Range select
H1	1.5MΩ	Ch. V Balance Range select



AMPLIFIERS SETTING

Excel spreadsheet: Wiring.xls

Mini Brushless Amplifiers (Copley Controls 7225X1)

Waist and Shoulder Joints

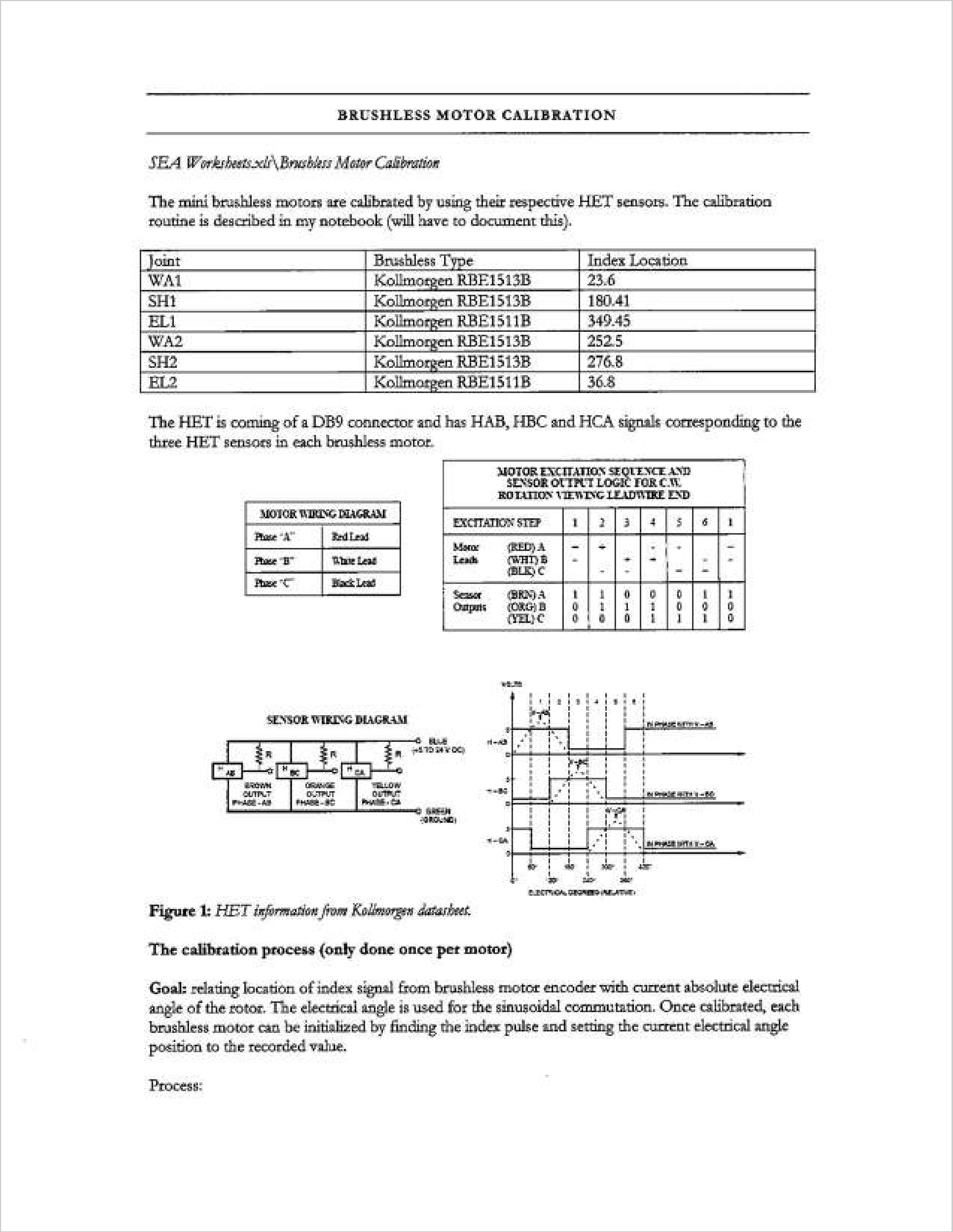
Peak continuous current	3 Amp
H4, H6	280 ohms
H13	330 kohms
Transconductance	1 A/V
H3, H5	30 kohms
I2T Limit	200 (T = 2.2 sec at 10A peak current)
H14	150 kohms

Elbow Joint

Peak continuous current	3.5 Amp
H4, H6	390 ohms
H13	200 kohms
Transconductance	1 A/V
H3, H5	30 kohms
I2T Limit	200 (T = 2.4 sec at 10A peak current)
H14	150 kohms

Macro Amplifiers (Copley Controls 4122)

Default configurations



BRUSHLESS MOTOR CALIBRATION

SEA Worksheets.xls\Brushless Motor Calibration

The mini brushless motors are calibrated by using their respective HET sensors. The calibration routine is described in my notebook (will have to document this).

Joint	Brushless Type	Index Location
WA1	Kollmorgen RBE1513B	23.6
SH1	Kollmorgen RBE1513B	180.41
EL1	Kollmorgen RBE1511B	349.45
WA2	Kollmorgen RBE1513B	252.5
SH2	Kollmorgen RBE1513B	276.8
EL2	Kollmorgen RBE1511B	36.8

The HET is coming of a DB9 connector and has HAB, HBC and HCA signals corresponding to the three HET sensors in each brushless motor.

MOTOR WIRING DIAGRAM	
Phase "A"	Red Lead
Phase "B"	White Lead
Phase "C"	Black Lead

MOTOR EXCITATION SEQUENCE AND SENSOR OUTPUT LOGIC FOR C.W. ROTATION VIEWING LEADWIRE END

EXCITATION STEP		1	2	3	4	5	6	1
Motor Leads	(RED) A	−	+		−	−		−
	(WHT) B	−		+	−		−	−
	(BLK) C		−	−		−	−	
Sensor Outputs	(BRN) A	1	1	0	0	0	1	1
	(ORG) B	0	1	1	1	0	0	0
	(YEL) C	0	0	0	1	1	1	0

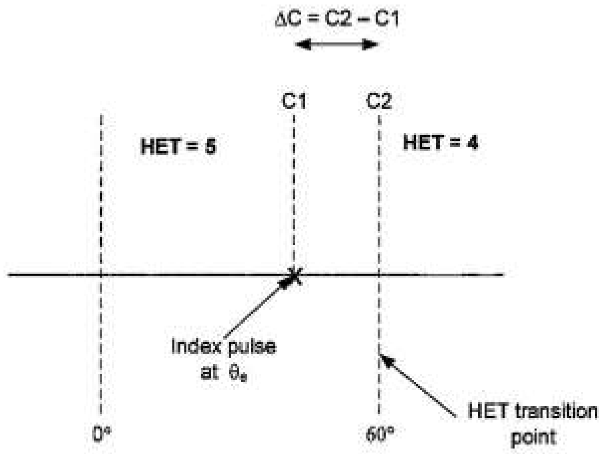
Figure 1: *HET information from Kollmorgen datasheet.*

The calibration process (only done once per motor)

Goal: relating location of index signal from brushless motor encoder with current absolute electrical angle of the rotor. The electrical angle is used for the sinusoidal commutation. Once calibrated, each brushless motor can be initialized by finding the index pulse and setting the current electrical angle position to the recorded value.

Process:

- Run *initWithIndex()* from bldc.cpp
- Rotate the joint to calibrate until an index pulse is found and a message is displayed on the screen. The current HET state (the lower 3 bits of PORTD) and the current encoder count (C1) are displayed.
- Keep rotating until HET changes state and take note the new encoder count (C2)
- The difference between C2 and C1 is the distance in count between the index pulse and the HET transition point. Calculate the corresponding angle by doing the conversion from encoder count to mechanical angle and then to electrical angle $\Delta\theta_e$ (1 mech. degree = 6 elec. degrees). The transition point angle can be found by noting the previous and current states of the HET.
- Look at the HET chart from the data sheet for the electrical angle at the transition point. Subtract (or add, depending on the direction of rotation) $\Delta\theta_e$ from it to get the current the current electrical angle θ_e



Example
For WA1:

$C_1=1361$

$C_2=1499$

$\Delta C=C_2-C_1=138$ counts

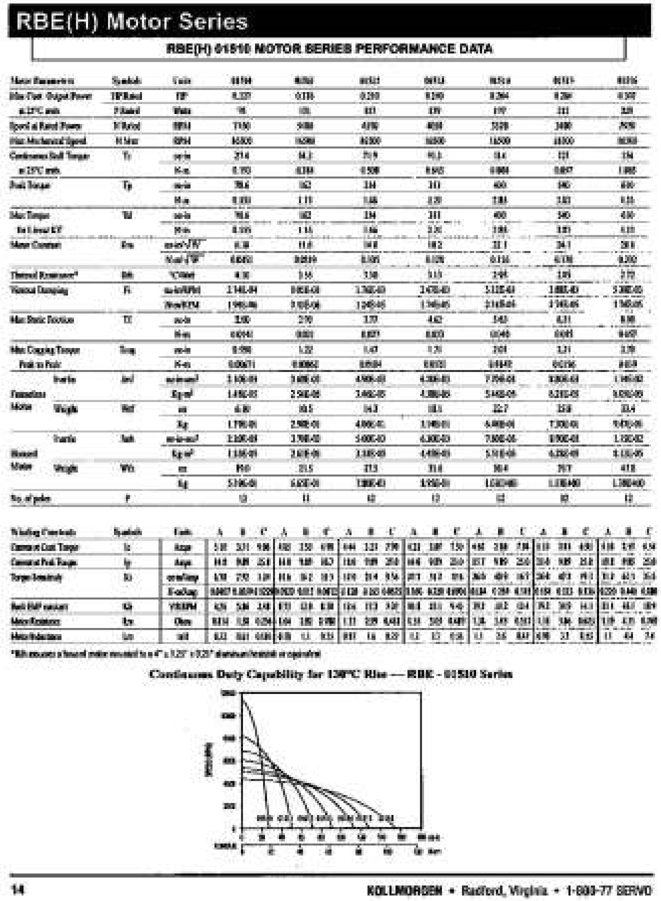
$$\Delta\theta_e=138\text{ counts}\,\frac{1\text{ rev}}{8192\text{ counts}}\,\frac{360\text{ mech deg}}{1\text{ rev}}\,\frac{6\text{ electrical deg}}{1\text{ mechanical deg}}$$

$= 36.38°$

Transition point=60° (from HET = 5 to HET = 4)

$\theta_e=60°-36.38°=23.62°$

$\therefore$ index for WA1 is at 23.62°



RBE(H) Motor Series

RBE(H) 01510 MOTOR SERIES PERFORMANCE DATA

Continuous Duty Capability for 130°C Rise — RBE - 01510 Series

14 KOLLMORGEN • Radford, Virginia • 1-800-77 SERVO

Elbow

3.5A peak H4, H6 390 ~~370~~ Ω

H13 200 kΩ

Transconductance 1 H3, H5 30k

I^2T 200 H14 150k

N-Series ServoDisc

PERFORMANCE DATA

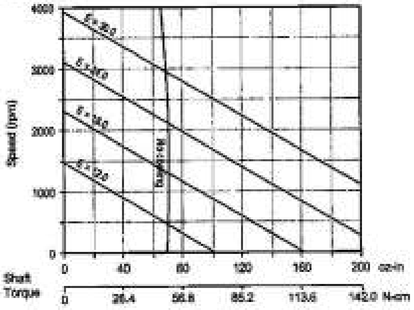
Performance Specifications	Symbol	Units	N9M4	N9M4T	N9M4LR	N9M4LRT	N12M4	N12M4T	N12M4LR	N12M4LRT
Peak Torque	T_p	oz-in	760	692	729	663	1598	1386	1522	1320
		N-cm	537	489	515	468	1128	979	1075	932
Rated Speed	N	RPM	3000	3000	3000	3000	3000	3000	3000	3000
Rated Continous Torque @ 25°C	T_{25}	oz-in	69	57	63	51	143	126	131	115
		N-cm	49	40	44	36	101	89	93	81
Rated Continuous Torque @ 40°C	T_{40}	oz-in	63	52	57	46	131	112	117	103
		N-cm	44	37	40	32	93	79	83	73
Rated Power Output	P	Watts	153	126	140	114	316	278	291	256
Maximum Recommended Speed	Nmax	RPM	6000	6000	6000	6000	6000	6000	6000	6000
Continous Stall Torque	T_s	oz-in	69	62	62	56	147	128	136	117
		N-cm	49	44	44	40	104	90	96	83
Cogging Torque	T_c	oz-in	0	0	0	0	0	0	0	0
Electrical Specifications										
Rated Terminal Voltage	E	Volts	30.0	28.0	16.0	14.0	51.0	45.0	26.0	23.0
Rated Continuous Current	I	Amps	7.80	7.10	14.00	12.90	8.00	8.10	14.80	15.00
Peak Current	I_p	Amps	79	77	151	147	83	83	159	159
Continuous Stall Current	I_s	Amps	7.5	7.3	13.7	13.3	8.0	8.0	14.7	14.7
Winding Specifications										
Terminal Resistance ± 10%	R_t	Ohms	0.850	0.850	0.370	0.370	0.750	0.750	0.310	0.310
Armature Resistance ± 10%	R_a	Ohms	0.660	0.660	0.180	0.180	0.610	0.610	0.170	0.170
Back EMF Constant ± 10%	K_e	V/KRPM	7.60	7.10	3.80	3.60	15.10	13.30	7.60	6.60
Torque Constant ± 10%	K_t	oz-in/Amp	10.30	9.60	5.10	4.80	20.40	17.80	10.20	8.90
		N-cm/Amp	7.27	6.78	3.60	3.39	14.41	12.57	7.20	6.28
Viscous Damping Constant	K_d	oz-in/KRPM	1.1	1.1	1.1	1.1	2.8	2.3	2.7	2.2
		N-cm/KRPM	0.8	0.8	0.8	0.8	2.0	1.6	1.9	1.5
Armature Inductance	L	µH	<0.03	<0.03	<0.03	<0.03	<0.05	<0.05	<0.05	<0.05
Temperature Coefficient of KE	C	%/°C Rise	-0.10	-0.10	-0.10	-0.10	-0.10	-0.10	-0.10	-0.10
Number of Commutator Bars	Z		117	117	117	117	141	141	141	141
Mechanical Specifications										
Moment of Inertia 0.00004	J_m	oz-in-sec²	0.0056	0.0083	0.0056	0.0083	0.0190	0.0260	0.0190	0.0260
		kg-cm²	0.40	0.59	0.40	0.59	1.34	1.84	1.34	1.84
Static Friction Torque	T_f	oz-in	4.0	4.5	4.0	4.5	5.5	5.5	5.5	5.5
		N-cm	2.8	3.2	2.8	3.2	3.9	3.9	3.9	3.9
Weight	W	lbs	3.1	3.2	3.1	3.2	5.3	5.3	5.3	5.3
		kg	1.4	1.5	1.4	1.5	2.4	2.4	2.4	2.4
Diameter	D	in	4.37	4.37	4.37	4.37	5.50	5.50	5.50	5.50
		mm	111.0	111.0	111.0	111.0	139.7	139.7	139.7	139.7
Length	LG	in	0.94	0.95	0.94	0.95	1.07	1.10	1.07	1.10
		mm	23.9	24.1	23.9	24.1	27.2	27.9	27.2	27.9
Figure of Merit										
Peak Acceleration	A_p	kRad/s²	135.7	83.3	130.1	79.9	84.1	53.3	80.1	50.8
Mechanical Time Constant	T_m	ms	4.90	8.30	5.20	8.80	3.90	7.10	4.20	7.70
Electrical Time Constant	T_e	ms	<0.05	<0.05	<0.17	<0.17	<0.07	<0.07	<0.27	<0.27
Continuous Power Rate	P_c	kW/sec	6.0	2.8	5.0	2.2	7.6	4.3	6.4	3.6
Thermal Specifications										
Thermal Resistance at Rated Speed	RAAR	°C/Watt	1.50	1.70	1.50	1.70	1.40	1.40	1.40	1.40
Thermal Resistance at Stall	RAAS	°C/Watt	2.00	2.10	2.00	2.10	1.90	1.90	1.90	1.90
Tachometer Specifications										
Output Voltage	V	Volts/KRPM	—	3.50	—	3.50	—	5.90	—	5.90
Maximum Ripple Peak to Peak	V_{rh}	%	—	3.0	—	3.0	—	3.0	—	3.0
Linearity of Output Voltage	LIN	%	—	0.06	—	0.11	—	0.11	—	0.11
Minimum Load Resistance	R_l	Ohms	—	370	—	370	—	494	—	494

Notes:

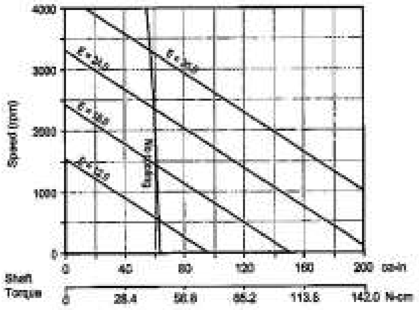
1. All values are based upon a 150°C armature temperature limit and with the motor mounted on an 8" x 16" x 3/8" aluminum heatsink with no forced air cooling. Other voltages, speeds, and torques, and duty cycles are achievable as long as the max armature temperature of 150°C is not exceeded.
2. Mass air flow (lbs/min) = air volume (CFM) x air density (lbs/ft³).
3. Terminal resistance is measured at 4.0 amps. RT varies as a function of applied current.
4. Unless otherwise noted, all specifications above apply at 25°C.
5. Peak torque and current is calculated based on max pulse duration of 50 milliseconds and a 1% duty cycle.
6. The operating voltage can be calculated as: I = (Shaft torque + TF + KD x N/1000) / KT.
7. The operating voltage can be calculated as: V = KE x (N/1000) + RT x I.
8. Tachometer ripple measured with a resistive load of 1 kohm and a single low pass filter with 3db cut off at 500 Hz.
9. Bidirectional tolerance of tachometer will not exceed 3%.

PERFORMANCE DATA

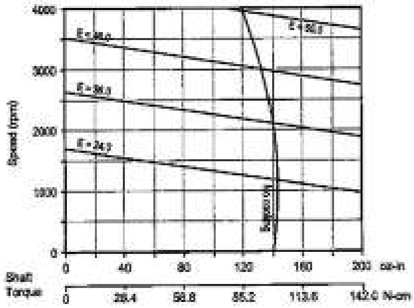
N9M4



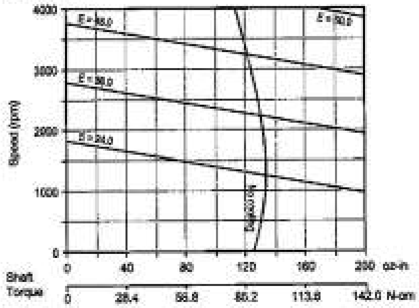
N9M4T



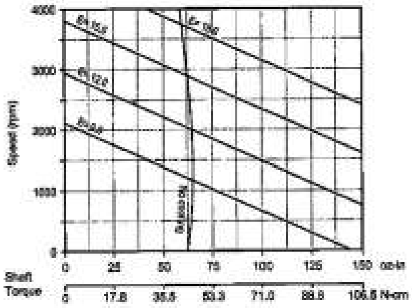
N12M4



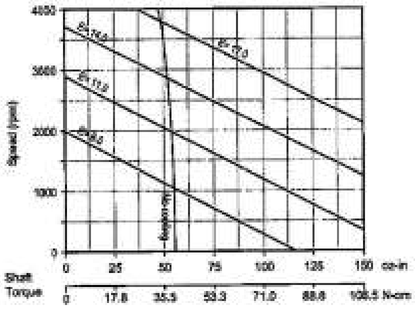
N12M4T



N9M4LR



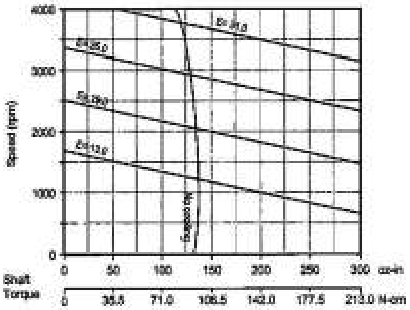
N9M4LRT



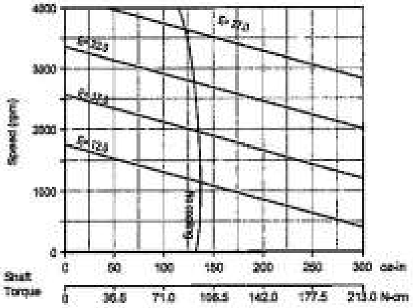
N-Series ServoDisc

DIMENSIONS

N12M4LR



N12M4LRT



Notes:

A. All curves are drawn for a fixed armature temperature of 150°C.

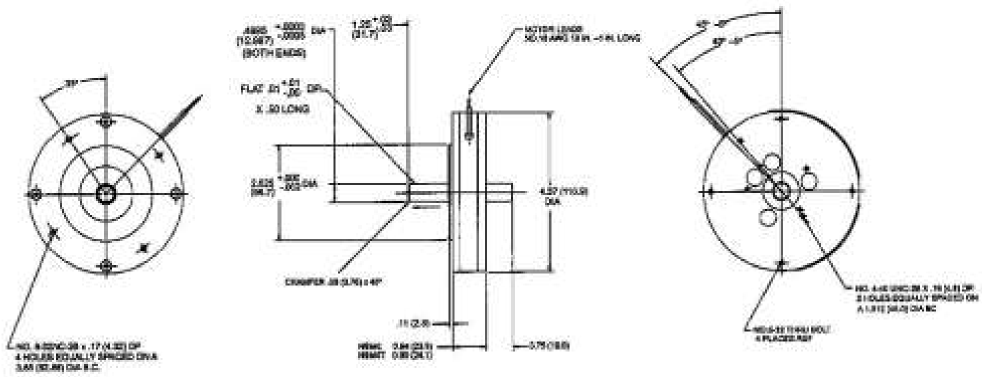
B. The motor can be operated at any point on the graph below 4000 RPM. Higher speeds are possible for some applications. Contact a Kollmorgen Sales Office for more details.

C. Determine voltage required for a desired combination of speed and torque by estimating it as a line parallel to one of the constant terminal voltage (E) lines.

D. The operating current can be calculated as:
I = (Shaft torque + TF + KD x N/1000)/KT.

E. The operating voltage can be calculated as:
V = KE x N/1000 + RT x I.

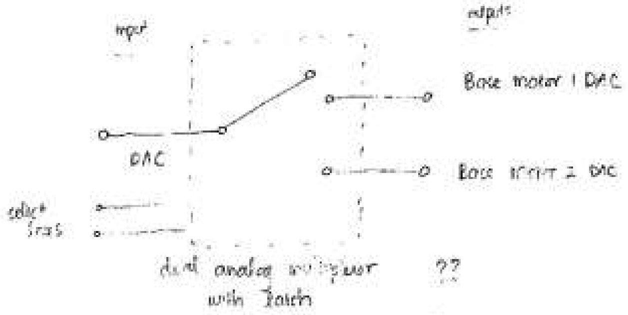
N9M4/N9M4T



AD : x 6 for LOWET

SEA 3-DOF PROJECT

Servotogo Breakout Board Layout



Amplifier connector

- How to check amplifier's setting

Brushless motors

2 DAC channels

2 x IDC16 ⟼ 1 x DB25

UL
IDC16.10 MF(-) u
IDC16.11 MF(+) u
IDC16.12 GND

υ
IDC16.10 MF(-) υ
IDC16.11 MF(+) υ
IDC16.12

DB25.2 u +
DB25.3 υ +
DB25.14 u −
DB25.15 υ −
DB25.12 GND

IDC16.6 HVBUS → DB25.5

multiplex

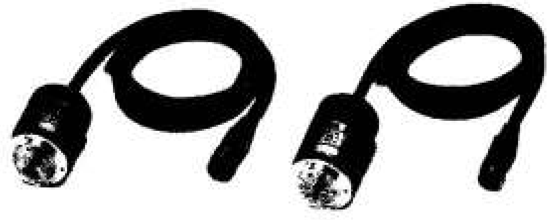
1 DAC channels ⟼ 2 output

BUT !!

- I need to be able to latch the output to whatever last value it has when it's deselected
- Need to be able to do ±10V analog

Canon Laser Rotary Encoder R-1O, R-1L

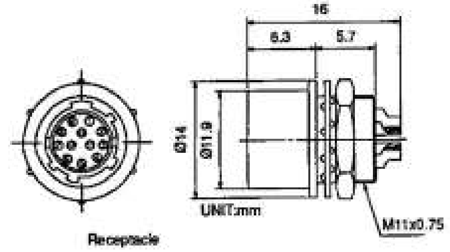
Thank you for purchasing the Canon Laser Rotary Encoder R-1. The R-1 is an ultra-compact, high precision rotary encoder using a semiconductor laser. The R-1 comes in two types: Type O, an open collector encoder, and type L, a line drive encoder, which output 2-phase square wave incremental signals at a rate of 81,000 pulses per rotation. In addition a Z-phase square wave outputs one pulse per rotation.
Please read the following instructions carefully to take advantage of this fine product to the fullest.



PRECAUTIONS

1. Use the encoder in the specified working environment.
2. The power supply voltage is ±5V DC, ±5%.
3. Never push or pull the cable strongly.
4. Be sure to use the protective cover when operating the R-1 in a dusty, greasy or damp environment.
5. When connecting the connector, confirm the function of each pin and connect it correctly.
6. Because this unit uses a semiconductor laser, never remove the unit's cover. Doing so may be hazardous.
7. Connect the R-1 with a flexible coupling.

CONNECTOR PIN NO. FUNCTION



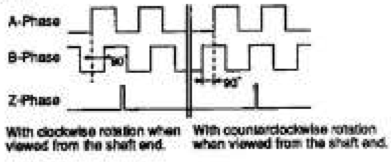
NO.	FUNCTION		NO.	FUNCTION	
	R-1O	R1L		R-1O	R-1L
1	A-phase	A-phase	7	+5V	+5V
2	GND	$\overline{A}$	8	GND	GND
3	B-phase	B-phase	9	-5V	-5V
4	GND	$\overline{B}$	10	GND	GND
5	Z-phase	Z-phase	11	NC	
6	GND	$\overline{Z}$	12	Case/Shield (cable)	

Plug : HR10A-10P-12P
Receptacle : HR10A-10R-12S (Manufactured by Hirose)
Cable Length : 1m

Note:
The cable shield wire is connected to the plug shell section and pin no. 12, which are connected to the encoder. Connect the shell section and pin no.12 to the circuit GND (Quiet GND with a small noise current). Be sure to avoid flow of pulse-like noise current in this line, as it may lead to deterioration of the semiconductor laser and cause errors from noise contamination of the signal.

OUTPUT WAVE FORM

R-1O,R-1L



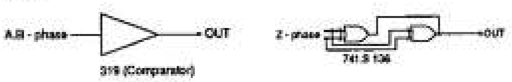
With clockwise rotation when viewed from the shaft end.

With counterclockwise rotation when viewed from the shaft end.

Note:
Z-phase is not synchronized with A-phase or B-phase.

OUTPUT CIRCUIT

R-1O

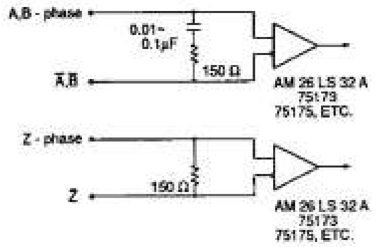


R-1L

RECOMMENDED RECEIVING CIRCUIT

R-1O

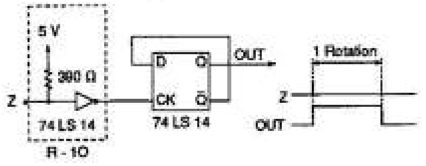
R-1L



ADDITONAL INFORMATION ABOUT THE Z-PHASE

1. The Z-phase output is one pulse per rotation. As the pulse width is narrow, the waveform cannot be viewed easily using a standard oscilloscope. Therefore, check by using equipment such as a high intensity oscilloscope or a storage oscilloscope.

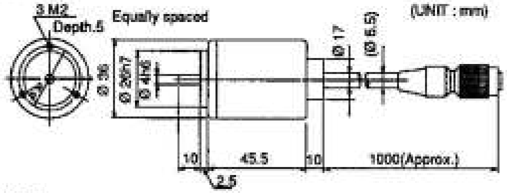
2. Another method for confirmation of the Z-phase output is to use a "D flip-flop." By using the D flip-flop, two pulses are combined to make the pulse width wider and therefore easier to recognize.



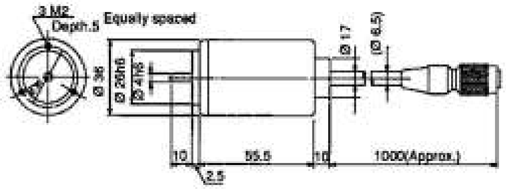
9. The pulse width of the Z-phase is 100 ~ 250 nsec. In case the receiving IC needs a wider pulse width increase the width using a one-shot multi-vibrator (74 LS221 etc.).

OUTLINE DIMENSIONS

R-1O



R-1L



SPECIFICATIONS

1. Unit pulse no./rotation:
   81,000 (without electric division)
2. Angel/pulse:
   16arc-sec. (Without electric division)
3. Max. response: 500 kHz (6rps)
4. Output signal:
   81,000 pulses per rotation; 2-phase square wave incremental signals, plus a Z-phase square wave signal.
   a. A-phase and B-phase:
      R-1O: Open collector output
         I_{OL} = 8mA Max. ($V_{OL} \leqq 0.4$)
         Recommended load resistance: 680Ω (5V)
         Pull-up voltage: 15V Max.
      R-1L: Line driver output (Balanced)
         Load current: ±20mA Max.
      Difference between A-phase and B-phase: 90° ±10°
      (Both R-1O and R-1L)
   b. Z-phase: 1 pulse/rotation:
      R-1O: Open collector output
         I_{OL} = 16mA Max. ($V_{OL} \leqq 0.4V$)
         Recommended load resistance: 390Ω (5V)
         Pull-up voltage: 5.25V Max.
         Pulse width: 100 ~ 250 nsec.
      R1L: Line driver output (Balanced)
         Load current: ±20mA Max.
         Pulse width: 100 ~ 250 nsec.
5. Outer diameter: 36 mm
6. Overall length: R-1O: 48 mm; R-1L: 58mm
7. Weight: R-1O: 80 g; R-1L: 95 g
8. Max permissible rotating speed: 5,000 rpm
9. Starting torque: 9 g-cm Max
10. Rotar's intertial moment (GD^2): 8g-cm^2
11. Power supply (voltage): ±5V DC, ±5%
12. Power supply (current):
    R-1O: +5V ... 200mA Max. (Recommended load)
          -5V ... 100mA Max.;
    R-1L: +5V ... 250mA Max. (No load)
          -5V ... 100mA.Max.
13. Light source:
    Semiconductor laser
    Wave length: 780nm,
    Max.output: 5mW
14. Working environment:
    Operating temperature: 0°C – 50°C
    Storage temperature: -30°C – 80°C
    Humidity: 90% RH Max. (No condensation)
    Vibration: 10G 500 Hz or less
    Impact: 30G 11msec. or less
15. Max. permissible shaft load:
    Radial 0.4 kg
    Thrust 1.0 kg

PRECAUTIONS FOR INCORPORATING THE R-1O, R-1L INTO EQUIPMENT

When connecting the R-1O, R-1L to the drive shaft of equipment, shaft misalignment, shaft deflection, or thrust fluctuation of the drive shaft causes excess force on the bearing causing reduced accuracy and serviceability of the encoder, and damage to the encoder.
Make sure the work load for each model is within the specified permissible load when connecting the encoder to equipment.

Permissible load

Radial	0.4kg
Thrust	1.0kg

When the encoder is rigidly connected to equipment, the shaft misalignment should be 2μm or less and the thrust fluctuation should be 1μm or less.
When precise shaft alignment is impossible, it is recommended to absorb the shaft misalignment, shaft deflection, or thrust fluctuation by using a flexible coupling.
A flexible coupling with enough torsional rigidity should be selected to obtain precise transmission. When attaching the flexible coupling to the encoder shaft, be very careful not to apply any unnecessary load to the encoder shaft.
The flexible coupling can be attached to the static encoder shaft even if the load caused by the deflection and inclination of the drive shaft exceeds the maximum setting. In this case, however, trouble may occur because undesired stress can be imposed on the encoder shaft when it rotates.

For further details concerning the R-1O, R-1L, please direct inquiries to the following in writing or by telephone.

Canon
CANON U.S.A., INC.
Components Division
One Canon Plaza, Lake Success, N.Y. 11042-9979 U.S.A.
Telephone 516-488-6700 Facsimile 516-354-1114
Telex96-1333
CANON EUROPA N. V.
Industrial Products Division
Unit 3, Brent Trading Centre, North Circular Road, Neasden.
London NW10 OJF, United Kingdom
Telephone (81) 451-4511 Facsimile (81) 459-0331
Telex 295776
CANON AUSTRALIA PTY LTD
Optical Products Division
I Thomas Holt Drive, North Ryde NSW 2113
Telephone (02) 887-0166 Facsimile (02) 887-4484
Telex AA23762
CANON INC.
7-1, Nishi-Shinjuku 2-Chome, Shinjuku-ku, Tokyo 163, Japan.
Mailing address: P.O. Box 5050. Dai-ichi Seimei Building, Tokyo 163, Japan.

Copley
Controls
Corp.

MODEL 7225X1

LINE-ISOLATED AC BRUSHLESS SERVO AMPLIFIER WITH +/-10V ANALOG U-V INPUTS

FEATURES

- **Compatible with controllers that output +/-10V analog torque commands for U & V phases**
- **Reduced offset drift**
- ***FAULT PROTECTIONS***
  Short-circuits
  - **output to output**
  - **output to HV (+)**
  - **output to HV (-)**

  Over / under voltage
  Over temperature
  Self-reset or latch-off
- ***No Transformer Required!***
  Operates from power supplies that rectify the line directly with full optical isolation between signal and power stages.
- **CURRENT LIMITING**
  User selectable, I^2T Limit with, indicator signal for control system
- ***Greater than 3 kHz Bandwidth***

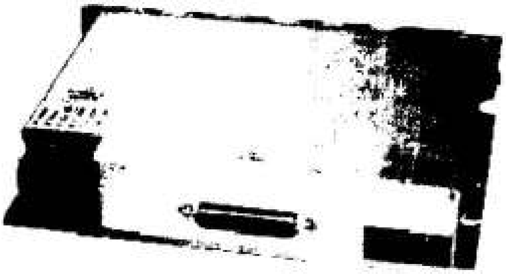
WORKS WITH POPULAR CONTROLLERS

- ***Technology 80*** **5651A**
- ***PMD*** **MC1231A Chipset**
- ***Delta Tau*** **PMAC**
- ***MEI*** **DPS Series**
- ***Galil*** **DMC-1700**

THE OEM ADVANTAGE

- **Internal solderless header configures amplifier for plug and play operation**

MODEL	POWER	I-CONT (A)	I-PEAK (A)
7225X1	24~180VDC	10	25



FEATURES

The 7xx5X1 model is a PWM servo-amplifier for AC Brushless servomotors that are commutated externally by digital control systems that output two +/-10V signals that represent the current command to the motor U and V windings. The amplifier synthesizes the current command for the W winding.

Control cards take feedback from an encoder on the motor and use various techniques to determine the rotor position. When this has been done, the controller is able to output two signals that correspond to the current in the U and V windings to produce torque in the motor. The amplifier synthesizes the W winding current from UV signals that are 120 electrical degrees apart.

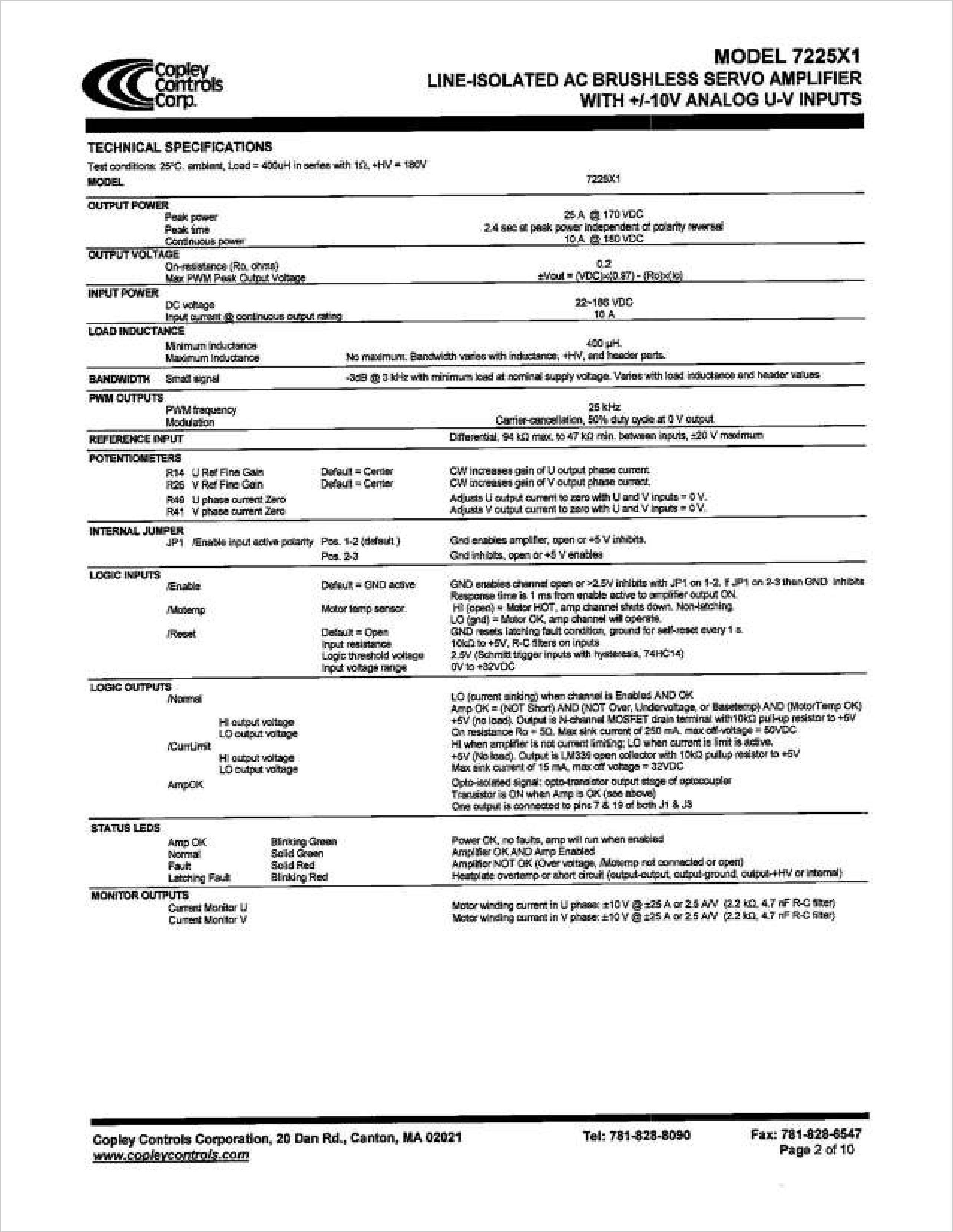
Amplifier adjustments with this system consist of inductance compensation, current limit, transconductance, and offset. Thereafter, the controller does all of the velocity and/or position control of the motor.

Internal solderless sockets let the user configure the various gain and current limit settings to customize the amplifiers for a wide range of loads and applications. Header components permit compensation over a wide range of load inductance's to maximize bandwidth with different motors.

The /Enable input active logic-level is jumper-selectable to ground or +5V to interface with all types of control cards.

MOSFET output stage deliver four quadrant power for bi-directional acceleration and deceleration of motors.

All models are protected against output short circuits (output to output, output to ground, output to +HV) and heatplate overtemperature. With the /Reset input open the amplifier will latch off until powered-down or the /Reset input is toggled. The amplifier will reset itself automatically from faults if the /Reset input is wired to GND.



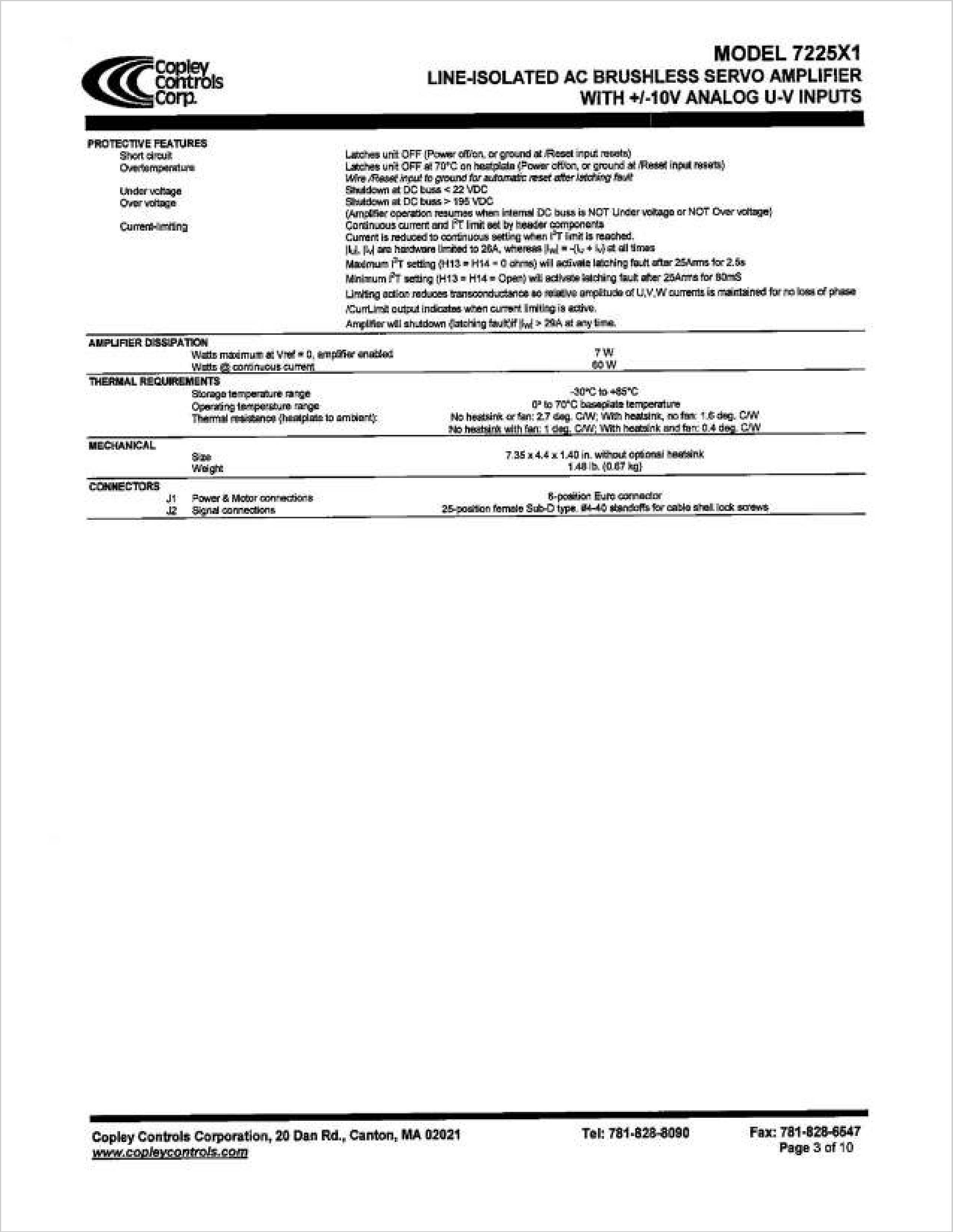
Copley Controls Corp.

MODEL 7225X1
LINE-ISOLATED AC BRUSHLESS SERVO AMPLIFIER WITH +/-10V ANALOG U-V INPUTS

TECHNICAL SPECIFICATIONS

Test conditions: 25°C. ambient, Load = 400uH in series with 1Ω, +HV = 180V

MODEL			7225X1
OUTPUT POWER	Peak power		25 A @ 170 VDC
	Peak time		2.4 sec at peak power independent of polarity reversal
	Continuous power		10 A @ 180 VDC
OUTPUT VOLTAGE	On-resistance (Ro, ohms)		0.2
	Max PWM Peak Output Voltage		±Vout = (VDC)x(0.97) - (Ro)x(Io)
INPUT POWER	DC voltage		22~186 VDC
	Input current @ continuous output rating		10 A
LOAD INDUCTANCE	Minimum inductance		400 µH.
	Maximum inductance		No maximum. Bandwidth varies with inductance, +HV, and header parts.
BANDWIDTH	Small signal		-3dB @ 3 kHz with minimum load at nominal supply voltage. Varies with load inductance and header values
PWM OUTPUTS	PWM frequency		25 kHz
	Modulation		Carrier-cancellation, 50% duty cycle at 0 V output
REFERENCE INPUT			Differential, 94 kΩ max. to 47 kΩ min. between inputs, ±20 V maximum
POTENTIOMETERS	R14 U Ref Fine Gain	Default = Center	CW increases gain of U output phase current.
	R26 V Ref Fine Gain	Default = Center	CW increases gain of V output phase current.
	R40 U phase current Zero		Adjusts U output current to zero with U and V inputs = 0 V.
	R41 V phase current Zero		Adjusts V output current to zero with U and V inputs = 0 V.
INTERNAL JUMPER	JP1 /Enable input active polarity	Pos. 1-2 (default)	Gnd enables amplifier, open or +5 V inhibits.
		Pos. 2-3	Gnd inhibits, open or +5 V enables
LOGIC INPUTS	/Enable	Default = GND active	GND enables channel open or >2.5V inhibits with JP1 on 1-2. If JP1 on 2-3 then GND inhibits Response time is 1 ms from enable active to amplifier output ON.
	/Motemp	Motor temp sensor.	HI (open) = Motor HOT, amp channel shuts down. Non-latching. LO (gnd) = Motor OK, amp channel will operate.
	/Reset	Default = Open	GND resets latching fault condition, ground for self-reset every 1 s.
		Input resistance	10kΩ to +5V, R-C filters on inputs
		Logic threshold voltage	2.5V (Schmitt trigger inputs with hysteresis, 74HC14)
		Input voltage range	0V to +32VDC
LOGIC OUTPUTS	/Normal		LO (current sinking) when channel is Enabled AND OK Amp OK = (NOT Short) AND (NOT Over, Undervoltage, or Basetemp) AND (MotorTemp OK)
		HI output voltage	+5V (no load). Output is N-channel MOSFET drain terminal with10kΩ pull-up resistor to +5V
		LO output voltage	On resistance Ro = 5Ω. Max sink current of 250 mA. max off-voltage = 50VDC
	/CurrLimit		HI when amplifier is not current limiting; LO when current is limit is active.
		HI output voltage	+5V (No load). Output is LM339 open collector with 10kΩ pullup resistor to +5V
		LO output voltage	Max sink current of 15 mA, max off voltage = 32VDC
	AmpOK		Opto-isolated signal: opto-transistor output stage of optocoupler Transistor is ON when Amp is OK (see above) One output is connected to pins 7 & 19 of both J1 & J3
STATUS LEDS	Amp OK	Blinking Green	Power OK, no faults, amp will run when enabled
	Normal	Solid Green	Amplifier OK AND Amp Enabled
	Fault	Solid Red	Amplifier NOT OK (Over voltage, /Motemp not connected or open)
	Latching Fault	Blinking Red	Heatplate overtemp or short circuit (output-output, output-ground, output-+HV or internal)
MONITOR OUTPUTS	Current Monitor U		Motor winding current in U phase: ±10 V @ ±25 A or 2.5 A/V (2.2 kΩ, 4.7 nF R-C filter)
	Current Monitor V		Motor winding current in V phase: ±10 V @ ±25 A or 2.5 A/V (2.2 kΩ, 4.7 nF R-C filter)



Copley Controls Corp.

MODEL 7225X1

LINE-ISOLATED AC BRUSHLESS SERVO AMPLIFIER WITH +/-10V ANALOG U-V INPUTS

PROTECTIVE FEATURES		
	Short circuit	Latches unit OFF (Power off/on, or ground at /Reset input resets)
	Overtemperature	Latches unit OFF at 70°C on heatplate (Power off/on, or ground at /Reset input resets) *Wire /Reset input to ground for automatic reset after latching fault*
	Under voltage	Shutdown at DC buss < 22 VDC
	Over voltage	Shutdown at DC buss > 195 VDC (Amplifier operation resumes when internal DC buss is NOT Under voltage or NOT Over voltage)
	Current-limiting	Continuous current and I^2T limit set by header components Current is reduced to continuous setting when I^2T limit is reached. $\lvert I_U \rvert$, $\lvert I_V \rvert$ are hardware limited to 26A, whereas $\lvert I_W \rvert = -(I_U + I_V)$ at all times Maximum I^2T setting (H13 = H14 = 0 ohms) will activate latching fault after 25Arms for 2.5s Minimum I^2T setting (H13 = H14 = Open) will activate latching fault after 25Arms for 80mS Limiting action reduces transconductance so relative amplitude of U,V,W currents is maintained for no loss of phase /CurrLimit output indicates when current limiting is active. Amplifier will shutdown (latching fault)if $\lvert I_W \rvert > 29A$ at any time.
AMPLIFIER DISSIPATION		
	Watts maximum at Vref = 0, amplifier enabled	7 W
	Watts @ continuous current	60 W
THERMAL REQUIREMENTS		
	Storage temperature range	-30°C to +85°C
	Operating temperature range	0° to 70°C baseplate temperature
	Thermal resistance (heatplate to ambient):	No heatsink or fan: 2.7 deg. C/W; With heatsink, no fan: 1.6 deg. C/W No heatsink with fan: 1 deg. C/W; With heatsink and fan: 0.4 deg. C/W
MECHANICAL		
	Size	7.35 x 4.4 x 1.40 in. without optional heatsink
	Weight	1.48 lb. (0.67 kg)
CONNECTORS		
J1	Power & Motor connections	6-position Euro connector
J2	Signal connections	25-position female Sub-D type. #4-40 standoffs for cable shell lock screws

Copley Controls Corporation, 20 Dan Rd., Canton, MA 02021 Tel: 781-828-8090 Fax: 781-828-6547
www.copleycontrols.com
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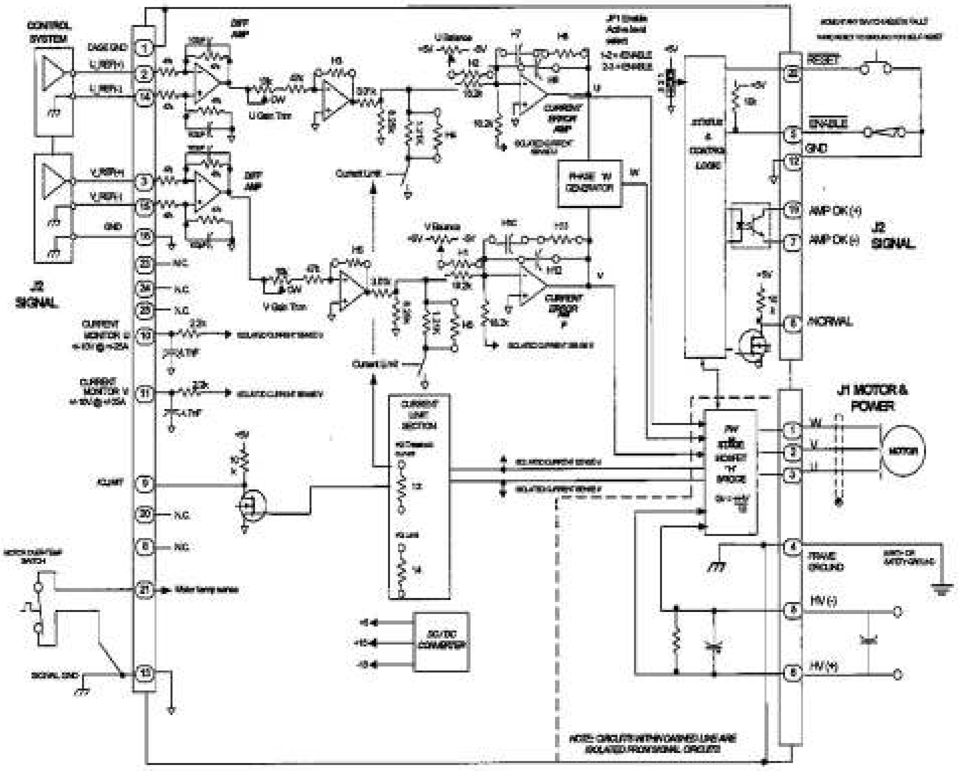
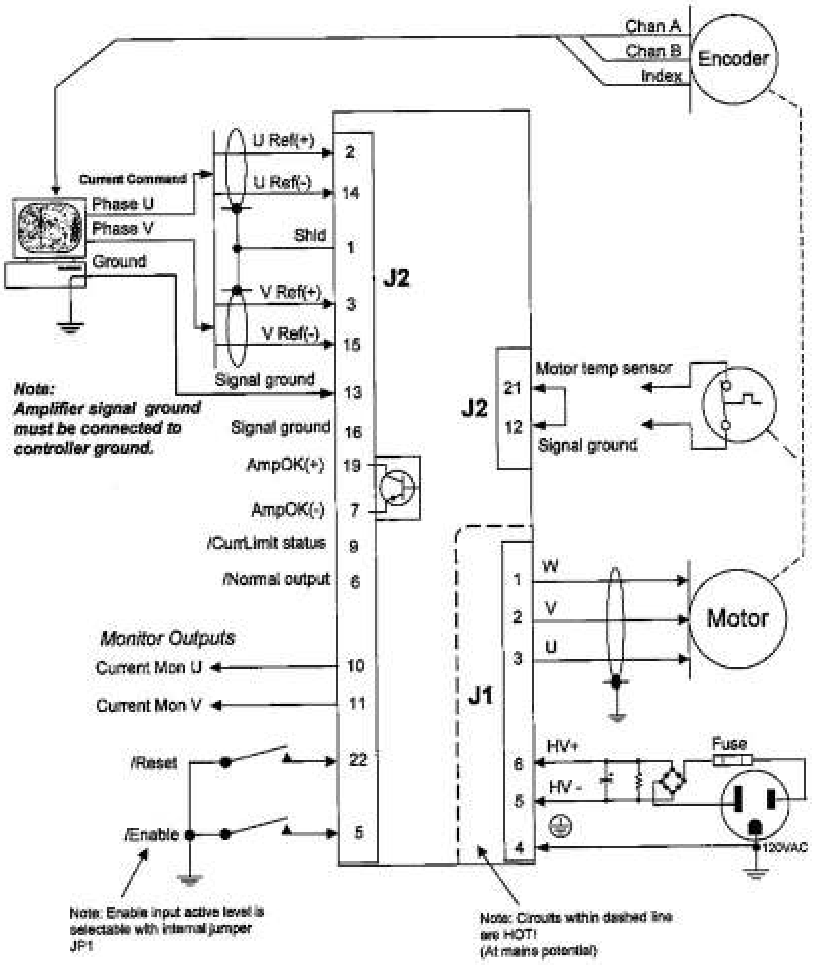
FUNCTIONAL DIAGRAM



TYPICAL AMPLIFIER CONNECTIONS

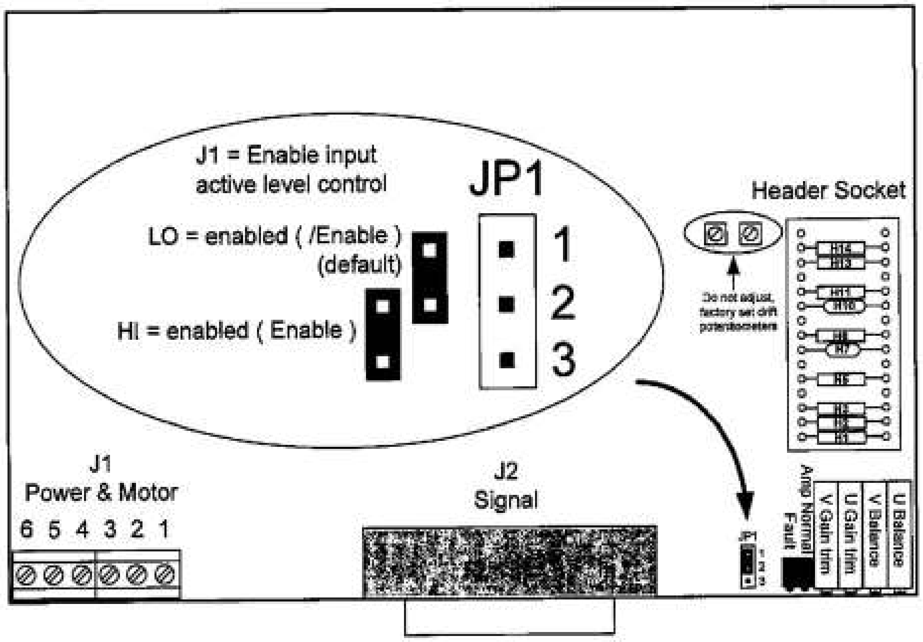


Note:
Amplifier signal ground must be connected to controller ground.

Note: Enable input active level is selectable with internal jumper JP1

Note: Circuits within dashed line are HOT! (At mains potential)

PC BOARD LAYOUT



HEADER SOCKET COMPONENTS

Part	Value	Remarks
H15	N/a	No function
H14	86.6kΩ	I^2T Current Limit select
H13	0Ω<short>	I^2T Threshold Current select
H12	<out>	Ch. V Current Error Amp hi-frequency roll off
H11	30.1kΩ	Ch. V Current Error Amp proportional gain
H10	100nF	Ch. V Current Error Amp integrator
H9	<out>	Ch. U Current Error Amp hi-frequency roll off
H8	30.1kΩ	Ch. U Current Error Amp proportional gain
H7	100nF	Ch. V Current Error Amp integrator
H6	<out>	Ch. V Continuous Current Limit
H5	75kΩ	Ch. V Transconductance
H4	<out>	Ch. U Continuous Current Limit
H3	75kΩ	Ch. U Transconductance
H2	1.5MΩ	Ch. U Balance Range select
H1	1.5MΩ	Ch. V Balance Range select

HEADER SOCKET COMPONENT SELECTION

LOAD INDUCTANCE

L (mH)	H8, H11 @ 80V	H8, H11 @ 160V	H7, H10
0.4	16.5k	11k	33nF
1	***32.4k***	***18.2k***	***33nF***
3	86.6k	42.4k	33nF
10	249k	124k	33nF
30	750k	392k	33nF

Note: Table values apply with components H9 & H12 not installed. Values in ***bold and italic*** are factory installed.

CURRENT LIMITS

A micro controller uses an I^2T algorithm to monitor to protect against overload conditions. The I^2T overload protection for each channel operates independent of the other. The algorithm detects when the current is any phase exceeds the continuous current limit level set by the header component H13. The I^2T algorithm tracks the energy of the overload (A^2 sec) and when the I^2T limit is reached, the output current is limited to a level set by H4 and H6. The following tables or equations can be used to select header component values to obtain the desired over-current protection setting.

Cont. Current (A)	H4 & H6 (Ohm_)	H13 (Ohm)
10	***<out>***	***0 Ohms (short)***
8	2.5k	16k
6	825	49k
4	383	150k
2	150	<out>

330 Ω

220 k

I^2T Limit (A^2sec)	H14 (Ohm)
1250	***0 (short)***
800	16k
450	49k
200	150k
50	<out>

$$\frac{I^2 t}{10^2 - 3.5^2} = T$$

T = 2 sec

I^2T = 2 180

10 - 3.5

$$H13 = 47.5k\text{ ohms} * \frac{(10 - I_{cont})}{(I_{cont} - 2)}$$

3.5 - 2

$$H14 = 47.5k\text{ ohms} * \frac{\left(6.25 - \sqrt{\frac{I^2T_{limit}}{32}}\right)}{\left(\sqrt{\frac{I^2T_{limit}}{32}} - 1.25\right)} = 180k$$

Example: The I^2T set point applies only to the energy delivered to the load over and above the continuous rating of the load. The amplifier's microchip is informed of the continuous current rating of the load via header resistor H13. The I^2T set point is set via header resistor H14. Using a 0 Ohm value for H14 gives an I^2T set point of 1250 A^2*S. If a 0 ohm value is also used for H13, the continuous current setting is set to 10A. This means for a 25 Arms current on either phase U,V, or W, the I^2*T protection will activate (current is forced to continuous limit as set by H4,H6 after a time T = 1250 A^2*S/(25^2-10^2) = 2.4 seconds.

BALANCE RANGE AND TRANSCONDUCTANCE SETTINGS

Header components H1 & H2 control the offset range. Default value is 1.5Mohm that gives a range of +/-350mA. The ratio between output current, and the reference voltage at the input is the *transconductance* of the amplifier. It is measured in Amps/Volt, and is controlled by components H3 & H5. The chart below gives some common settings.

± 4000 counts / 10 V

Gain (A/V)	H3 & H5
2.5	102k
2.0	***75.0k***
1.5	59k
1	29.4k
0.5	14.3k

$i_m \cdot K_T \cdot M = \tau_i$

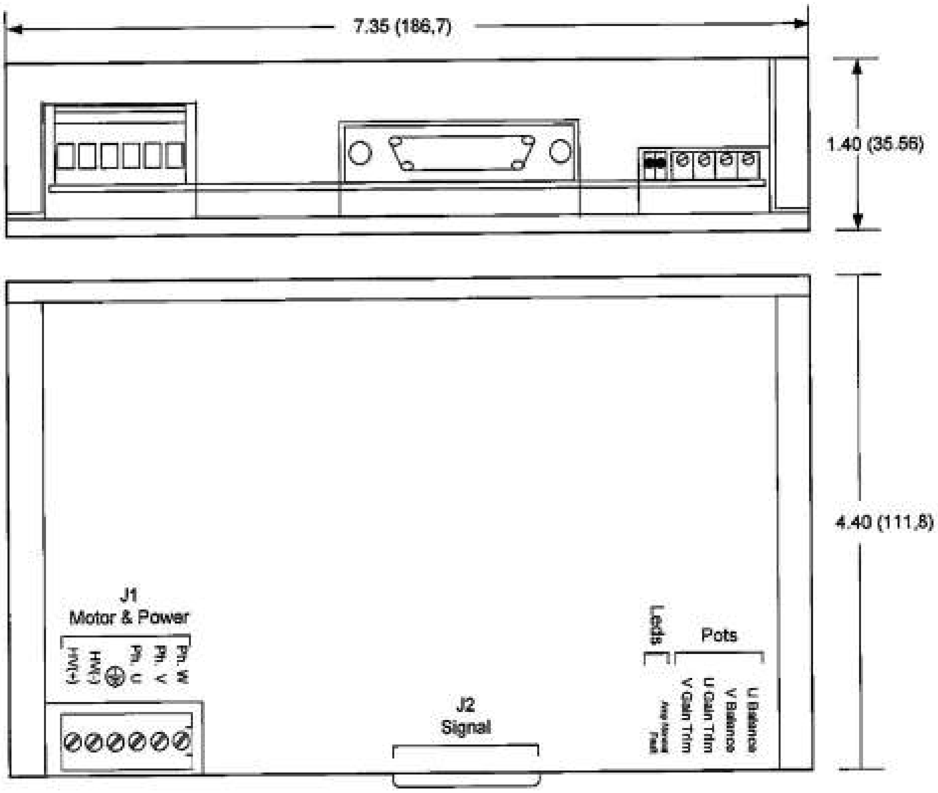
$V_{DAC} = \frac{count}{4092} \cdot 10$

$\tau_m = \frac{count}{4092}(10)(K_{amp})(K_T)$

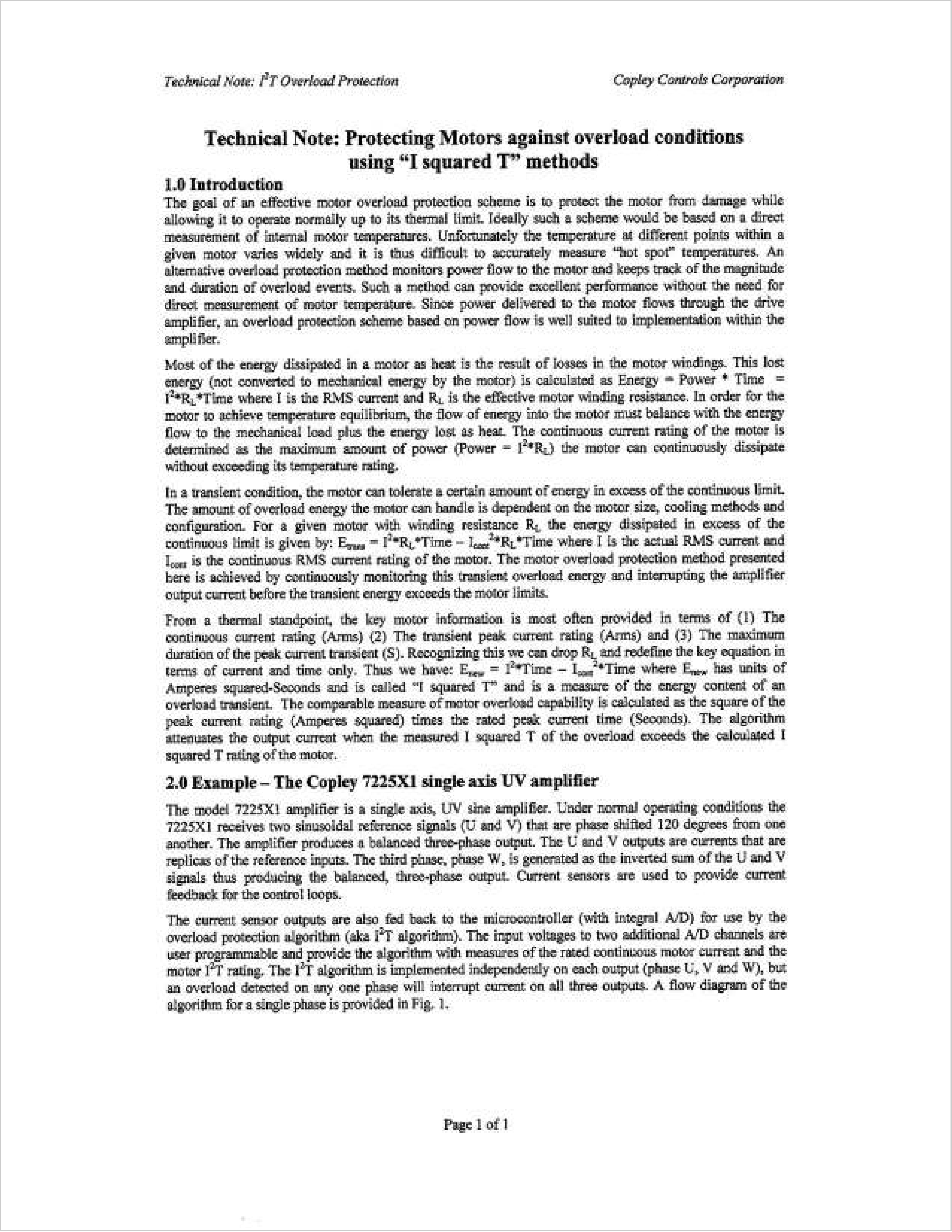
$i_m = V_{DAC} \cdot K_{amp}$

DIMENSIONS

Note: Dimensions in inches (mm.)



NOTES



Technical Note: Protecting Motors against overload conditions using "I squared T" methods

1.0 Introduction

The goal of an effective motor overload protection scheme is to protect the motor from damage while allowing it to operate normally up to its thermal limit. Ideally such a scheme would be based on a direct measurement of internal motor temperatures. Unfortunately the temperature at different points within a given motor varies widely and it is thus difficult to accurately measure "hot spot" temperatures. An alternative overload protection method monitors power flow to the motor and keeps track of the magnitude and duration of overload events. Such a method can provide excellent performance without the need for direct measurement of motor temperature. Since power delivered to the motor flows through the drive amplifier, an overload protection scheme based on power flow is well suited to implementation within the amplifier.

Most of the energy dissipated in a motor as heat is the result of losses in the motor windings. This lost energy (not converted to mechanical energy by the motor) is calculated as Energy = Power * Time = I^2*R_L*Time where I is the RMS current and R_L is the effective motor winding resistance. In order for the motor to achieve temperature equilibrium, the flow of energy into the motor must balance with the energy flow to the mechanical load plus the energy lost as heat. The continuous current rating of the motor is determined as the maximum amount of power (Power = I^2*R_L) the motor can continuously dissipate without exceeding its temperature rating.

In a transient condition, the motor can tolerate a certain amount of energy in excess of the continuous limit. The amount of overload energy the motor can handle is dependent on the motor size, cooling methods and configuration. For a given motor with winding resistance R_L the energy dissipated in excess of the continuous limit is given by: $E_{trans} = I^2*R_L$*Time – $I_{cont}^2*R_L$*Time where I is the actual RMS current and I_{cont} is the continuous RMS current rating of the motor. The motor overload protection method presented here is achieved by continuously monitoring this transient overload energy and interrupting the amplifier output current before the transient energy exceeds the motor limits.

From a thermal standpoint, the key motor information is most often provided in terms of (1) The continuous current rating (Arms) (2) The transient peak current rating (Arms) and (3) The maximum duration of the peak current transient (S). Recognizing this we can drop R_L and redefine the key equation in terms of current and time only. Thus we have: $E_{new} = I^2$*Time – I_{cont}^2*Time where E_{new} has units of Amperes squared-Seconds and is called "I squared T" and is a measure of the energy content of an overload transient. The comparable measure of motor overload capability is calculated as the square of the peak current rating (Amperes squared) times the rated peak current time (Seconds). The algorithm attenuates the output current when the measured I squared T of the overload exceeds the calculated I squared T rating of the motor.

2.0 Example – The Copley 7225X1 single axis UV amplifier

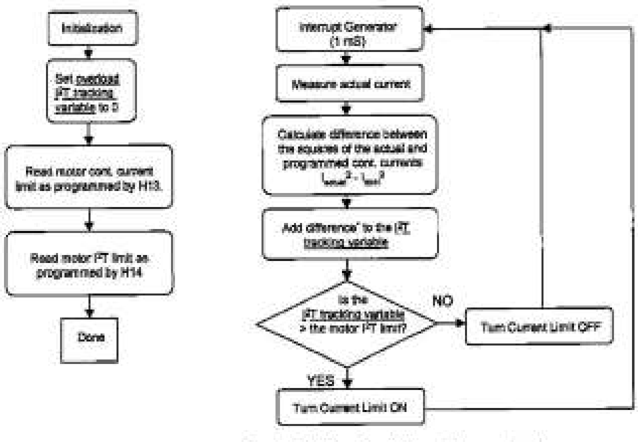
The model 7225X1 amplifier is a single axis, UV sine amplifier. Under normal operating conditions the 7225X1 receives two sinusoidal reference signals (U and V) that are phase shifted 120 degrees from one another. The amplifier produces a balanced three-phase output. The U and V outputs are currents that are replicas of the reference inputs. The third phase, phase W, is generated as the inverted sum of the U and V signals thus producing the balanced, three-phase output. Current sensors are used to provide current feedback for the control loops.

The current sensor outputs are also fed back to the microcontroller (with integral A/D) for use by the overload protection algorithm (aka I^2T algorithm). The input voltages to two additional A/D channels are user programmable and provide the algorithm with measures of the rated continuous motor current and the motor I^2T rating. The I^2T algorithm is implemented independently on each output (phase U, V and W), but an overload detected on any one phase will interrupt current on all three outputs. A flow diagram of the algorithm for a single phase is provided in Fig. 1.

I^2T Motor Overload Protection Algorithm

Setup

1. Program amp w/ motor continuous current limit (Header component H4, H6 and H13)
2. Program amp w/ motor I^2T Limit (Header component H14)

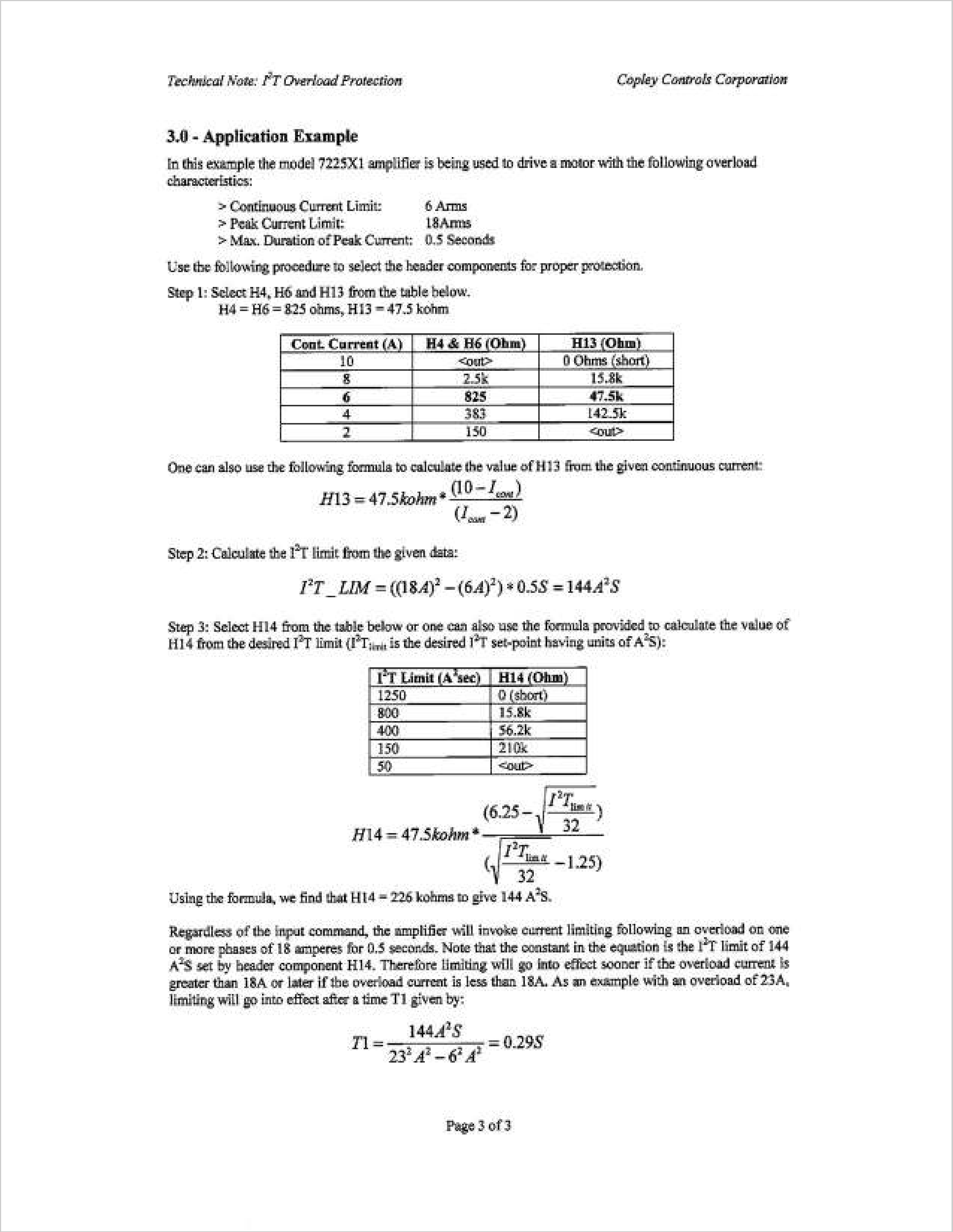


*Note that the "difference" can be either positive or negative, but the I²T tracking variable itself is limited to no less than zero

Figure 1 - I squared T overload protection algorithm as implemented for the Copley Model 7225X1 UV sine amplifier.

The flowchart is broken up into two sections. The first section on the left shows the initialization routine that occurs every time the unit is powered up. Three major events happen at initialization. First the main variable in the routine, the I^2T tracking variable, is given an initial value of zero. Second, the motor continuous current limit is read via the voltage present at one of the A/D inputs of the microcontroller. This voltage is determined by the header component H13. Third, the motor I^2T limit is read via the voltage present at another of the microcontroller A/D inputs. This voltage is determined by the header component H14.

The "run-time" portion of the routine is shown in the other section of the flowchart. This routine is interrupt driven by a 1mS timer and is implemented on each phase (U, V and W). The flowchart shows the implementation of just one phase for simplicity. At each interrupt, the output current is sampled via the A/D converter on board the microcontroller. The algorithm then calculates the difference between the square of the sample and the square of the continuous current limit (determined by H13 and read during initialization). This difference is then added to the present value of the I^2T tracking variable to determine an updated value for the I^2T tracking variable. Note that this difference can be either positive or negative, thus the I^2T tracking variable can grow or fall, but it can never have a value below zero. Once the I^2T tracking variable is updated it is then compared with the I^2T set-point (determined by H14 and read during initialization). If the I^2T tracking variable is greater than the set-point, the microcontroller invokes curent limiting and the amplifier output current is forced to a level no greater than the continuous limit set by H4 and H6. If the I^2T tracking variable is less than the set-point, no action is taken and the amplifier operates normally.



3.0 - Application Example

In this example the model 7225X1 amplifier is being used to drive a motor with the following overload characteristics:

> Continuous Current Limit:	6 Arms
> Peak Current Limit:	18Arms
> Max. Duration of Peak Current:	0.5 Seconds

Use the following procedure to select the header components for proper protection.

Step 1: Select H4, H6 and H13 from the table below.
H4 = H6 = 825 ohms, H13 = 47.5 kohm

Cont. Current (A)	H4 & H6 (Ohm)	H13 (Ohm)
10	<out>	0 Ohms (short)
8	2.5k	15.8k
6	**825**	**47.5k**
4	383	142.5k
2	150	<out>

One can also use the following formula to calculate the value of H13 from the given continuous current:

$$H13 = 47.5kohm * \frac{(10 - I_{cont})}{(I_{cont} - 2)}$$

Step 2: Calculate the I²T limit from the given data:

$$I^2T\_LIM = ((18A)^2 - (6A)^2) * 0.5S = 144A^2S$$

Step 3: Select H14 from the table below or one can also use the formula provided to calculate the value of H14 from the desired I²T limit (I^2T_{limit} is the desired I²T set-point having units of A^2S):

I²T Limit (A²sec)	H14 (Ohm)
1250	0 (short)
800	15.8k
400	56.2k
150	210k
50	<out>

$$H14 = 47.5kohm * \frac{(6.25 - \sqrt{\frac{I^2T_{limit}}{32}})}{(\sqrt{\frac{I^2T_{limit}}{32}} - 1.25)}$$

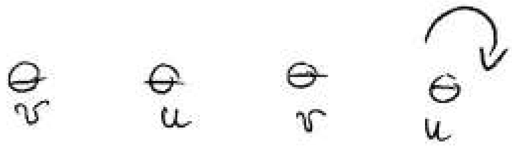
Using the formula, we find that H14 = 226 kohms to give 144 A²S.

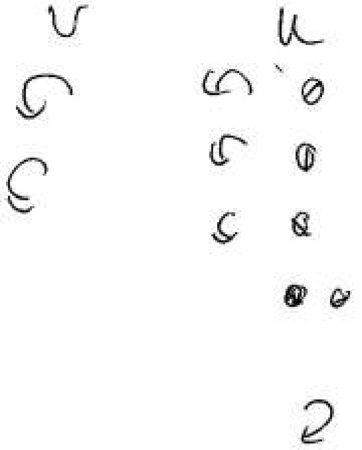
Regardless of the input command, the amplifier will invoke current limiting following an overload on one or more phases of 18 amperes for 0.5 seconds. Note that the constant in the equation is the I²T limit of 144 A²S set by header component H14. Therefore limiting will go into effect sooner if the overload current is greater than 18A or later if the overload current is less than 18A. As an example with an overload of 23A, limiting will go into effect after a time T1 given by:

$$T1 = \frac{144A^2S}{23^2A^2 - 6^2A^2} = 0.29S$$

It is important to note the algorithm behavior when the limit has been reached. Once the current limiting occurs, the output current in all phases will fall to a level less than or equal to the continuous current settings. When the output current has fallen below the continuous limit, the difference between the actual current and the continuous limit current will be negative and the I^2T tracking variable will fall. When the I^2T tracking variable falls below the I^2T limit, current limiting is turned off and the amplifier output then follows the output normally.

Note that under some conditions, the amplifier can limit cycle meaning that the limiting turns on and off periodically. This occurs when the U and./or V *inputs* remain at a level above the continuous current limit after limiting is invoked. In this case the limiting will force the current to a level below the continuous limit and thus cause the I^2T tracking variable to fall. When the I^2T tracking variable falls below the I^2T limit, limiting is turned off and the current will increase back to a level corresponding to the input. Since this level is greater than the continuous limit, the I^2T tracking variable will increase back up and eventually reach the I^2T limit again. As long as the input signals remain at a level higher than the continuous current limit, this cyclical operation will continue. The cycle period is influenced by all of the factors used in computing the I^2T algorithm, thus the greater the overload, the shorter the cycle period.





$\frac{v}{u} = 1.6185$

$\frac{u}{v} = 1.964$

$A/V = 0.5$

$V =$

I_{motor}

$V = I_{motor} \cdot \ell_{amp} \frac{V}{A} \cdot 4096 \frac{counts}{V}$

$I_{motor} = \cdot counts \cdot \frac{5}{4096} \frac{V}{counts} \cdot \frac{25}{10} \frac{A}{V}$

RBE(H) Motor Series

COMMUTATION AND CONNECTION DIAGRAMS

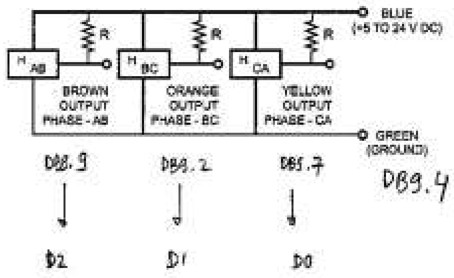
MOTOR WIRING DIAGRAM	
Phase "A"	Red Lead
Phase "B"	White Lead
Phase "C"	Black Lead

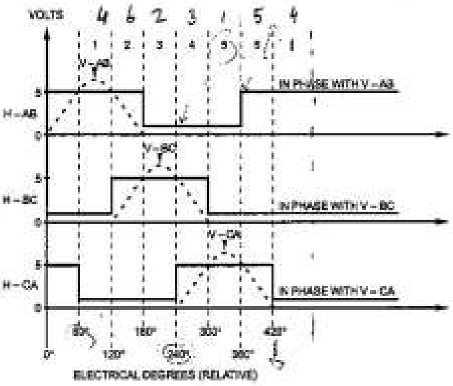
red
yellow
black

MOTOR EXCITATION SEQUENCE AND SENSOR OUTPUT LOGIC FOR C.W. ROTATION VIEWING LEADWIRE END

EXCITATION STEP		1	2	3	4	5	6	1
Motor Leads	(RED) A	+	+		-	-		+
	(WHT) B	-		+	+		-	-
	(BLK) C		-	-		+	+	
Sensor Outputs	(BRN) A	1	1	0	0	0	1	1
	(ORG) B	0	1	1	1	0	0	0
	(YEL) C	0	0	0	1	1	1	0
		4	⑥	2	3	①	5	

SENSOR WIRING DIAGRAM

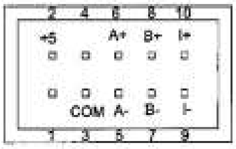




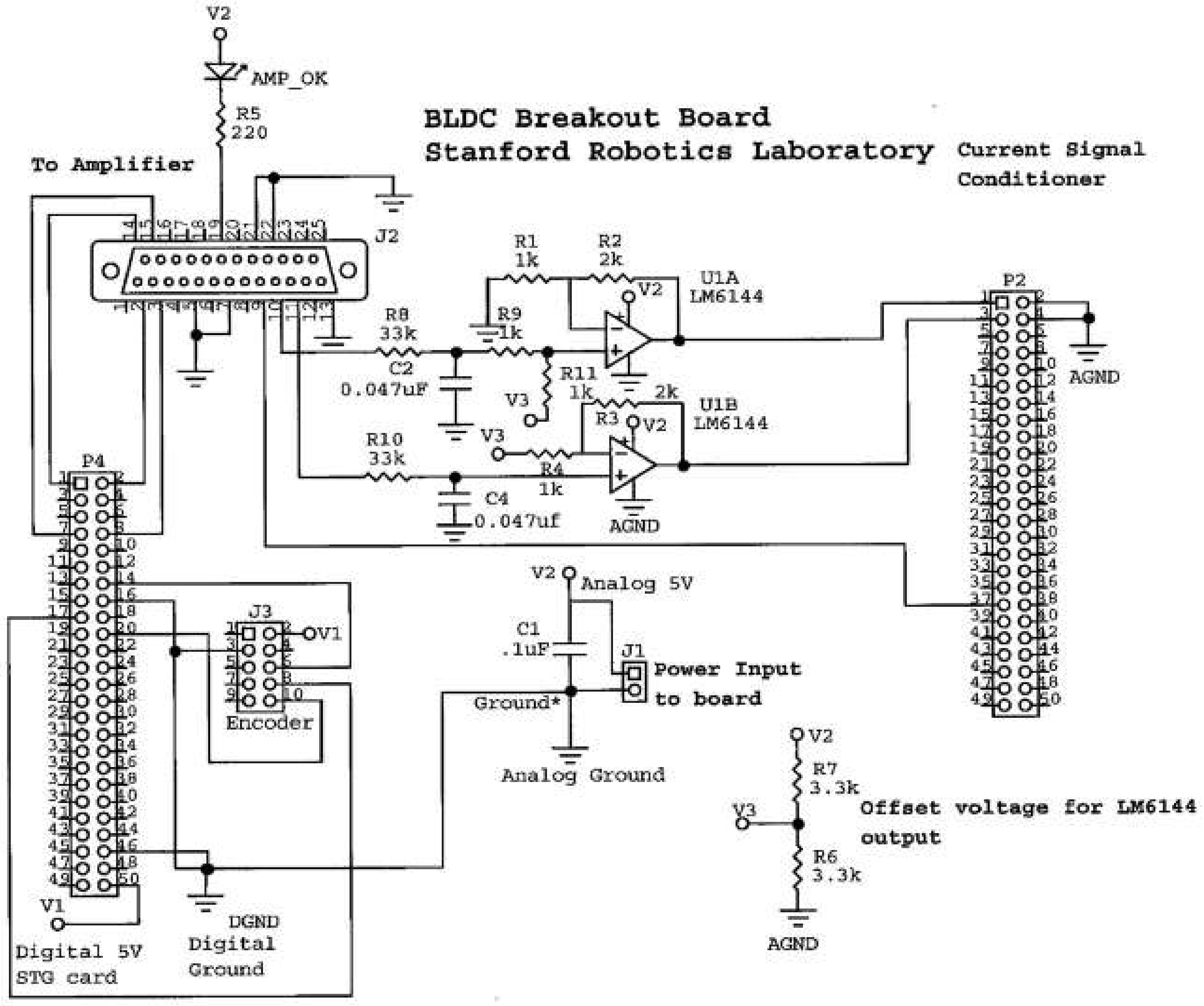
SENSOR AND MOTOR PHASE OUTPUT

ENCODER WIRING TABLE	
CH A LEADS CH B, C.C.W.	
PIN #	FUNCTION
PIN 1	NC
PIN 2	+5 V
PIN 3	COMMON
PIN 4	NC
PIN 5	DATA Ā
PIN 6	DATA A
PIN 7	DATA B̄
PIN 8	DATA B
PIN 9	INDEX Z̄
PIN 10	INDEX Z

△ ON CONECTOR IDENTIFIES
PIN #1 ON BOTH LD
AND MODULE



W = U + V


BLDC Breakout Board
Stanford Robotics Laboratory
Current Signal Conditioner
To Amplifier
V2
AMP_OK
R5
220
J2
R1
1k
R2
2k
U1A
LM6144
R8
33k
R9
1k
C2
0.047uF
V3
R11
1k
2k
R3
U1B
LM6144
R10
33k
R4
1k
C4
0.047uf
AGND
P2
P4
J3
V1
Encoder
Analog 5V
C1
.1uF
J1
Power Input to board
Ground*
Analog Ground
R7
3.3k
R6
3.3k
Offset voltage for LM6144 output
Digital 5V
STG card
DGND
Digital Ground

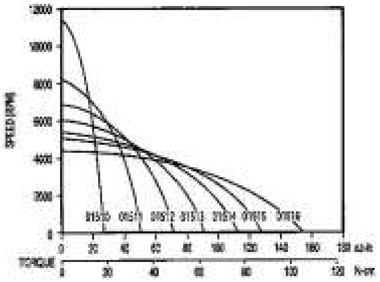
RBE(H) 01510 MOTOR SERIES PERFORMANCE DATA

Motor Parameters	Symbols	Units	01510	01511	01512	01513	01514	01515	01516
Max Cont. Output Power	HP Rated	HP	0.127	0.176	0.210	0.240	0.264	0.284	0.307
at 25°C amb.	P Rated	Watts	95	131	157	179	197	212	229
Speed at Rated Power	N Rated	RPM	7450	5400	4550	4050	3570	3400	2970
Max Mechanical Speed	N Max	RPM	16500	16500	16500	16500	16500	16500	16500
Continuous Stall Torque	Tc	oz-in	27.4	54.3	71.9	91.3	114	127	154
at 25°C amb.		N-m	0.193	0.384	0.508	0.645	0.808	0.897	1.085
Peak Torque	Tp	oz-in	78.6	162	234	313	403	540	610
		N-m	0.555	1.15	1.66	2.21	2.85	3.81	4.31
Max Torque	Tsl	oz-in	78.6	162	234	313	403	540	610
for Linear KT		N-m	0.555	1.16	1.66	2.21	2.85	3.81	4.31
Motor Constant	Km	oz-in/√W	6.38	11.6	14.8	18.2	22.1	24.1	28.6
		N-m/√W	0.0451	0.0819	0.105	0.128	0.156	0.170	0.202
Thermal Resistance*	Rth	°C/Watt	4.10	3.55	3.30	3.13	2.95	2.85	2.72
Viscous Damping	Fi	oz-in/RPM	2.74E-04	1.05E-03	1.76E-03	2.47E-03	3.32E-03	3.88E-03	5.30E-03
		N-m/RPM	1.94E-06	7.43E-06	1.24E-05	1.74E-05	2.34E-05	2.74E-05	3.74E-05
Max Static Friction	Tf	oz-in	2.00	2.93	3.77	4.62	5.63	6.31	8.00
		N-m	0.0141	0.021	0.027	0.033	0.040	0.045	0.057
Max Cogging Torque	Tcog	oz-in	0.950	1.22	1.47	1.71	2.01	2.21	2.70
Peak to Peak		N-m	0.00671	0.00862	0.0104	0.0121	0.0142	0.0156	0.019
Frameless Motor Inertia	Jmf	oz-in-sec²	2.10E-03	3.60E-03	4.90E-03	6.20E-03	7.70E-03	8.80E-03	1.14E-02
		Kg-m²	1.48E-05	2.54E-05	3.46E-05	4.38E-05	5.44E-05	6.21E-05	8.05E-05
Frameless Motor Weight	Wtf	oz	6.30	10.5	14.3	18.1	22.7	25.8	33.4
		Kg	1.79E-01	2.98E-01	4.06E-01	5.14E-01	6.44E-01	7.30E-01	9.47E-01
Housed Motor Inertia	Jmh	oz-in-sec²	2.20E-03	3.70E-03	5.00E-03	6.30E-03	7.80E-03	8.90E-03	1.15E-02
		Kg-m²	1.55E-05	2.61E-05	3.53E-05	4.45E-05	5.51E-05	6.28E-05	8.12E-05
Housed Motor Weight	Wth	oz	19.0	23.5	27.5	31.6	36.4	39.7	47.8
		Kg	5.39E-01	6.65E-01	7.80E-01	8.95E-01	1.03E+00	1.13E+00	1.38E+00
No. of poles	P		12	12	12	12	12	12	12

Winding Constants	Symbols	Units	A	B	C	A	B	C	A	B	C	A	B	C	A	B	C	A	B	C	A	B	C
Current at Cont. Torque	Ic	Amps	5.10	3.71	9.06	4.85	5.53	4.98	4.44	3.23	7.90	4.22	3.07	7.50	4.62	2.94	7.18	5.13	2.83	6.91	5.18	2.59	6.34
Current at Peak Torque	Ip	Amps	14.0	9.89	25.0	14.0	9.89	16.7	14.0	9.89	25.0	14.0	9.89	25.0	15.7	9.89	25.0	21.0	9.89	25.0	19.8	9.89	25.0
Torque Sensitivity	Kt	oz-in/Amp	6.78	7.92	3.24	11.6	16.2	11.5	17.0	23.4	9.56	22.7	31.2	12.8	26.0	40.9	16.7	26.0	47.2	19.3	31.2	62.5	25.5
		N-m/Amp	0.0407	0.0559	0.0229	0.0833	0.115	0.0812	0.120	0.165	0.0675	0.160	0.220	0.0903	0.184	0.289	0.118	0.184	0.333	0.136	0.220	0.440	0.180
Back EMF constant	Kb	V/KRPM	4.26	5.86	2.40	8.73	12.0	8.50	12.6	17.3	7.07	16.8	23.1	9.43	19.2	30.2	12.4	19.2	34.9	14.3	23.1	46.1	18.9
Motor Resistance	Rm	Ohms	0.814	1.58	0.256	1.04	2.02	0.988	1.53	2.59	0.438	1.55	3.03	0.489	1.38	3.45	0.597	1.16	3.86	0.623	1.19	4.75	0.789
Motor Inductance	Lm	mH	0.32	0.61	0.101	3.58	1.1	0.55	0.87	1.6	0.27	1.2	2.3	0.38	1.1	2.6	0.47	0.99	3.3	0.55	1.1	4.4	7.4

*Rth assumes a housed motor mounted to a 4" x 3.25" x 0.25" aluminum heatsink or equivalent

Continuous Duty Capability for 130°C Rise — RBE - 01510 Series



Motor Parameters

Motor parameters are listed on the individual data page for each motor. These parameters are dependent upon the size and shape of the model, but are independent of the winding used. Following is a brief description of the motor parameters.

Maximum Continuous Output Power at 25°C Ambient (HP Rated). This is the maximum continuous power output based on a 130°C temperature rise and a standard aluminum heat sink. (Standard heat sink size is listed just above the continuous performance curves). The maximum continuous power output can be increased if additional cooling is provided.

Speed at Rated Power (N Rated) is the speed at which the maximum continuous power is output.

Maximum Mechanical Speed (N Max) is the maximum speed which will not compromise rotor integrity.

Continuous Stall Torque at 25°C Ambient (Tc) is the maximum constant torque without rotation resulting in a steady state winding temperature rise of 130°C with the standard aluminum heat sink. The size of the standard heat sink is listed above the continuous performance curve for each RBE(H) series. The continuous stall torque can be increased if additional cooling is provided.

Peak Torque (Tp) is the maximum torque available from a given size of motor and is the torque the motor will provide when peak current **Ip** is provided. Peak torque is based on the maximum current density in the winding and is available for a maximum duration of 10 seconds.

Maximum Torque for Linear KT (Tsl) is the maximum torque for which **Kt** will be greater than 90 percent of **Kt** at low torque. As the torque increases above **Tsl, Kt** will drop below 90 percent of **Kt** at low torque and an incremental increase in current will yield a reduced increase in torque.

Motor Constant (Km) is the ratio of peak torque to the square root of power input at 25°C and at stall:

$Km = Tp/(Pp)^{.5}$

This ratio is useful during the initial selection of a motor, because it indicates the ability of a motor to convert electrical power into torque. A common use of Km is to determine how much power a motor will dissipate in order to generate a certain amount of torque by using the following equation:

$\text{Watts Dissipated} = \text{Torque}^2 / \mathbf{Km}^2$

Thermal Resistance (Rth) is the ratio of winding temperature rise to average power losses continuously dissipated from the stator. Motor **Rth** values assume a standard aluminum heat sink which is specified above the continuous speed torque curve for each RBE(H) series. Customer supplied supplemental cooling can reduce the **Rth** value significantly resulting in increased continuous speed and torque operation.

Viscus Damping (Fi) is the torque loss due to rotational losses, mostly eddy current, which is proportional to speed. A lower **Fi** indicates less loss during high speed operation.

Maximum Static Friction (Tf) is the sum of the retarding torques at start-up or at stall within the motor. In a frameless brushless motor, retarding torques consist of magnetic frictional torque and cogging torque. Housed motor TF includes bearing and other retarding torques.

Maximum Cogging Torque (Tcog) is a torque disturbance based on the magnets in the field attraction to the teeth in the armature. Cogging torque is minimized in the motor design by strategic selection of slot / pole combinations and by skewing the laminations in the armature.

Number of Poles (P) is the number of magnetic poles in the field. The electrical cycles per revolution is equal to the number of poles to the number of poles divided by 2.

$$\frac{\dot{q}}{q} = \frac{s\omega}{s + \omega}$$

$$\omega = 2\ \text{Hz} \approx 12\ \text{rad}$$

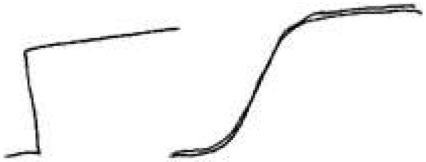
$$T = 100\ \text{Hz}$$

181.71 ?

$$\frac{\dot{q}_k}{q_k} = \frac{6.26z - 6.26}{z - 0.9937}$$

$$\dot{q}_k = 0.9937\,\dot{q}_{k-1} + 6.26(q_k - q_{k-1})$$

$$k_v = 2\omega_n\zeta \qquad \omega_n = \sqrt{k_p}$$



k_v

$$C = k_p(q_d - q) - k_v(\dot{q})$$

$$\dot{q}_d = \frac{k_p}{k_v}(q_d - q)$$

$$C = -k_v\left(\dot{q} - \frac{k_p}{k_v}(q_d - q)\right)$$

$$C = k_v(\dot{q}_d - \dot{q})$$

USING THE INPUT CAPTURE MODULE

Hall Sensors A, B and C are connected to IC1, IC2 and IC3, respectively, on the Input Capture (IC) module. The Input Capture module is used in "Input Capture on State Change" mode. In this mode, the IC module interrupts every transition on any of the IC pins. Also, Timer5 is captured on every transition and cleared at the beginning of the next clock cycle. The captured Timer5 value is useful in determining the speed of the motor. Measuring the speed and controlling the motor in closed loop is discussed in detail in the section **"Closed-Loop Control Using Hall Sensors"**.

Upon IC interrupt, in the IC Interrupt Service Routine, the status of all three input capture pins is read and the combination is used to pick up the correct sequence from the table.

Table 1 shows a typical switching sequence used to run the motor in the clockwise direction and Table 2 shows the counterclockwise sequence. These tables are taken directly from the motor data sheet[1].

Note 1: Motor Data Sheet
Manufacturer: Bodine Electric Company
Type Number: 22B4BEBL
Series: 3304
Web Site: www.bodine-electric.com

If the motor you have uses a different sequence, it should be entered in the firmware. Figure 2 shows the relationship between the motor phase current and the Hall Sensor inputs and the corresponding PWM signals to be activated to follow the switching sequence, which in turn, runs the motor in the clockwise direction.

TABLE 1: SEQUENCE FOR ROTATING THE MOTOR IN CLOCKWISE DIRECTION WHEN VIEWED FROM NON-DRIVING END

Sequence Number	Hall Sensor Input			Active PWMs		Phase Current		
	A	B	C			A	B	C
1	0	0	1	PWM1(Q1)	PWM4(Q4)	DC+	Off	DC-
2	0	0	0	PWM1(Q1)	PWM2(Q2)	DC+	DC-	Off
3	1	0	0	PWM5(Q5)	PWM2(Q2)	Off	DC-	DC+
4	1	1	0	PWM5(Q5)	PWM0(Q0)	DC-	Off	DC+
5	1	1	1	PWM3(Q3)	PWM0(Q0)	DC-	DC+	Off
6	0	1	1	PWM3(Q3)	PWM4(Q4)	Off	DC+	DC-

TABLE 2: SEQUENCE FOR ROTATING THE MOTOR IN COUNTERCLOCKWISE DIRECTION WHEN VIEWED FROM NON-DRIVING END

Sequence Number	Hall Sensor Input			Active PWMs		Phase Current		
	A	B	C			A	B	C
1	0	1	1	PWM5(Q5)	PWM2(Q2)	Off	DC-	DC+
2	1	1	1	PWM1(Q1)	PWM2(Q2)	DC+	DC-	Off
3	1	1	0	PWM1(Q1)	PWM4(Q4)	DC+	Off	DC-
4	1	0	0	PWM3(Q3)	PWM4(Q4)	Off	DC+	DC-
5	0	0	0	PWM3(Q3)	PWM0(Q0)	DC-	DC+	Off
6	0	0	1	PWM5(Q5)	PWM0(Q0)	DC-	Off	DC+

1
6
4
5
2
3

Introduction to
Hurst Brushless DC Motors

Brushless Motor Construction

DC brushless motors are similar in performance and application to brush-type DC motors. Both have a speed vs. torque curve which is linear or nearly linear. The motors differ, however, in construction and method of commutation. A brush-type permanent magnet DC motor usually consists of an outer permanent magnet field and an inner rotating armature. A mechanical arrangement of commutator bars and brushes switches the current in the armature windings to maintain rotation. A DC brushless motor has a wound stator, a permanent magnet rotor assembly, and internal or external devices to sense rotor position. The sensing devices provide signals for electronically switching (commutating) the stator windings in the proper sequence to maintain rotation of the magnet assembly.

The rotor assembly may be internal or external to the stator in a DC brushless motor. The combination of an inner permanent magnet rotor and outer windings offers the advantages of lower rotor inertia and more efficient heat dissipation than DC brush-type construction. The elimination of brushes reduces maintenance, increases life and reliability and reduces noise and EMI generation.

Brushless Motor Commutation

The possible number of phases and winding arrangements for the DC brushless stator are quite varied. As in the case of brush-type DC motors, increasing the number of phases reduces torque ripple. However, an important practical consideration for DC brushless motors is the number of electronic switches required to commutate the phases. Three phase motors provide a compromise in this regard and are popular in many applications.

The winding arrangement for a three phase motor may be either a Y or a Δ configuration. The most efficient operation of the motor requires current flow in more than one phase at any instant and current reversal in each of the phases at some point during 360 electrical degrees of rotation. This in turn requires a minimum of two electronic switches per phase. It may be noted that the Y configuration with a lead common to the three phases can be commutated in a unipolar mode with only three electronic switches. However, the motor torque is reduced with this scheme.

Hall Effect Commutation

Rotor position sensing is essential for proper commutation of DC brushless motors. Magnetic sensing with inexpensive Hall effect switches is frequently adequate. The devices require little space and can easily be placed within the motor. Optical encoders or resolvers may also be used. Cost, operating environment of the motor, intended application and performance all influence the choice.

As an example of Hall effect commutation, consider the typical commutation scheme shown in the table for the Hurst motors. Commutation of a three phase motor with current reversal requires that the windings be switched every 60 electrical degrees. If three sensing devices are spaced 120 electrical degrees apart, and if each device has a 50% duty cycle, six discrete 3-bit signal states are produced at 60 degree intervals as the rotor turns. Each change of state triggers switching of the stator windings to a particular terminal pair and polarity.

Reversing

The sensors are located so that switching occurs 30 electrical degrees before the peak in the torque vs. angle curve for that terminal pair. Operation of the motor in the reverse direction requires only a change in the switching sequence. A number of manufacturers offer an integrated circuit to perform the sensor signal decoding and provide signals to sequence the power switches for the motor phases.

Brushless Motor Control

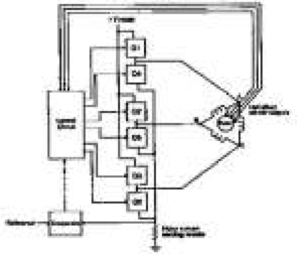
DC brushless motors are used in the same types of applications as DC brush-type motors, e.g., servo, constant speed, variable speed, controlled torque, etc. The methods of control are similar to those for brush-type motors. Most will involve some type of current control whether in an open loop mode or a closed loop mode with position and perhaps velocity sensing. A description of a typical drive circuit will illustrate the possibilities.

The power switches for a three phase motor are usually arranged in three half bridges as shown in the figure. Power MOSFETs of bipolar devices may be used. Provision is normally made in the control circuit to delay turn on of one device in each leg until turn off of the other device is complete to prevent shorting the power supply. Motor current may be sensed with a single resistor in the common lead. For more information about controls, please consult the factory.

Current Limiting

In current limiting or speed control applications the sensed current may then be compared with a fixed or variable reference and the resulting signal used to control motor current (speed, torque, etc.). Current control schemes include pulse width modulation or linear operation of the power switches. Dynamic braking is accomplished with this circuit by turning off the top switch and turning on the bottom switch in each leg of the bridge. The winding currents are then short circuited through the bottom switches. As DC brushless motor applications have increased, many of these control features have been incorporated into integrated circuit controllers.

DC Brushless Motor Control Block Diagram

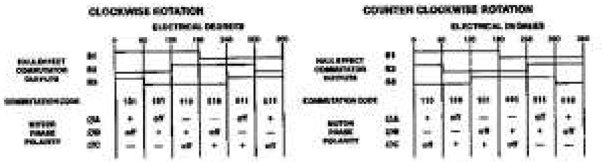


COMMUTATOR CHARACTERISTICS

PARAMETER	VALUE
COMMUTATOR TYPE	HALL EFFECT SWITCH
INPUT VOLTAGE (Vcc)	4.5 TO 24VDC*
PHASING	120° ELECTRICAL
OUTPUT TYPE	OPEN COLLECTOR
SINK CURRENT	20ma MAX

*Maximum input voltage is 5.5VDC for motors with encoders.

COMMUTATION SEQUENCE



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LOW COST HIGH EFFICIENCY SENSORLESS DRIVE FOR BRUSHLESS DC MOTOR USING MC68HC(7)05MC4

By Leos Chalupa
Roznov System Application Laboratory
Motorola, Roznov pod Radhostem, Czech Republic

1 INTRODUCTION

This application note presents a low-cost sensorless speed control system for Brushless DC (BLDC) motors. Especially suitable when high efficiency and low-price are required, the drive system is based on a high efficiency Brushless DC motor, sensorless rotor position detection, low cost components and the motor control dedicated low-cost microcontroller (the Motorola's MC68HC05MC4).

Today more and more variable speed drives are put in appliance products to increase the whole system efficiency and the product performance.The low dynamic drive is a solution in many cases for common appliance applications, whereby the load or speed is changed quite slowly in comparison with the system mechanical time constant. Quite simple algorithms can perform the control tasks. Moreover, the necessary computing power can be minimised by using a dedicated microcontroller and efficient use of its internal function blocks (such as A/D converter, dedicated PWM outputs, input capture and output compare functions...).

Three phase Brushless DC (BLDC) motors are good candidates because of their high efficiency capability and easy to drive features. The disadvantage of this kind of motor is the fact that commutation of motor phases relies on its rotor position. Although the rotor position is usually sensed by sensors, the requirement for sensorless solutions becomes more and more real.

The sensorless rotor position technique developed detects the zero crossing points of Back-EMF induced in the motor windings. The phase Back-EMF zero crossing points are sensed while one of the three phase windings is not powered. The obtained information is processed in order to control the phase voltage, using Pulse Width Modulation.

This application note tries to give a fundamental mathematical method for modelling, torque calculation and control concept of the presented drive. The drive was developed in order to drive simple applications (e.g. pumps, compressors, fans...) within certain ranges of speed and load. Results from simulation show the drive behaviour at different working conditions and better explains the drive strategy.

1.1 Classical System

As is well known, the Brushless DC motor (BLDC motor) is also named electronically commuted motor. There are no brushes on the rotor and the commutation is performed electronically at certain rotor positions. The three phase voltage system (see Figure 1-1.), with a rectangular shape, is used to create a rotational field. This easy to create shape of applied voltage ensures the simplicity of control and drive. The rotor position must be known at certain angles in order to align the applied voltage with the Back-EMF, which is induced in the stator winding due to the movement of the permanent magnets on the rotor. The alignment between Back-EMF and commutation events is very important. At this condition the motor behaves as a DC motor and runs at the best working point. The simplicity of control and good performance makes this motor a natural choice in low-cost and high-efficiency applications.

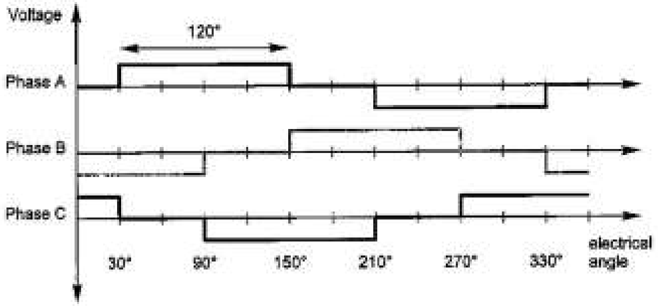


Figure 1-1. Three Phase Voltage System

1.2 Why Sensorless?

As explained in the previous section, the rotor position must be known in order to drive a Brushless DC motor. If any sensors are used to detect rotor position, then sensed information must be transferred to a control unit (see Figure 1-2.). Therefore additional connections to the motor are necessary. This may not be acceptable for some kind of applications. There are at least two reasons why you might want to eliminate the position sensors:

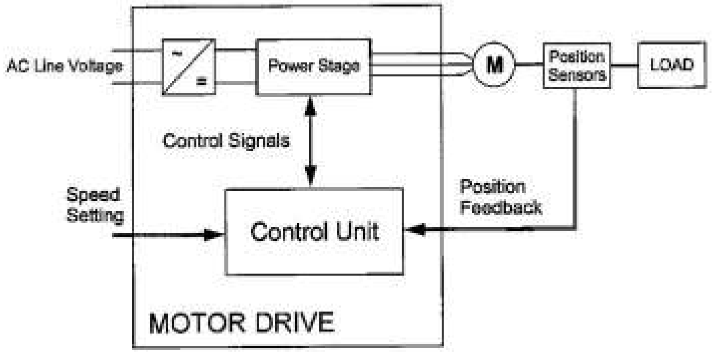


Figure 1-2. Classical System

1. Real impossibility to make additional connections between position sensors and the control unit
2. Cost of the position sensors & wiring

The first point might be solved by integration of the driver within the motor body. However a significant number of applications requiring a sensorless solution still remain.

2 THEORY

2.1 Power Stage - Motor System Model

In order to explain and simulate the idea of Back-EMF sensing techniques a simplified mathematical model, based on the basic circuit topology (see Figure 2-1.), has been created.

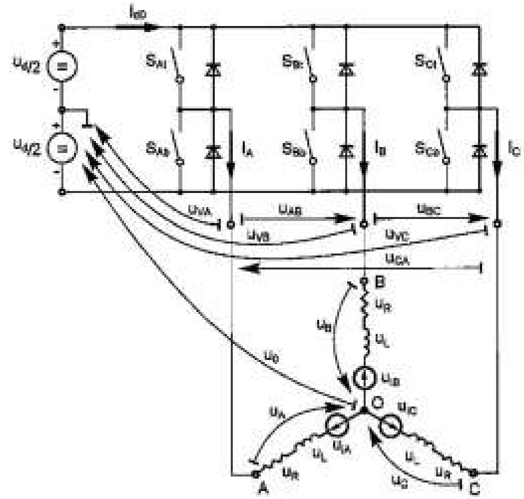
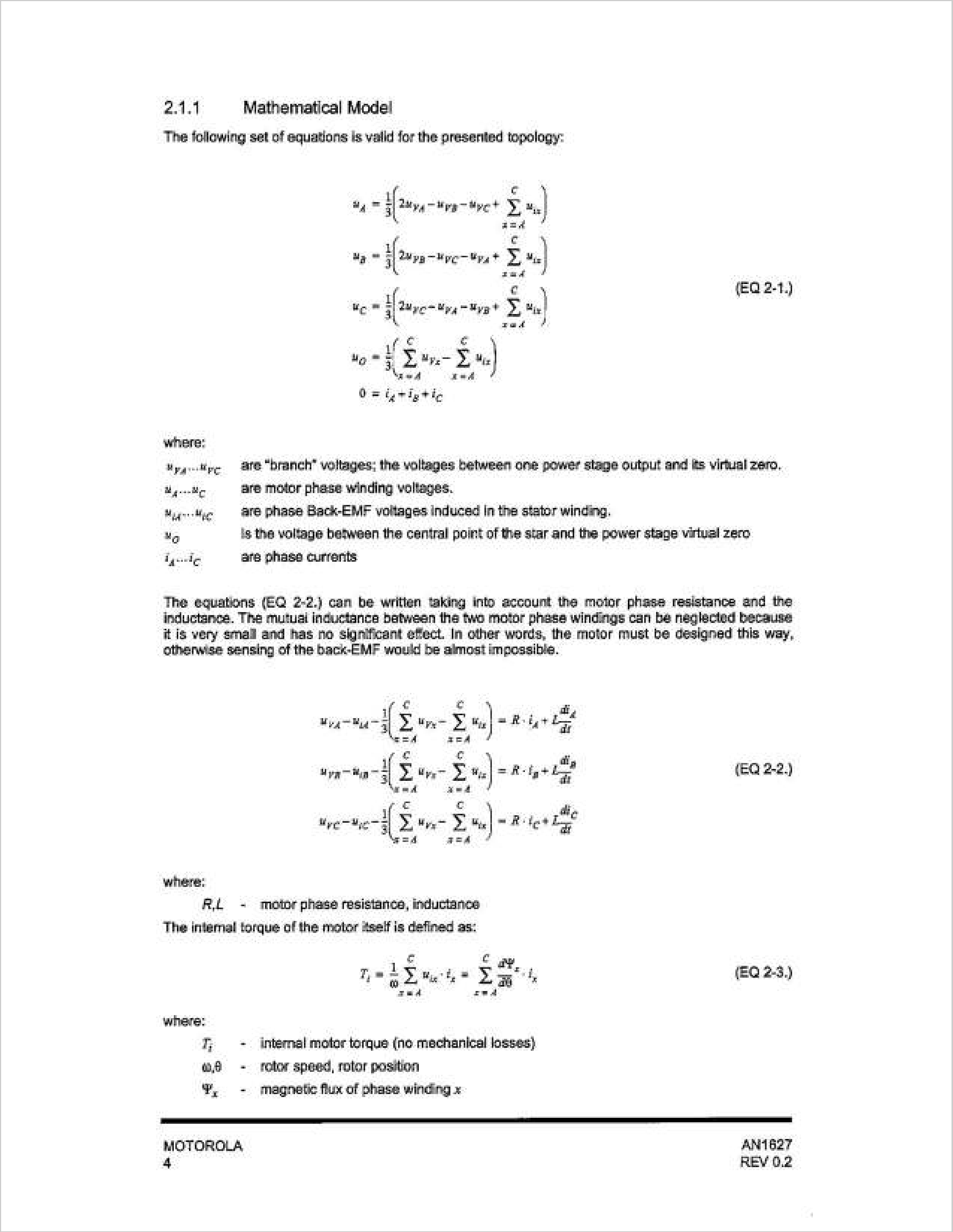


Figure 2-1. Power Stage - Motor Topology

The second goal of the model is to find how the motor characteristics depend on the switching angle. The **switching angle** is the angular difference between a real switching event and an ideal one (at the point where the **phase to phase** Back-EMF crosses zero).

The motor-drive model consists of a normal three phase power stage plus a Brushless DC motor. The power for the system is provided by a voltage source (U_d). Six semiconductor switches ($S_{A/B/C\ t/b}$), controlled elsewhere, allow the rectangular voltage waveforms (see Figure 1-1.) to be applied. The semiconductor switches and diodes are simulated as ideal devices. The zero voltage level of the whole model is put at one half of the DC-Bus voltage. This simplifies the mathematical expressions when output waveforms from the power stage are calculated.



2.1.1 Mathematical Model

The following set of equations is valid for the presented topology:

$$
\begin{aligned}
u_A &= \frac{1}{3}\left(2u_{VA} - u_{VB} - u_{VC} + \sum_{x=A}^{C} u_{ix}\right)\\
u_B &= \frac{1}{3}\left(2u_{VB} - u_{VC} - u_{VA} + \sum_{x=A}^{C} u_{ix}\right)\\
u_C &= \frac{1}{3}\left(2u_{VC} - u_{VA} - u_{VB} + \sum_{x=A}^{C} u_{ix}\right)\\
u_O &= \frac{1}{3}\left(\sum_{x=A}^{C} u_{Vx} - \sum_{x=A}^{C} u_{ix}\right)\\
0 &= i_A + i_B + i_C
\end{aligned}
\qquad \text{(EQ 2-1.)}
$$

where:

$u_{VA} \ldots u_{VC}$	are "branch" voltages; the voltages between one power stage output and its virtual zero.
$u_A \ldots u_C$	are motor phase winding voltages.
$u_{iA} \ldots u_{iC}$	are phase Back-EMF voltages induced in the stator winding.
u_O	is the voltage between the central point of the star and the power stage virtual zero
$i_A \ldots i_C$	are phase currents

The equations (EQ 2-2.) can be written taking into account the motor phase resistance and the inductance. The mutual inductance between the two motor phase windings can be neglected because it is very small and has no significant effect. In other words, the motor must be designed this way, otherwise sensing of the back-EMF would be almost impossible.

$$
\begin{aligned}
u_{VA} - u_{iA} - \frac{1}{3}\left(\sum_{x=A}^{C} u_{Vx} - \sum_{x=A}^{C} u_{ix}\right) &= R \cdot i_A + L\frac{di_A}{dt}\\
u_{VB} - u_{iB} - \frac{1}{3}\left(\sum_{x=A}^{C} u_{Vx} - \sum_{x=A}^{C} u_{ix}\right) &= R \cdot i_B + L\frac{di_B}{dt}\\
u_{VC} - u_{iC} - \frac{1}{3}\left(\sum_{x=A}^{C} u_{Vx} - \sum_{x=A}^{C} u_{ix}\right) &= R \cdot i_C + L\frac{di_C}{dt}
\end{aligned}
\qquad \text{(EQ 2-2.)}
$$

where:

R,L - motor phase resistance, inductance

The internal torque of the motor itself is defined as:

$$T_i = \frac{1}{\omega}\sum_{x=A}^{C} u_{ix} \cdot i_x = \sum_{x=A}^{C} \frac{d\Psi_x}{d\theta} \cdot i_x \qquad \text{(EQ 2-3.)}$$

where:

T_i - internal motor torque (no mechanical losses)

ω,θ - rotor speed, rotor position

Ψ_x - magnetic flux of phase winding x

It is important to understand how the Back-EMF can be sensed and how the motor behaviour depends on the alignment of the Back-EMF to commutation events. This is explained in the next sections.

2.2 Back-EMF Sensing

The Back-EMF sensing technique is based on the fact that only two phases of a DC Brushless motor are connected at a time (see Figure 1-1.), so the third phase can be used to sense the Back-EMF voltage.

Let us assume the situation when phases A and B are powered and phase C is free. No current is going through this phase. This is described by the following conditions:

$$\begin{aligned} &S_{At}, S_{Bt} \leftarrow PWM \\ &u_{VA} = \mp\frac{1}{2}u_d, u_{VB} = \pm\frac{1}{2}u_d \\ &i_A = -i_B, i_C = 0, di_C = 0 \\ &u_{iA} + u_{iB} + u_{iC} = 0 \end{aligned} \qquad \text{(EQ 2-4.)}$$

The branch voltage C can be calculated when considering the above conditions:

$$u_{VC} = \frac{3}{2}u_{iC} \qquad \text{(EQ 2-5.)}$$

It is seen from the Figure 2-1. that the branch voltage of phase C can be sensed between the power stage output C and the zero voltage level. Thus the Back-EMF voltage is obtained and the zero crossing can be recognized.

The same expressions can also be found for phase A and B:

$$u_{Vx} = \frac{3}{2}u_{ix} \ldots x = A, B, C \qquad \text{(EQ 2-6.)}$$

There are two necessary conditions which have to be met -

- Top and bottom switch (in diagonal) have to be driven with the same PWM signal
- No current is going through the phase used to sense the Back-EMF

The Figure 2-2. shows branch and motor phase winding voltages during a 0-360°electrical interval. Shaded rectangles designate the validity of the equation (EQ 2-6.). In other words, the Back-EMF voltage can be sensed during designated intervals.

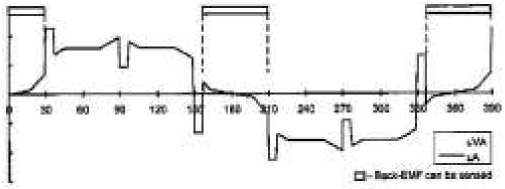


Figure 2-2. Phase Voltage Waveforms

2.3 Motor Behaviour at steady state condition

The previously described model is used to obtain the typical waveforms at different switching angles. See Figure 2-3., Figure 2-4. and Figure 2-5.

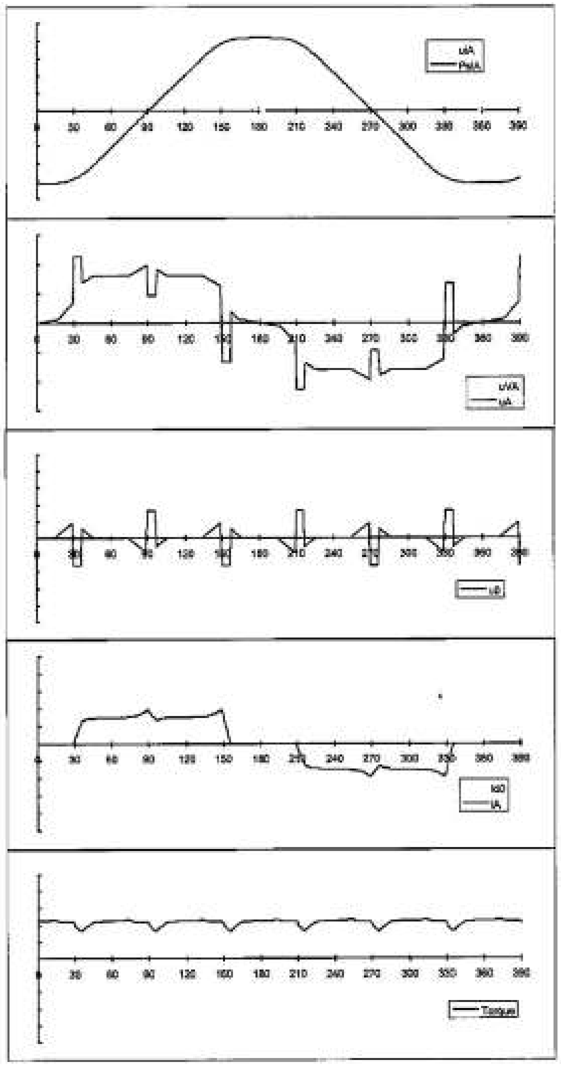


Figure 2-3. 0° switching angle

The x-axis represents electrical degrees of rotation. The y-axis represents relative amplitudes only. u_{VA} is voltage created by the power stage and applied to the motor. u_A is the voltage across the motor phase winding. A detailed look at waveforms u_{VA} and u_A shows one interesting fact. These voltages can differ quite a lot because the three phase system, created by the power stage, is not symmetrical. If it was symmetrical, u_0 would equal zero. The expression of u_0 (see (EQ 2-1.)) explains how this depends on the motor construction ($u_{iA,B,C}$) and the implemented three phase system ($u_{vA,B,C}$). The Back-EMF of phase A can be sensed within the following intervals: 330°-30° and 150°-210°.

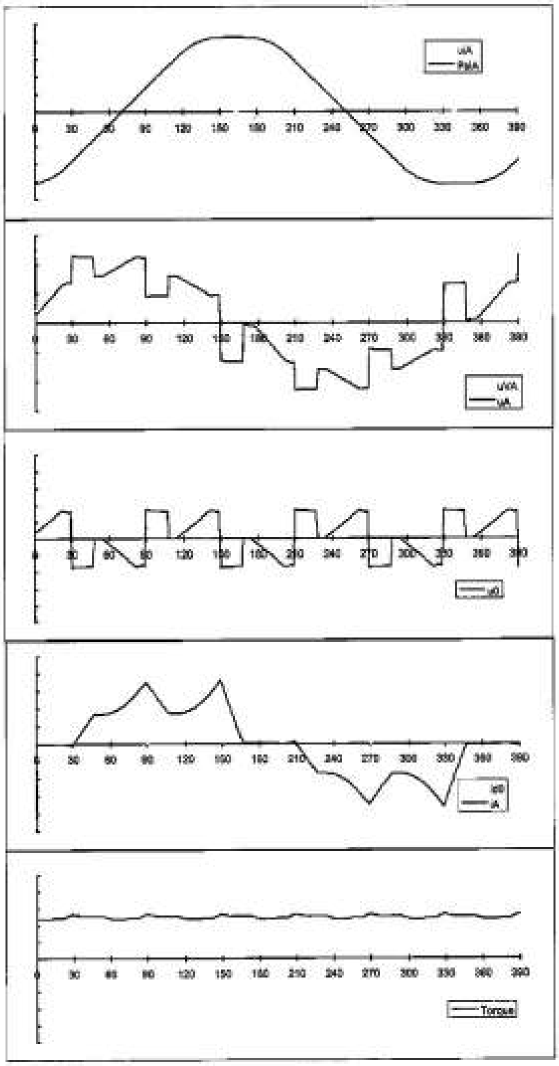


Figure 2-4. -20° switching angle

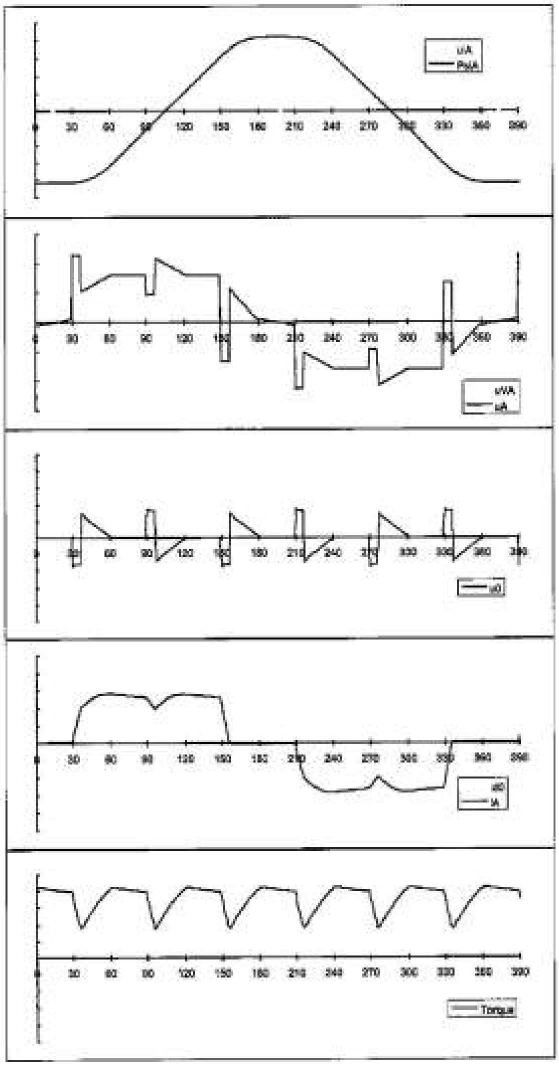


Figure 2-5. 15° switching angle

Let's suppose that the motor is fed from a voltage source, the fundamental frequency created by the power stage is constant. Figure 2-6. shows how the average motor torque (y-axis, relative amplitudes only) depends on the switching angle (x-axis) at constant speed and applied voltage (PWM duty cycle value) as a parameter. The working point can be found for a certain load torque and PWM duty cycle. If the load torque changes and the PWM duty cycle remains constant, the switching angle changes accordingly. The zero crossing event is used to calculate the actual switching angle. Once it is known, a corrective action can be made to keep the switching angle within the desired interval. This will allow control of the motor at the best efficiency point without need of position sensors.

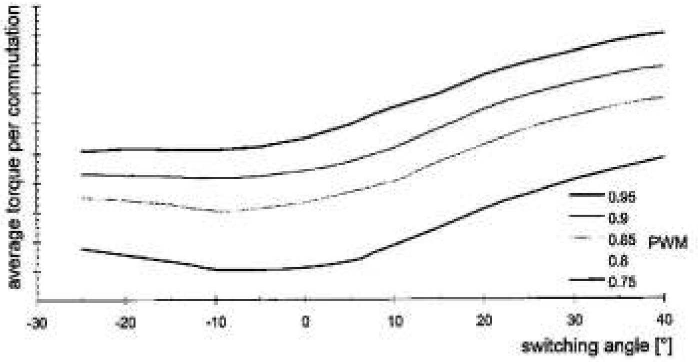


Figure 2-6. The Average Motor Torque

3 USUAL REQUIREMENTS

As was shortly introduced in the previous sections, the low-dynamic application is the field where this drive can be used. The speed and load vary during a much longer time than the system mechanicalt ime constant.

The Table 3-1. summarizes the requirements of a typical (low-cost high-volume) application with a Brushless DC motor.

Motor type:	6 stator poles, three-phases, four rotor poles, EC PM DC Motor	Specifications
Motor characteristic:	Speed range:	300-3000 RPM
	Torque range:	10%-120% of nominal
	Max. electrical power:	<300W
Drive input characteristic:	Input "inrush" max.:current	<30A
	Target efficiency:	0.9
	Min efficiency:	0.8 over whole range
	Fulfil European Community Regulation IEC555-1	
Load characteristic:	Type	Varying
	Start-up torque	Max. 150% of nominal
	Max. acceleration time: (0 to rated speed)	10sec

Table 3-1.

4 SYSTEM CONCEPT

The concept below was chosen. The cost target, especially, forced the design to be as simple as possible.

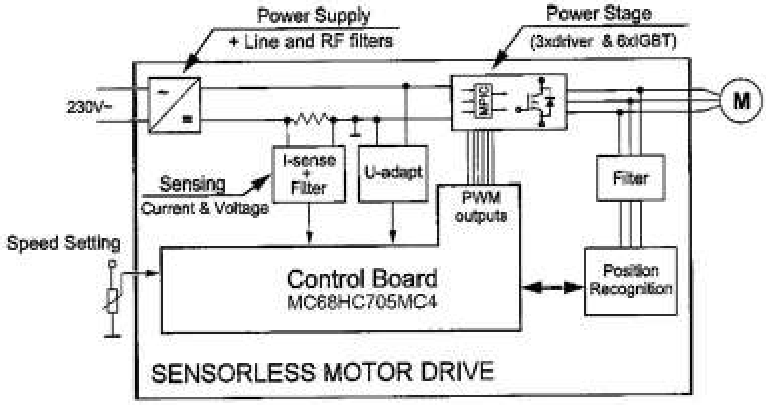


Figure 4-1. System

The 68HC705MC4 is an HC05-based MCU designed for three-phase Brushless DC motor (permanent magnet) drive applications. Generalf eatures include 3.5 kb of EPROM, 176 bytes of RAM, a 16-bit timer including an output compare and two input captures, 4 general-purpose I/O pins, and an SCI (UART) port in a 28-pin SOIC or DIP package. In addition the MC4 has specific features that target Brushless DC motor control, including a 2-channel, 8-bit PWM module, a high current source port, and a 6-channel, 8-bit A/D module.

Key features of the 6-pin, 2-channel PWM module include 16 PWM rates between 122Hz and 23.4kHz for each PWM channel, buffered data registers with an interlocking mechanism for coherent updates of the pulse width outputs on each PWM channel, and a 3-output commutation MUX on each PWM channel for easy control of outputs to the motor drive. The six PWM outputs of the commutation MUX are connected to an output port with 10 mA current source capability per pin, thus reducing the cost of external components required for building motor drives.

The Back-EMF zero crossing detection enables position recognition, as explained in previous sections. The resistor network is used to divide sensed voltages down to a 0-15V voltage level. Simple filtering prevents the comparators being disturbed by high voltage spikes produced by the switching of the IGBT's. The multiplexer selects the phase comparator output which corresponds to the current commutation stage. This signal is transferred to the microcontroller Input Capture pin (TCAP2).

The voltage drop resistor (0,6Ohm/2W) is used to measure the DC-bus current which is chopped by the PWM. The obtained signal is rectified and amplified (0-5V). The internal microcontroller A/D converter is synchronized with the PWM signal. This synchronization avoids spikes when the IGBT's are switching and simplifies the electric circuit.

The A/D converter is also used to sense the DC-Bus Voltage and speed setting. The DC-Bus voltage is divided down to a 5V signal level by a resistor network.

The six IGBT's (copack with built-in fly back diode) and high voltage gate drivers create a compact power stage. The drivers provide the level shifting that is required to drive high side bridge circuits commonly used in motor drives. PWM technique is used to the control motor phase voltage.

A Simple power supply (rectified mains) contains a Line and a RF filter in order to fulfil the European Community Regulation (IEC555-1).

As apparent, this concept uses a common ground (no isolation is used).

WARNING

It is strongly recommended to use opto-isolation (optocouplers and optoisolation amplifiers) during the development time to avoid any damage to the development equipment.

5 SOFTWARE

5.1 Control Algorithm

Two methods were tried. The first idea was to calculate the commutation events from the obtained zero crossing times. This method was shown to be not robust enough. Any disturbance to the feedback signal (due to noise or motor asymmetry) caused high torque ripples and motor stoppage.

Because of these problems, the second method was developed. Here, the motor is running as a synchronous motor. The operating value of the voltage is created in such a way, that the Back-EMF is aligned with the commutation. The phase Back-EMF zero crossing point is kept within a time window ($90° < \alpha < 180°$ see Figure 6-2.; this can be changed with respect to type of application). This principle is close to a simple vector controlm ethod. This does not require calculation of the next commutation event, directly based on the Back-EMF sensing. Therefore this is a more stable algorithm in case of feedback signal disturbance. No motor stoppage can occur. When using this algorithm the motor speed variation is very low.

The control flow consist of five phases: Alignment stage, Ramp-up stage, Stabilizing stage, PLL (Phase Locked Loop) Acquisition stage and Running (PLL Locked) stage.

5.1.1 Alignment

Before the motor starts, there is a short time (which depends on the motors electrical time constant) when two phases are supplied by power. The Current Controller keeps the current within predefined limits. This stage is necessary in order to create a high start-up torque (see (EQ 2-3.)).

5.1.2 Ramp-up

Here the motor is starting and ramped up (an "S" curve ramp creates a smooth transient) until it reaches the working speed. The Current Controller keeps current at the maximum limit in order to ensure that the rotor will not become locked. The Back-EMF sensing technique enables a sensorless detection of the rotor position, however the drive must be started without this feedback. It is caused by the fact that the amplitude of the induced voltage is proportional to the motor speed. Hence, the Back-EMF cannot be sensed at a very low speed and a special start-up algorithm must be performed.

5.1.3 Stabilizing

The motor is now running for a short time at constant speed. The motor speed is stabilized before the synchronization with the Back-EMF feedback takes place.

5.1.4 PLL Acquisition

The current controller is switched off. Only the over-current detection is left on. The motor is supplied from a voltage source. This transition must be done very carefully. The actual PWM duty cycle (phase voltage) is decreased until several zero crossing points can be sensed within the target time window ($90° < \alpha < 180°$ see Figure 5-2.). The rate of decrease must be tuned for the application and must not create oscillation of the whole drive.

5.1.5 Running (PLL Locked)

The following conditions have to be met to enter this stage:

1. Zero crossing events have to be sensed within a time window ($90° < \alpha < 180°$) several times (e.g. 6×).
2. Current peaks during one commutation period must be below a certain limit

Then the PLL controller keeps the phase shift α at the right value by controlling the phase voltage. The motor is now running with a good efficiency. Current measurement, over-current detection and motor stall detection is still continuously applied.

5.2 Main routines

The software is divided into several main routines: INIT, START, BODY and INTERRUPT SERVICE ROUTINES (ISR). Several common subroutines are called throughout these routines and are described in more details later in Section 5.2.4. The Figure 5-1. shows the sequence of the routines.

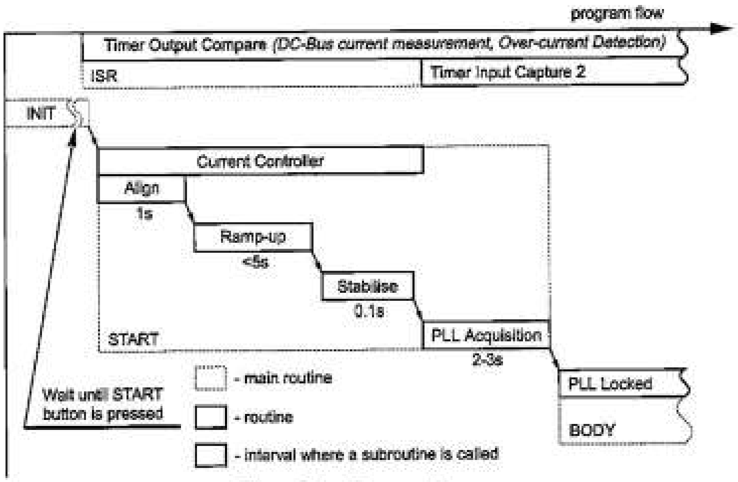


Figure 5-1. Program Flow

5.2.1 INIT

INIT does all necessary initialization before the drive is started.

Ports initialization:	PA1-6 - PWM output PB5-7 - output PC0-7 - inputs PD6-7 - inputs
A/D initialization:	The DC-Bus current zero value and DC-Bus voltage limits (high and low) are tested in order to prevent the drive from being damaged. The speed setting is sampled. After the START switch is turned on (see Figure 5-1.) the initialization continues.
PWM initialization:	Sets the PWM frequency (3.9kHz) and the PWM polarity (POLA=POLB=0). The time point for synchronization of the current measurement with the PWM is taken. The pointer for commutation look-up table is initialized.
Timer Output Compare Function initialization:	Software Timer1 is synchronized with PWM in order to handle the internal A/D converter. Software Timer2 (creates commutation period) is initialized.
Input Capture Function initialization:	The Input Capture Function and input pin (TCAP2) are initialized.
Current Controller initialization:	All registers used by PI type of Current Controller are cleared.

Finally, interrupts are enabled.

5.2.2 START

This routine performs the start algorithm and consists of the parts mentioned in the Control Algorithm Section.

5.2.2.1 Alignment

The required value of the current is set. The Current Controller keeps the current at the required value. The Current Controller subroutine is called every 512µs after new value of the DC-Bus current have been obtained. The PWM signals are applied onto only two motor phases (no commutation). The motor current and rotor position is stabilized.

5.2.2.2 Ramp-up

The commutation period is decreased in order to reach the maximals peed within 5 seconds (see Figure 5-4.).

The following subroutines are called: Current Controller (512µs period), Speed Setting & DC-Bus voltage sensing, Commutation and Ramp.

When finished ramping-up (the motor is running at the required speed), the program flow continues.

5.2.2.3 Stabilizing

Here, the motor speed is stabilized before the PLL Acquisition takes place. The motor is running with constant speed. After a certain time (which depends on the application), the program flow continues to the next stage.

The Current Controller (512µs period), the Speed Setting & DC-Bus voltage sensing and the Commutation subroutines are called from this part.

5.2.2.4 PLL Acquisition

The current peak during one commutation is detected, if it is higher than the preset limit (2.5A), then the PWM duty cycle is decreased. The Timer Input Capture interrupt is enabled ~20µs after

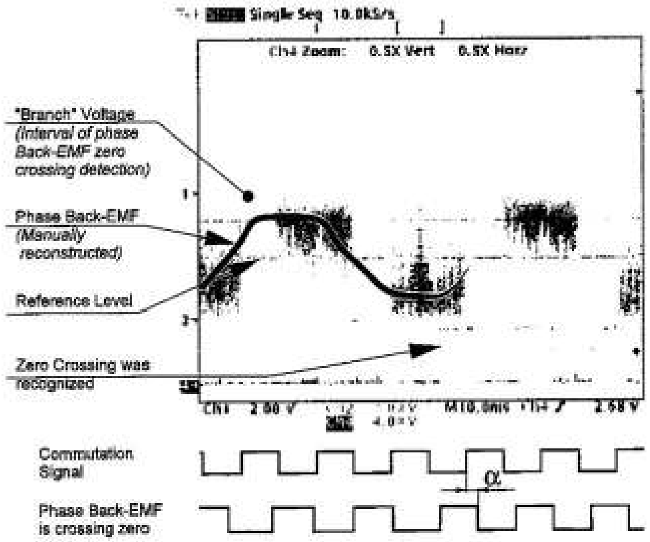


Figure 5-2. PLL Acquisition

the commutation, disabled before the successive commutation (OC ISR) or disabled when the edge of the Zero Crossing signal is detected by the IC function (see Figure 5-3.). If a zero crossing edge is detected, then the time interval from the commutation is calculated and condition the 90°<α<180° is evaluated. If this condition is met six times and the current peak is within the limit then the program flow continues. The following subroutines are called: Speed Setting & DC-Bus voltage sensing, Commutation.

The Figure 5-2. shows the moment when the zero crossing points can be sensed. The actual angle α is less than 90°, so the entry condition to the Running stage is not met and the DC-Bus voltage is decreased further.

5.2.3 BODY

This routine performs the Running stage. Speed Setting & DC-Bus voltage sensing and Commutation subroutines are called. The zero crossing feedback is handled and the angle α is evaluated. If no feedback is captured between commutations, then the zero crossing comparator feedback (TCAP2 pin) is sampled directly, just before the commutation takes place (see Figure 5-3., ⓪ samples).

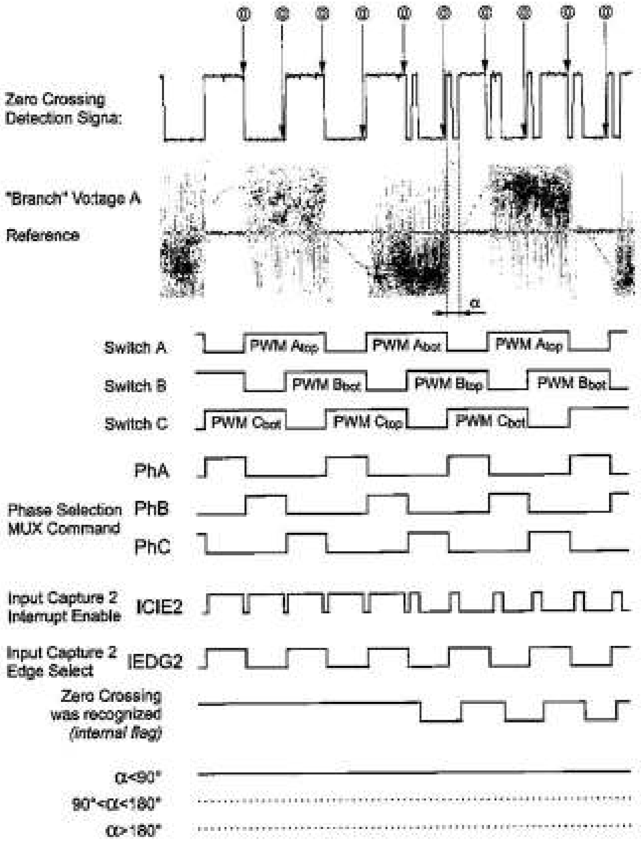


Figure 5-3. Control signals

The following formula is then used to evaluate if $\alpha<90^\circ$ or $\alpha>180^\circ$:

```
if (((Rising edge is expected) AND (TCAP2 pin is Hi)) OR ((Falling edge is expected) AND (TCAP2 pin is Lo))) then α<90°
if (((Rising edge is expected) AND (TCAP2 pin is Lo)) OR ((Falling edge is expected) AND (TCAP2 pin is Hi))) then α>180°
```

All three cases ($\alpha<90^\circ$, $90^\circ<\alpha<180^\circ$ and $\alpha>180^\circ$) are counted (by EMF_Lo, EMF_OK and EMF_Hi counters) during one revolution (18 commutations per revolution). The results are then used in the PLL Controller.

The stall detection is performed using another formula:

```
if (EMF_Lo>Stall limit)  then call Emergency Service routine.
```

5.2.3.1 PLL Controller

Based on above counters values the PLL Controller changes PWM duty cycle (PWMA and PWMB registers)

```
if (EMF_OK>6)  then stay with the same PWM duty cycle
               else if (EMF_Lo>EMF_Hi)  then decrease PWM duty cycle
                                        else increase PWM duty cycle
```

The PLL Controller can be more sophisticated for other applications (the time window can be divided into several subintervals etc.). In any case some kind of statistical evaluation is strongly recommended.

NOTE

If the current is higher, then the alignment of electrical and magnetic fields should be adjusted. The best α is no longer 90° but higher. The used algorithm does this improvement automatically.
Explanation: When load torque is increased, the phase current is also increased automatically and the fly-back diodes are open for a longer time. Thus the time window (when zero crossing can be sensed) is shrunk. This reduction can be so big that the time window becomes less than 30° long (also $\alpha>90^\circ$). If the previously explained control algorithm is applied, then the switching angle is increased so that the alignment of electrical and magnetic fields is improved (see Figure 2-3., Figure 2-4. and Figure 2-5.).

5.2.4 Subroutines

5.2.4.1 Emergency Stop Subroutine

This routines turns OFF all PWM outputs and waits. After some time the drive tries to start again (optional).

5.2.4.2 Speed Setting & DC-Bus Voltage Measurement

The result from an A/D conversion of either the Speed Setting or the DC-Bus voltage is stored into memory. The successive A/D channel is selected and an A/D conversion is started.

5.2.4.3 Current Controller Subroutine

The Current controller subroutine is called every 512µs. The PI controller calculates the PWM value based on the required and measured values of the DC-Bus current. The result from the PI controller is a scaled value of the duty cycle which is directly used by the PWM registers.

5.2.4.4 Commutation Subroutine

The values for the PWM control registers and the MUX command are found in look-up tables. The MUX command is a three bit word which controls the multiplexer (MUX). The multiplexer handles the phase zero crossing signals (see Figure 6-1.). The values obtained from the tables are stored into internal temporary variables in order to be ready when commutation is performed. The polarity edge sensitivity (IEDG2) of the Timer Input Capture 2 Function is toggled (see Figure 5-3.) in order to detect the rising or falling edge of the zero crossing signal. The edge polarity corresponds to the rotor position and the actual commutation stage.

5.2.4.5 Ramp Subroutine

The Ramp routine calculates commutation period (Timer1) such that an "S" speed ramp is achieved. No arithmetic division is required.

The formulas implemented are given below:

Acceleration: $$Period_{\{n+1\}} \leftarrow Period_{\{n\}} - \frac{(Period_{\{n\}} - PeriodStop) \cdot Const}{100hex} - 1 \qquad \text{(EQ 5-1.)}$$

Deceleration: $$Period_{\{n+1\}} \leftarrow Period_{\{n\}} + \frac{(Period_{\{n\}} - PeriodStart) \cdot Const}{100hex} + 1 \qquad \text{(EQ 5-2.)}$$

$Const$ is a parameter which changes the slope of the Ramp

$PeriodStart$ is the value of the commutation period when ramping is started.

$PeriodStop$ is the value of the commutation period when ramping is finished.

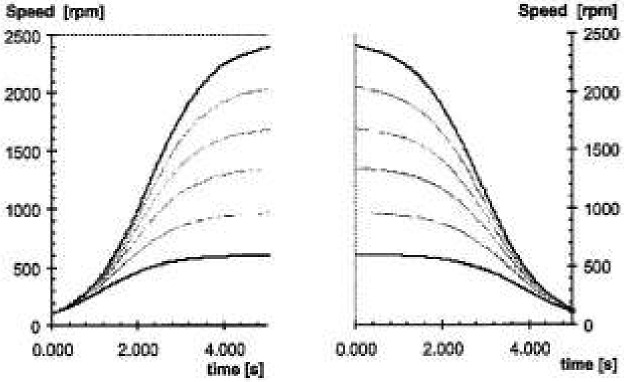


Figure 5-4. Ramp

5.2.5 ISR - Interrupt Services Routines

5.2.5.1 Timer Output Compare Interrupt Service Routine

Timer1 (A/D conversion) and Timer2 (commutation) are asynchronous software timers served by this routine.

If the OC Interrupt is caused by Timer1 the A/D conversion of DC-Bus current is done here first in order to be synchronized (within 4µs) with the PWM signal (see Figure 5-5.). The obtained value goes through the Over-current and Peak-current detection routines to make sure that it is within the limits. The flag for the Current Controller is set. This flag causes the Current Controller subroutine to be called with new data measured in this ISR.

On the other hand, if OC Interrupt is caused by Timer2, then the preset values (calculated in Commutation subroutine) are put into the PWM registers (CtlA, CtlB) and the MUX command is output too.

The IC Interrupt (ICIE2) is disabled here to preserve the Back-EMF detection from the disturbances produced during the commutation (see Figure 5-3.). It is enabled again after approximately 50µs.

If these two events are going to happen closer than 133µs, then Timer2 obtains priority and the commutation is performed at the right moment (no speed variation) while the A/D conversion is postponed by 256µs (see Figure 5-5.).

A new value for the Timer Output Compare Register (OCRH/L) is calculated, based on the states of Timer1 and Timer2.

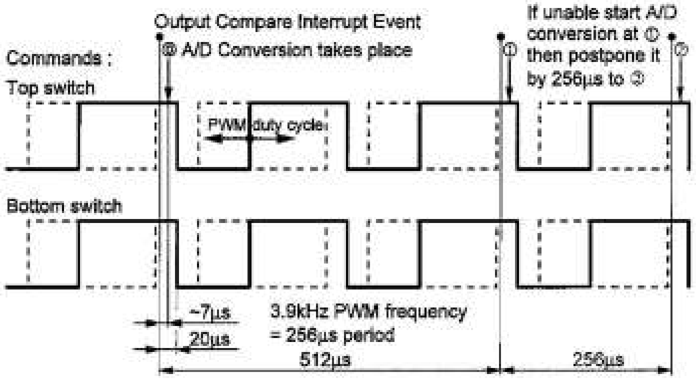


Figure 5-5. Measurement Synchronization with PWM

5.2.5.2 Timer Input Capture Interrupt Service Routine

The signal generated by the Position Recognition Logic causes this interrupt. Special care must be taken due to noise, which can disturb the incoming signal. The respective pin (TCAP2) is sampled in suitable places within the BODY and START routines. The period is less than 90µs and is typically 30µs. When this ISR is initiated, then another three samples of the TCAP2 are taken and the state of the TCAP2 pin is then evaluated. Based on the last sample before the ISR is executed, the selected edge sensitivity (IEDG2) and the samples after ISR is initiated, and the Input Capture event is verified. If it is acceptable then the captured time is stored in memory and the Input Capture Interrupt ISR is finished. The captured time is then used to calculate the angle α in the PLL.

5.2.5.3 IRQ

The wake up signal is serviced here. This allows the system to perform communication while the microcontroller is in the stand-by mode.

5.3 Program Load

The cycle time of the main routines was measured. The values obtained are given in the following table.

Routine	**Cycle time** (6MHz clock frequency)		
	Min.	**typical**	**Max.**
Start	22µs	~70µs	170µs
Body	25µs	~80µs	180µs
ISR-OC	55µs	84µs	90µs
ISR-IC	—	~25µs	—

Table 5-1.

5.4 Memory Map Usage

The figure below shows the overall memory usage for the system. It can be seen that very little ROM space is required by the system.

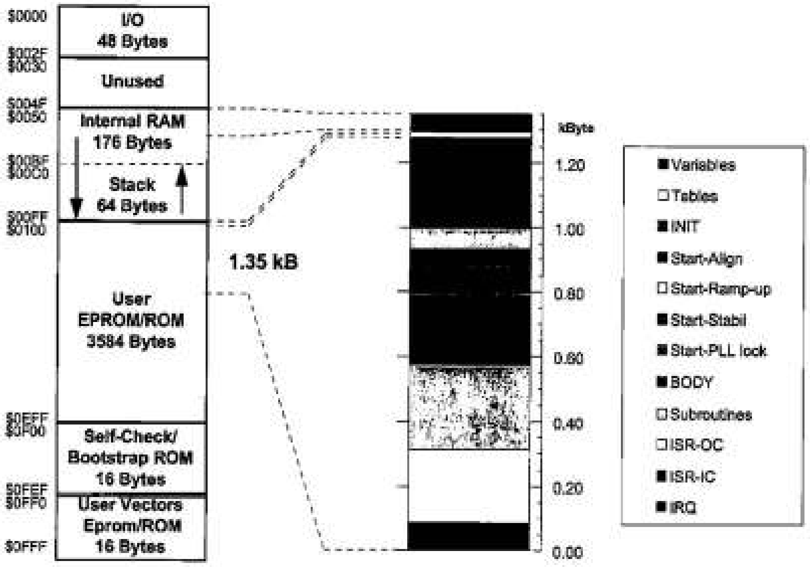


Figure 5-6. Memory Mapping

6 HARDWARE

6.1 Power Stage Board

6.1.1 Back-EMF sensing

An outline of the Back-EMF sensing circuit is shown below.

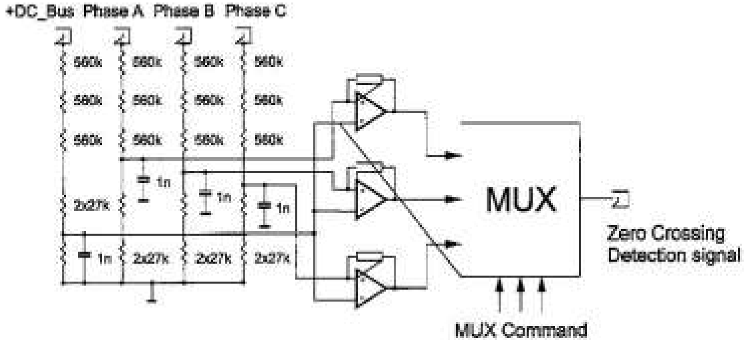


Figure 6-1. Back-EMF Sensing Circuit Diagram

As explained in the theoretical part of this application note, the phase zero crossing event can be detected at the moment when the branch voltage (of a free phase) crosses the half DC-bus voltage level. The resistor network is used to divide sensed voltages down to a 0-15V voltage level. The comparators sense the zero voltage difference of the input signal. The multiple resistors reduce the voltage across each resistor component to an acceptable level. A simple RC (54kΩ×1nF) filter prevents the comparators being disturbed by high voltage spikes produced by IGBT switching. The MUX selects the phase comparator output, which corresponds to the current commutation stage. This Zero Crossing Detection signal is transferred to the microcontroller's Input Capture pin (TCAP2).

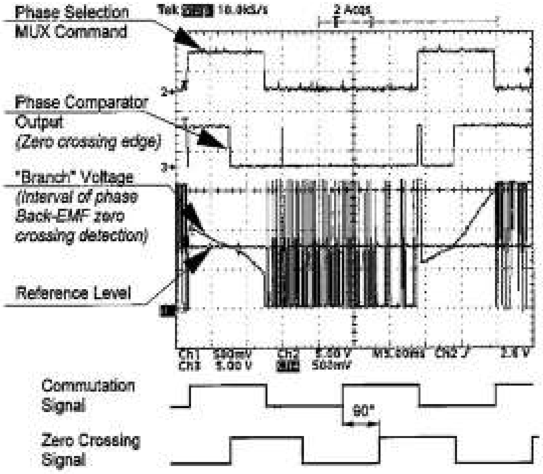


Figure 6-2. The Zero Crossing Detection

6.1.2 Current Sensing

In a star connected motor winding, the phase current can be sensed on the DC-Bus rail as each motor phase pair is energised. The voltage drop resistor (0,6Ohm/2W) is used to measure the DC-Bus current (0-4A), which is chopped by the PWM. The obtained signal is rectified and amplified (0-5V). The internal A/D converter is synchronized with the PWM signals (see Figure 5-5.). The current signal is converted 20µs before the end of the PWM "ON-time". This synchronization avoids spikes when the IGBT's are switching and simplifies the electric circuit. The over-current limit (3,6A) is detected by a comparator. This output is used to directly switch off the IGBT's and for the IRQ interrupt input.

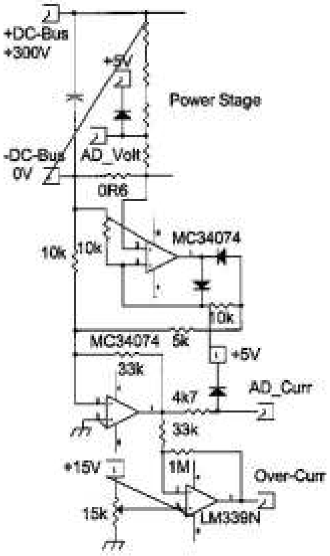


Figure 6-3. Current & Voltage Sensing

6.1.3 Voltage Sensing

The DC-Bus voltage is divided down-to a 5V signal level using a serial resistor network. This signal is used for over-voltage and under-voltage detection.

6.1.4 Power Devices

There are six MGP4N60ED IGBT's (copack; with built-in fly back diode) driven by three IR2112 high voltage gate drivers.

Switching times, saturation voltage, and short circuit time of the IGBT's have been optimized for motor drives. In addition, the gate threshold voltage is higher than usual for MOS gated power devices. The higher threshold eliminates the need for negative gate drive.

The high voltage driver IC's are designed to directly drive the gates of power MOS devices. They provide the level shifting that is required to drive high side bridge circuits commonly used in motor drives and other power applications. They are capable of withstanding floating supply offset voltages up to 600 volts. The Shut-down input of the IR2112 is used to switch off aill GBT's when over-current detection is activated.

Detailed information about the power stage board can be found in the application note: AN1590 - "High Voltage Medium Power Board for Three Phase Motors".

6.2 µC 68HC(7)05MC4 Board

The MC68HC(7)05MC4 is a low cost microcontroller especially dedicated for motor control. Due to this fact, the board schematic is very simple.

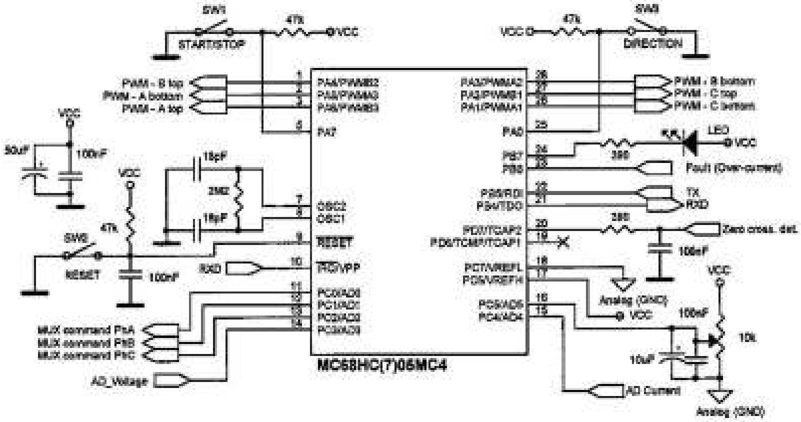


Figure 6-4. Microcontroller Board Schematic

There are no special requirements as long as common rules for microcontroller design are followed.

6.3 Power Supply Board

The rectified line voltage is filtered by a 100µF (400V) capacitor. The RFI filter and the Low-frequency (harmonic) filter are used to meet the new EMC regulation (IEC555-1).

Any type of power supply system (e.g. SMPS or classical centre-taped transformer) can be used to produce the auxiliary voltage levels (+15V, -15V and +5V).

This solution does have one limitation. The DC-Bus voltage can be 260V - 350V (the best case) which might not be sufficient for very large required speed range. The DC-Bus voltage should be three times higher (in star connection) than the phase Back-EMF voltage at maximal speed.

6.3.1 Meeting European Community Regulation IEC555-1

The application was designed to meet this regulation. The cheapest way often is to use one RF filter and one LF filter to improve the line current shape. The value of the inductance in the LF filter is a trade-off between power factor and current shape. It must be also kept in mind that any inductance before or after the mains rectifier decreases the DC-Bus voltage. This reduces the speed range this application can be used for.

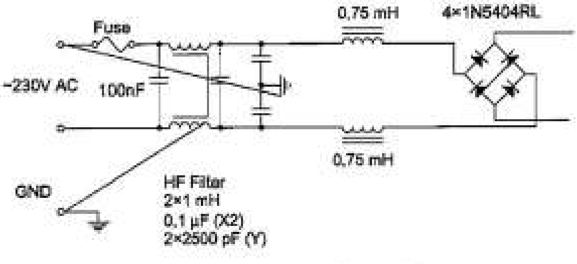
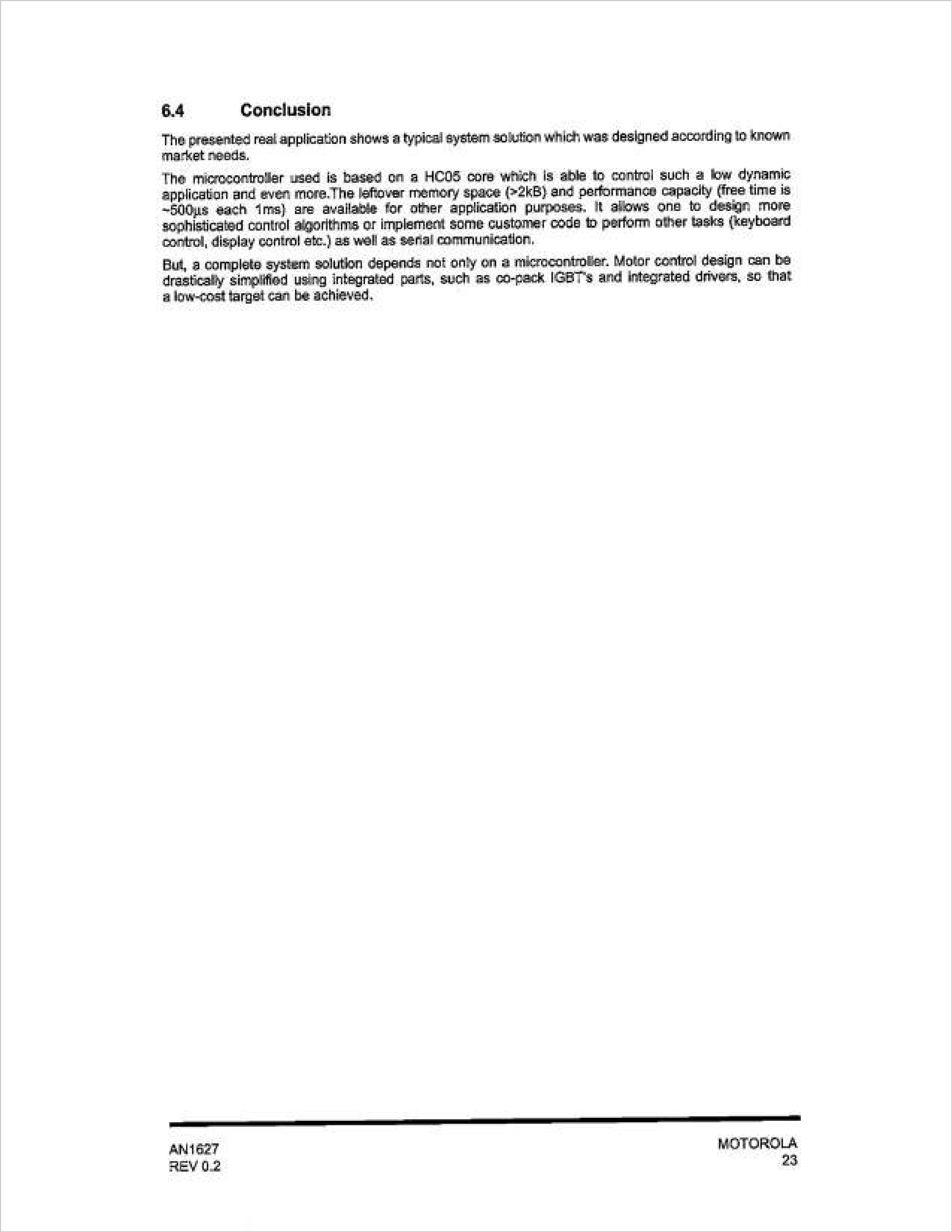


Figure 6-5. Main Power Supply Filters



6.4 Conclusion

The presented real application shows a typical system solution which was designed according to known market needs.

The microcontroller used is based on a HC05 core which is able to control such a low dynamic application and even more.The leftover memory space (>2kB) and performance capacity (free time is ~500µs each 1ms) are available for other application purposes. It allows one to design more sophisticated control algorithms or implement some customer code to perform other tasks (keyboard control, display control etc.) as well as serial communication.

But, a complete system solution depends not only on a microcontroller. Motor control design can be drastically simplified using integrated parts, such as co-pack IGBT's and integrated drivers, so that a low-cost target can be achieved.

APPENDIX A
MEASURED RESULTS

A.1 Power Stage Board

During the development it was important to find out if the power devices are suited for the application. Therefore a 30min test was performed. The temperature of the devices was checked to be within the range allowed by the manufacturer (105°C).

The gate resistance was 22Ohm. All IGBT's were in an upright position without heatsink and the temperature was measured on the centralp in. The following table summarizes the obtained results. The maximal temperature does not exceed the allowed limit and the devices passed the test.

Device: MGP4N60ED TO220 Copack			
Settings		**Results**	
Load torque	0.97Nm	Motor phase current	1.13A mean 1.36A RMS 2.14A peak
Motor speed	1002RPM		
DC-Bus voltage	305.5V	Temperature at the start of the test	30.8°C
PWM frequency	3.9kHz	Temperature at the end of the test	80.0°C
Ambient temperature	23°C	Max.temperature during 30min test	81.1°C

Table 1.

A typical oscillogram of the phase current (Ch1; 1A/10mV), the branch voltage (Ch3) and one half of the DC-Bus voltage (Ch2) waveforms (Ch2 and Ch3 are divided by resistor network) was taken during the test. In the points where the channel 3 is crossing the channel 2, the Zero Crossing points are

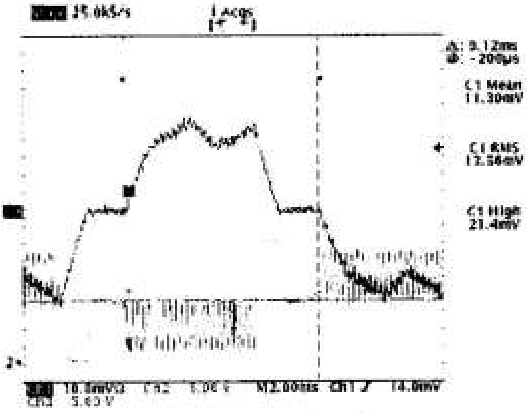


Figure 1. Measured waveforms

detected (see 2.2 Back-EMF Sensing). The alignment of the Back-EMF with the commutation events is good and the motor runs with high efficiency. The cursors define the time interval where the measurement of mean, RMS and peak values of the phase current are done. See results on the right side of the oscillogram.

A.1.1 Motor Load

The start-up capability of the drive was tested at more than 150% of the nominal load. The drive successfully reached the working speed (2400rpm) within 5s, as required. Figure 2. shows the obtained waveforms at the steady state condition.

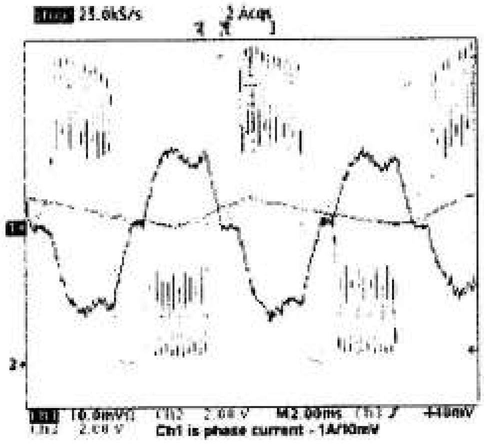


Figure 2.

Ch1 is the motor phase current, Ch2 is one half of the DC-Bus voltage and CH3 is the branch voltage Ch2 and Ch3 are divided by the resistor network down-to 0-15V level). The DC-Bus voltage is created by rectifying the mains and therefore channel 2 varies according to charging/discharging of the DC-Bus capacitor (10ms period). This variation has no impact on the Back-EMF zero crossing detection.

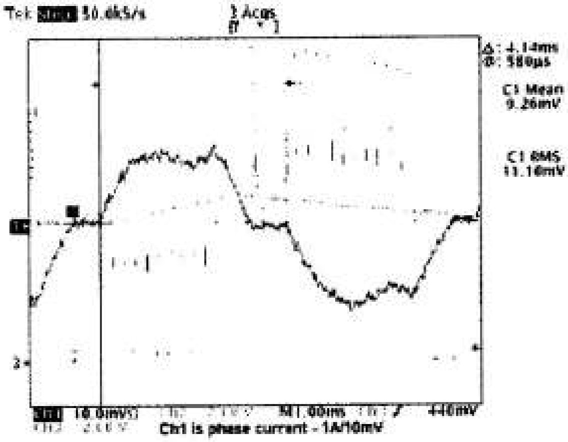
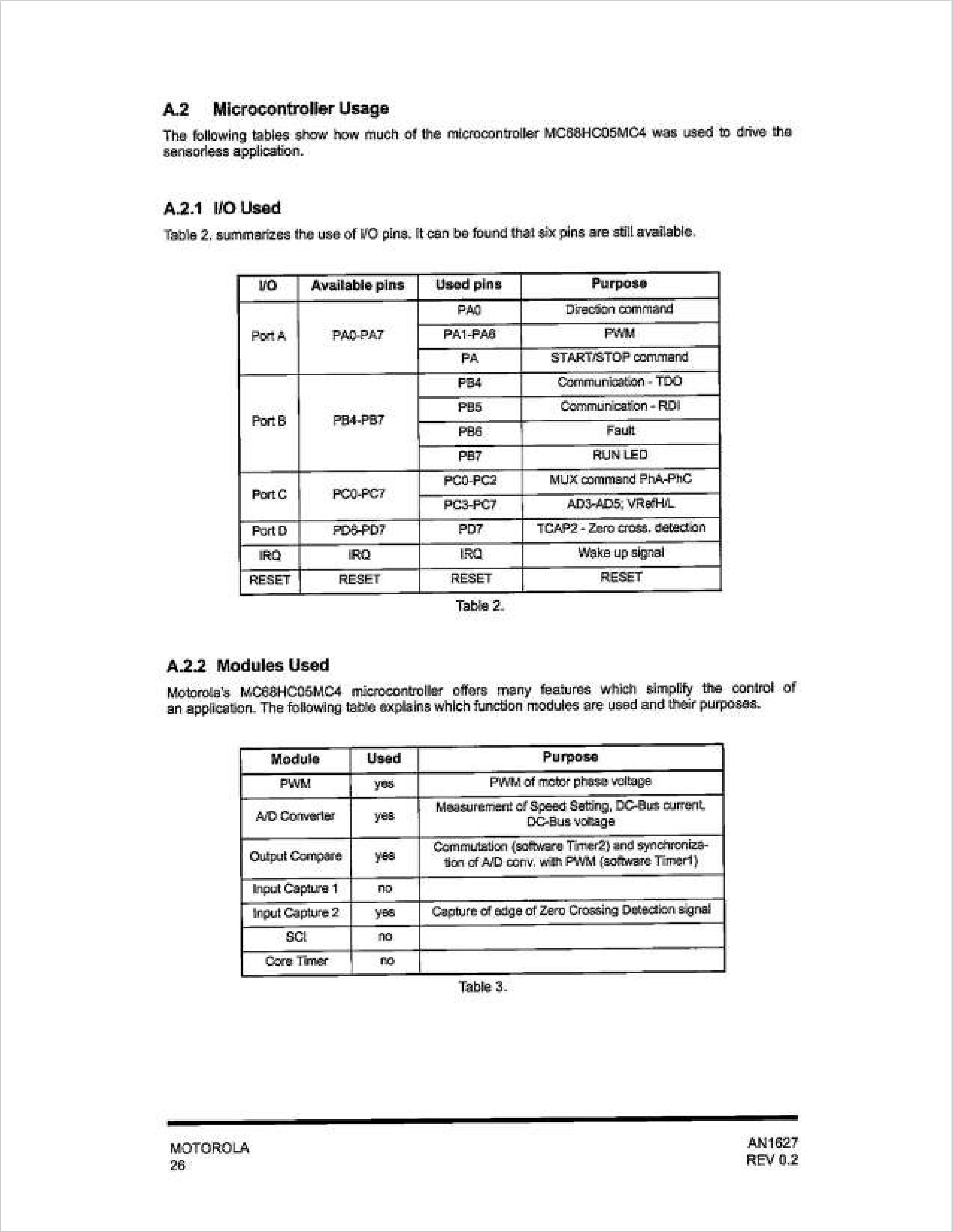


Figure 3.



A.2 Microcontroller Usage

The following tables show how much of the microcontroller MC68HC05MC4 was used to drive the sensorless application.

A.2.1 I/O Used

Table 2. summarizes the use of I/O pins. It can be found that six pins are still available.

I/O	Available pins	Used pins	Purpose
Port A	PA0-PA7	PA0	Direction command
		PA1-PA6	PWM
		PA	START/STOP command
Port B	PB4-PB7	PB4	Communication - TDO
		PB5	Communication - RDI
		PB6	Fault
		PB7	RUN LED
Port C	PC0-PC7	PC0-PC2	MUX command PhA-PhC
		PC3-PC7	AD3-AD5; VRefH/L
Port D	PD6-PD7	PD7	TCAP2 - Zero cross. detection
IRQ	IRQ	IRQ	Wake up signal
RESET	RESET	RESET	RESET

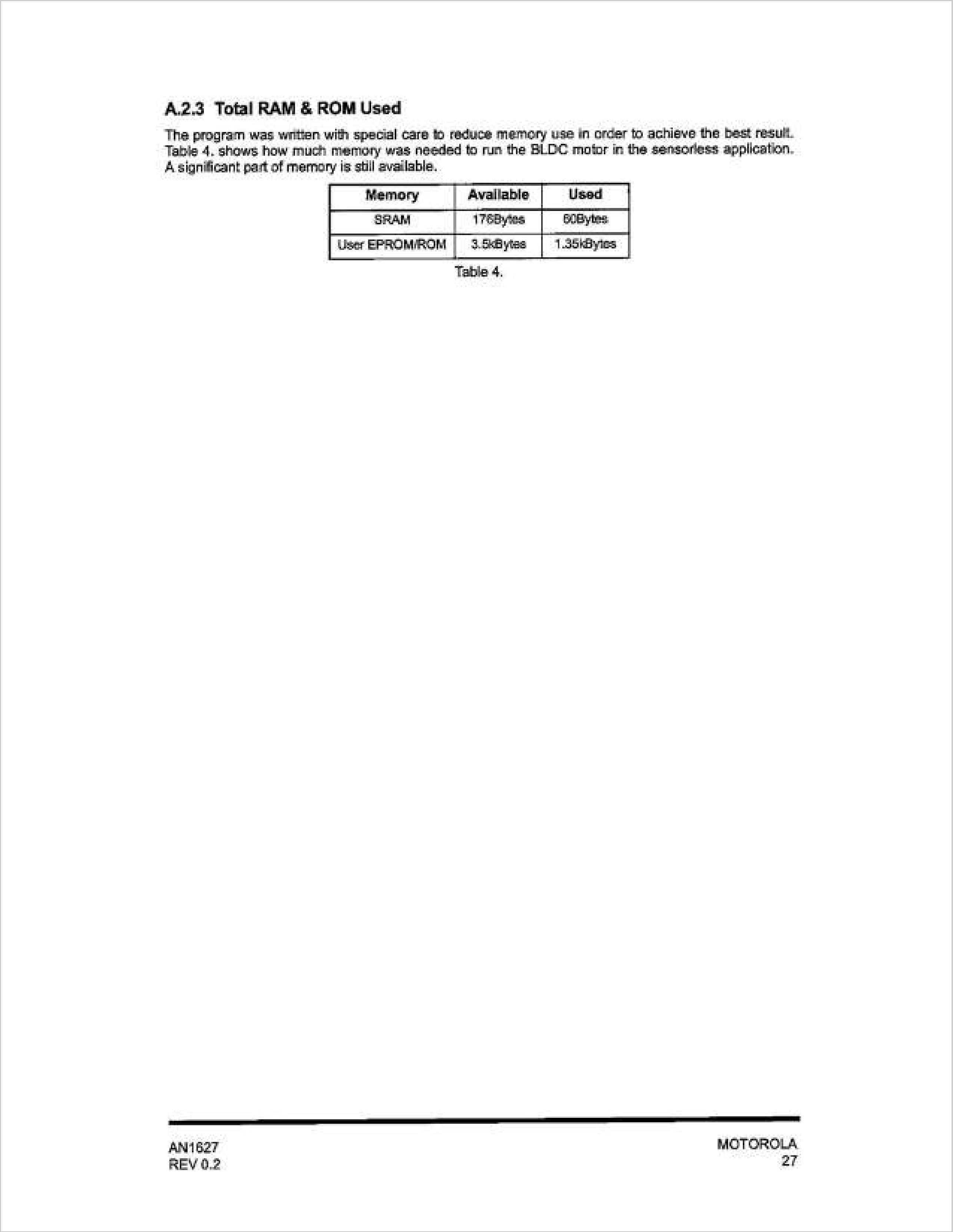
Table 2.

A.2.2 Modules Used

Motorola's MC68HC05MC4 microcontroller offers many features which simplify the control of an application. The following table explains which function modules are used and their purposes.

Module	Used	Purpose
PWM	yes	PWM of motor phase voltage
A/D Converter	yes	Measurement of Speed Setting, DC-Bus current, DC-Bus voltage
Output Compare	yes	Commutation (software Timer2) and synchronization of A/D conv. with PWM (software Timer1)
Input Capture 1	no	
Input Capture 2	yes	Capture of edge of Zero Crossing Detection signal
SCI	no	
Core Timer	no	

Table 3.



A.2.3 Total RAM & ROM Used

The program was written with special care to reduce memory use in order to achieve the best result. Table 4. shows how much memory was needed to run the BLDC motor in the sensorless application. A significant part of memory is still available.

Memory	Available	Used
SRAM	176Bytes	60Bytes
User EPROM/ROM	3.5kBytes	1.35kBytes

Table 4.

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555 Timer

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What is the 555 timer?

The 555 timer is one of the most remarkable integrated circuits ever developed. It comes in a single or dual package and even low power cmos versions exist - ICM7555. Common part numbers are LM555, NE555, LM556, NE556. The 555 timer consists of two voltage comparators, a bi-stable flip flop, a discharge transistor, and a resistor divider network.

I am particularly indebted to **Philips Components and Semiconductors Australia** for their most generous assistance in giving me access to material presented on this page.

Philips describe their 555 monlithic timing circuit as a "highly stable controller capable of producing accurate time delays, or oscillation. In the time delay mode of operation, the time is precisely controlled by one external resistor and capacitor. For a stable operation as an oscillator, the free running frequency and the duty cycle are both accurately controlled with two external resistors and one capacitor. The circuit may be triggered and reset on falling waveforms, and the output structure can source or sink up to 200mA."

What are the 555 timer applications?

Applications include precision timing, pulse generation, sequential timing, time delay generation and pulse width modulation (PWM).

Pin configurations of the 555 timer

Here are the pin configurations of the 555 timer in figure 1 below.

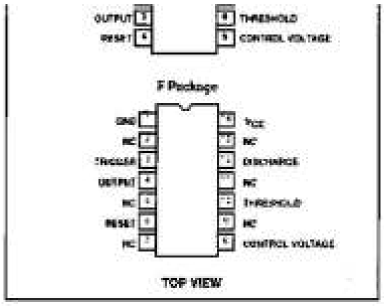


Figure 1 - 555 timer pin configurations

Pin Functions - 8 pin package

Ground (Pin 1)

Not surprising this pin is connected directly to ground.

Trigger (Pin 2)

This pin is the input to the lower comparator and is used to set the latch, which in turn causes the output to go high.

Output (Pin 3)

Output high is about 1.7V less than supply. Output high is capable of I_{source} up to 200mA while output low is capable of I_{sink} up to 200mA.

Reset (Pin 4)

This is used to reset the latch and return the output to a low state. The reset is an overriding function. When not used connect to V+.

Control (Pin 5)

Allows access to the 2/3V+ voltage divider point when the 555 timer is used in voltage control mode. When not used connect to ground through a 0.01 uF capacitor.

Threshold (Pin 6)

This is an input to the upper comparator. See data sheet for comprehensive explanation.

Discharge (Pin 7)

This is the open collector to Q14 in figure 4 below. See data sheet for comprehensive explanation.

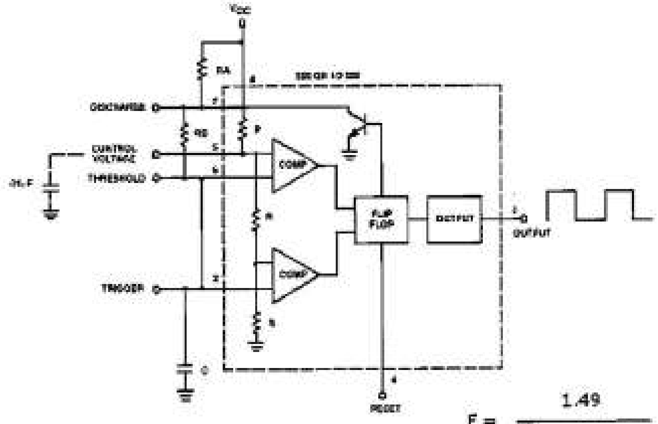
V+ (Pin 8)

This connects to Vcc and the Philips databook states the ICM7555 cmos version operates 3V - 16V DC while the NE555 version is 3V - 16V DC. Note comments about effective supply filtering and bypassing this pin below under "General considerations with using a 555 timer"

555 timer in astable operation

When configured as an oscillator the 555 timer is configured as in figure 2 below. This is the free running mode and the trigger is tied to the threshold pin. At power-up, the capacitor is discharged, holding the trigger low. This triggers the timer, which establishes the capacitor charge path through Ra and Rb. When the capacitor reaches the threshold level of 2/3 Vcc, the output drops low and the discharge transistor turns on.

The timing capacitor now discharges through Rb. When the capacitor voltage drops to 1/3 Vcc, the trigger comparator trips, automatically retriggering the timer, creating an oscillator whose frequency is determined by the formula in figure 2.

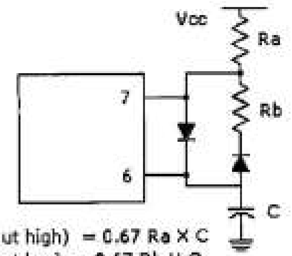


Astable Operation

$$F = \frac{1.49}{(Ra + 2Rb) * C}$$

Figure 2 - 555 timer in astable operation

There are difficulties with duty cycle here and I will deal with them below. It should also be noted that a minimum value of 3K should be used for Rb.



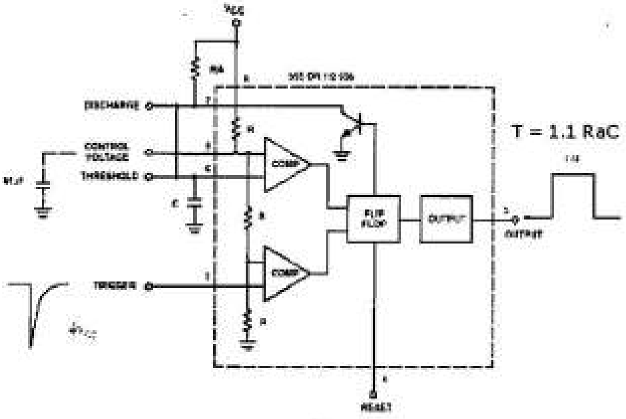
t1 (output high) = 0.67 Ra X C
t1 (output low) = 0.67 Rb X C
T = t1 + t2 (total period) $F = \frac{1}{T}$

Figure 2a - modified duty cycle in astable operation

Here two signal diodes (1N914 types) have been added. This circuit is best used at Vcc = 15V.

555 timer in monostable operation

Another popular application for the 555 timer is the monostable mode (one shot) which requires only two external components, Ra and C in figure 3 below. Time period is determined by 1.1 X Ra C.



Monostable Operation

Figure 3 - 555 timer in monostable operation

General considerations with using a 555 timer

Most devices will operate down to as low as 3V DC supply voltage. However correct supply filtering and bypassing is critical, a capacitor between .01 uF to 10 uF (depending upon the application) should be placed as close as possible to the 555 timer supply pin. Owing to internal design considerations the 555 timer can generate large current spikes on the supply line.

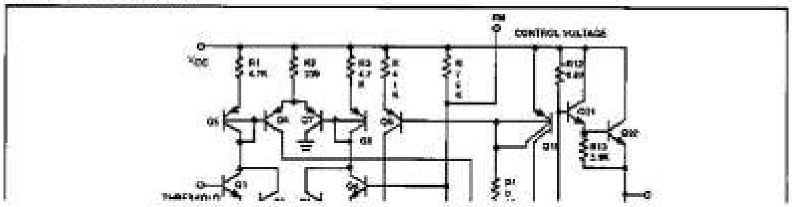
While the 555 timer will operate up to about 1 Mhz it is generally recommended it not be used beyond 500 Khz owing to temperature stability considerations.

Owing to low leakage capacitor considerations limit maximum timing periods to no more than 30 minutes.

External components when using a 555 timer

Care should be taken in selecting stable resistors and capacitors for timing components in the 555 timer. Also the data sheet should be consulted to determine maximum and minimum component values which will affect accuracy. Capacitors must be low leakage types with very low *Dielectric Absorption* properties. Electrolytics and Ceramics are not especially suited to precision timing applications.

EQUIVALENT SCHEMATIC



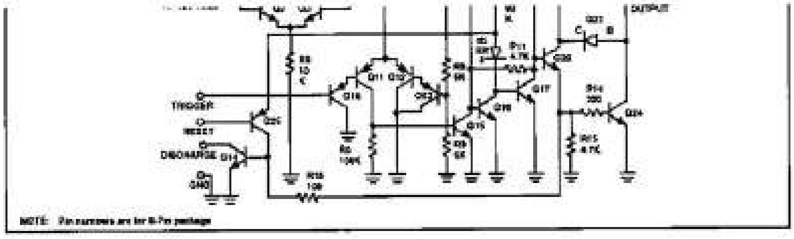


Figure 4 - equivalent schematic of a 555 timer

Very low frequency timing

There have been a few people who have written to me about problems in using the 555 timer in very low frequency applications. One particular case was at 10 Hz:

"I made an oscillator using 555 IC running in astable mode, at 10.0HZ. My problem is extreme over heating and ultimate burn out. Can you give a stable circuit as 555 to run at 10HZ without over heating and extremly stable"

My advice was to consider using another device such as the **74HC4060** "14-stage binary ripple counter with oscillator". You can use more than one device in series (cascading) for higher division numbers (with some often inconvenient gaps).

The 74HC/HCT4060 are 14-stage ripple-carry counter/dividers and oscillators with three oscillator terminals (RS, RTC and CTC), ten buffered outputs (Q3 to Q9 and Q11 to Q13) and an overriding asynchronous master reset (MR).

The oscillator configuration allows design of either RC or crystal oscillator circuits. The oscillator may be replaced by an external clock signal at input RS. In this case keep the other oscillator pins (RTC and CTC) floating.

One device is capable of dividing by Q13 (that's Philips notation for 14 as they consider Q0 as a valid first number division). If you don't understand binary division it may pay to look at my page on **digital basics**. You can cascade as many devices as you like although you will get some gaps in divisions available to you - see data sheet.

To achieve a highly accurate 10 Hz output I'd suggest doing this:

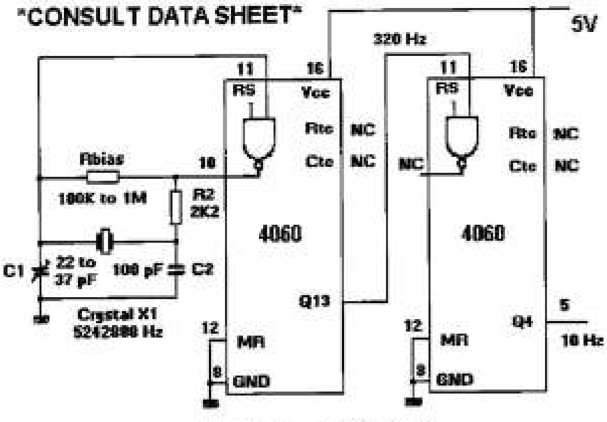


Figure 5 - 74HC/HCT4060 14-stage ripple-carry counter/dividers used to achieve 10 Hz output

Now if I've got it right and, it's been a long while since I did this, that schematic should work.

I've selected the Q13 output of the first device (which is actually the 14th stage) and this gives us 5242880 Hz divided by 16384 = 320 Hz. This is fed into the RS input of the second device where the output is taken from the Q4 stage which is 320 Hz divided by 32 = 10 Hz.

To some people the confusing part is Philips use the system of Q0, Q1, Q2.., Q12 and Q13. So Q13 is actually the **14th stage** and Q4 is actually the **5th stage**. Adding the two together, 14 + 5 = 19 (oh duh), means if we successively divided 5242880 by 2 for nineteen times we will get down to 10 Hz.

Several points to bear in mind:

- maximum clock frequency is a very generous 80+ Mhz
- you can cascade as many devices as you like
- the accuracy of your output frequency is entirely the same as your source
- this only one example - use your imagination

It's an interesting device to play around with. I've used it to play around with hee-haw sirens and other amazing yet, foolish projects.

Don't forget to consult the data sheet, print it out and look at my page on **digital basics**.

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elbow 1
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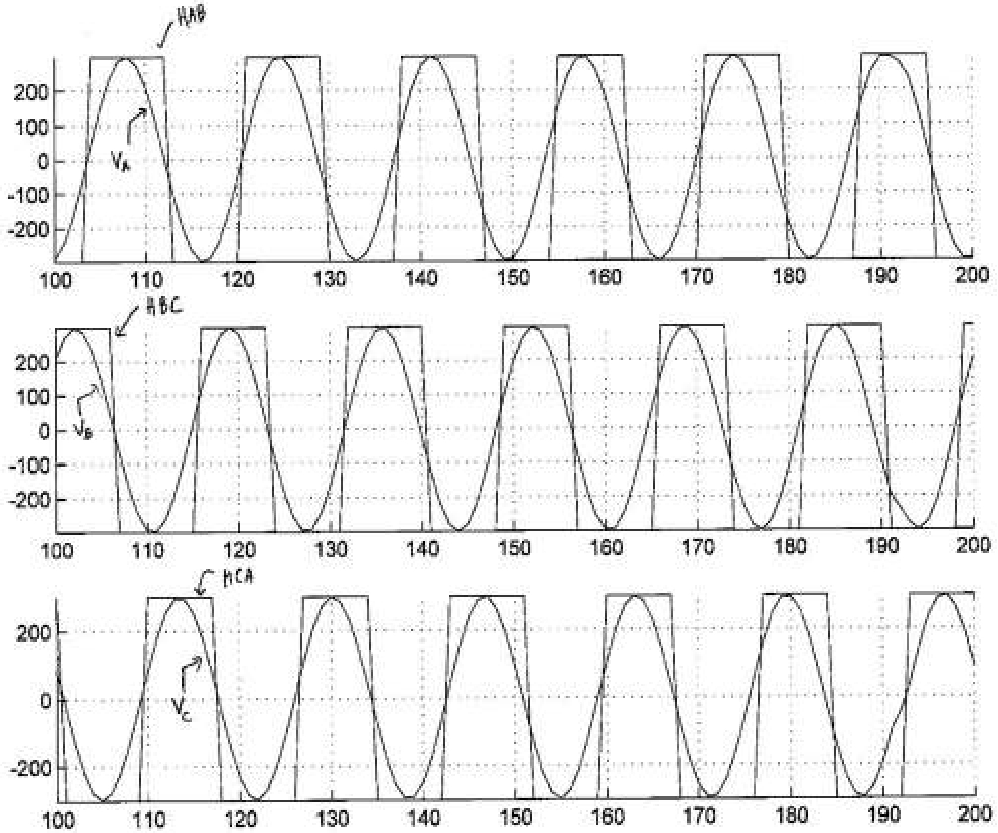
<u>CCW</u>

(V_A vs t)
(HAB vs t)



Brushless Motor Commutation

Sinusoidal :

$$u = \tau \sin(\theta + \text{offset})$$

$$v = \tau \sin(\theta \underline{- 120°} + \text{offset})$$

$\hookrightarrow$ not +120° ?

I looked at the plot of (u vs. HAB) and (v vs. HBC) and found that the phase shift between u and v needs to be (−120°) and NOT (+120°) !!

"offset" → I found that as time goes by, the u signal starts to lag the HAB signal causing the motor to stop eventually.

Fortunately this "offset" appears to be linear. This is calculated by looking at the graph of (position vs time) and noticing the point where velocity starts to drop. Get the phase diff between u and HAB at this time and also the time (or tick)

$$\text{phase offset}/\text{tick} = \frac{(\text{phase diff in degrees})}{\#\ \text{ticks}}$$

for example, if at t = 40000, u now lags by 45°

$$\text{offset} = \frac{45°}{40000} = 0.0011\ °/\text{tick}$$

the sign is negative (or positive ?? trial and error this)

Sinosoidal Com

C.W. direction ↦ "offset" starts at 0° ; offset = 0.0018 °/tick

C.C.W. direction ↦ "offset" starts at 30° ; offset = 30° − 0.0011 °/tick

↳ trial and error determination

I determined this by first setting "offset" to zero and the do sinosoidal going C.C.W. Then look at the plot of (u and HAB vs t) and check the phase shift. Ideally the should be in phase

---

to shift ⟵| add phase shift

to shift |⟶ substract phase shift

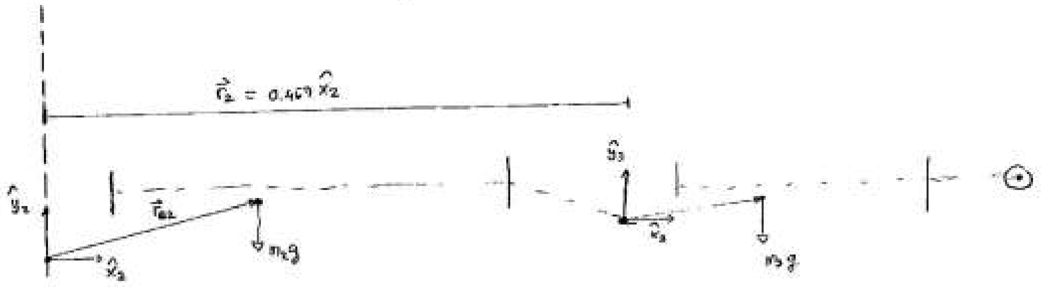
---

rpm calculation $= \underbrace{\left(\text{slope of } \overset{\text{angle}}{\text{position}} \text{ vs. tick}\right)}_{\text{degrees / tick}} \; 1000 \frac{\text{ticks}}{\text{second}} \; 60 \frac{\text{seconds}}{\text{minute}} \; \frac{1}{360} \frac{\text{revolution}}{\text{degrees}}$

RBE 1511B → EL 3

RBE 1513B → SH3, WA1

$\vec{r}_2 = 0.469\,\hat{x}_2$



$0.168\,\hat{x}_2 + 0.025\,\hat{y}_2$

$0.084\,\hat{x}_3 + 0.007\,\hat{y}_3$

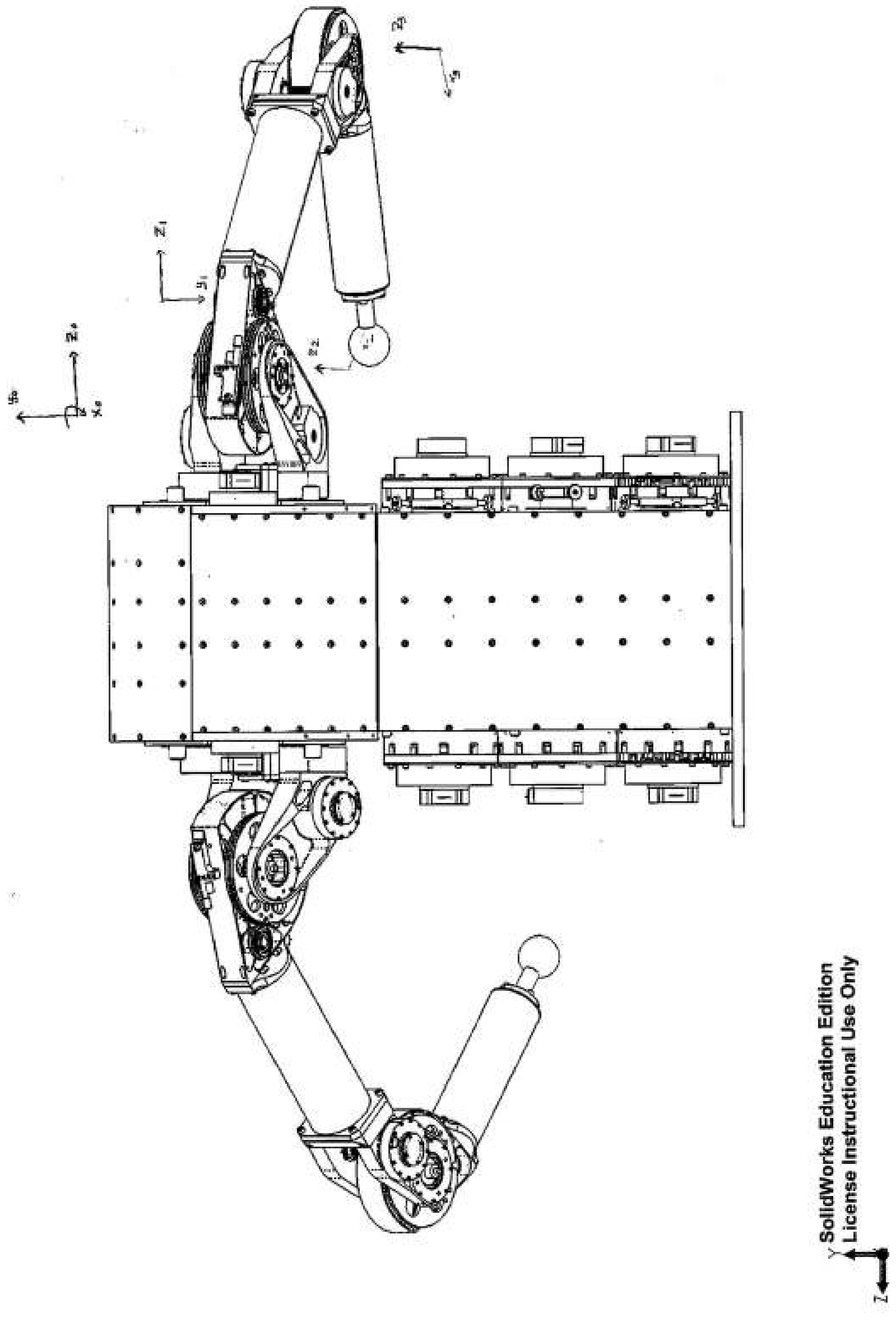
$\vec{F}_2 = -m_2 g\,\hat{y}_2 = -31.3\,\hat{y}_2$

$\vec{F}_3 = -m_3 g\,\hat{y}_3 = -14.18\,\hat{y}_3$

$\tau_2 = \vec{r}_{E2} \times \vec{F}_2 + \left(\vec{r}_{c3} + \vec{r}_2\right) \times \vec{F}_3$

$\tau_2 = (0.168)(-31.3) + (0.457 + 0.084)(-14.18)$

$\tau_2 = -12.93$ N.m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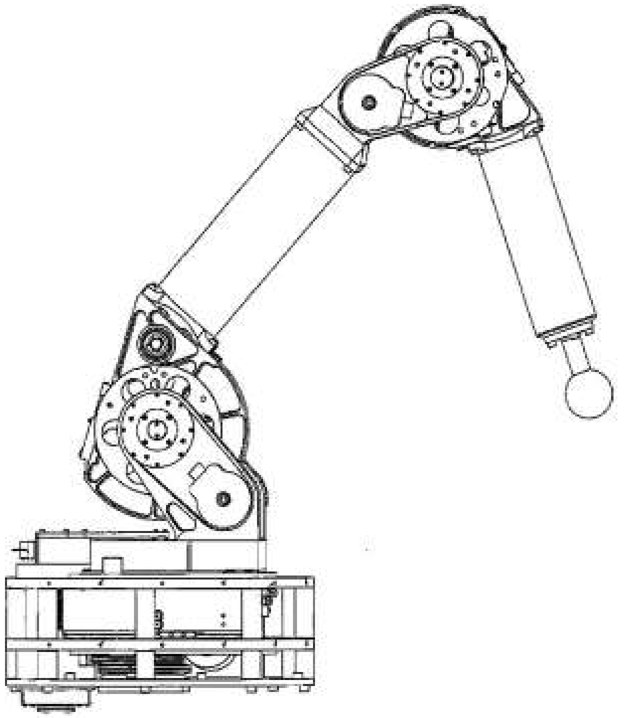


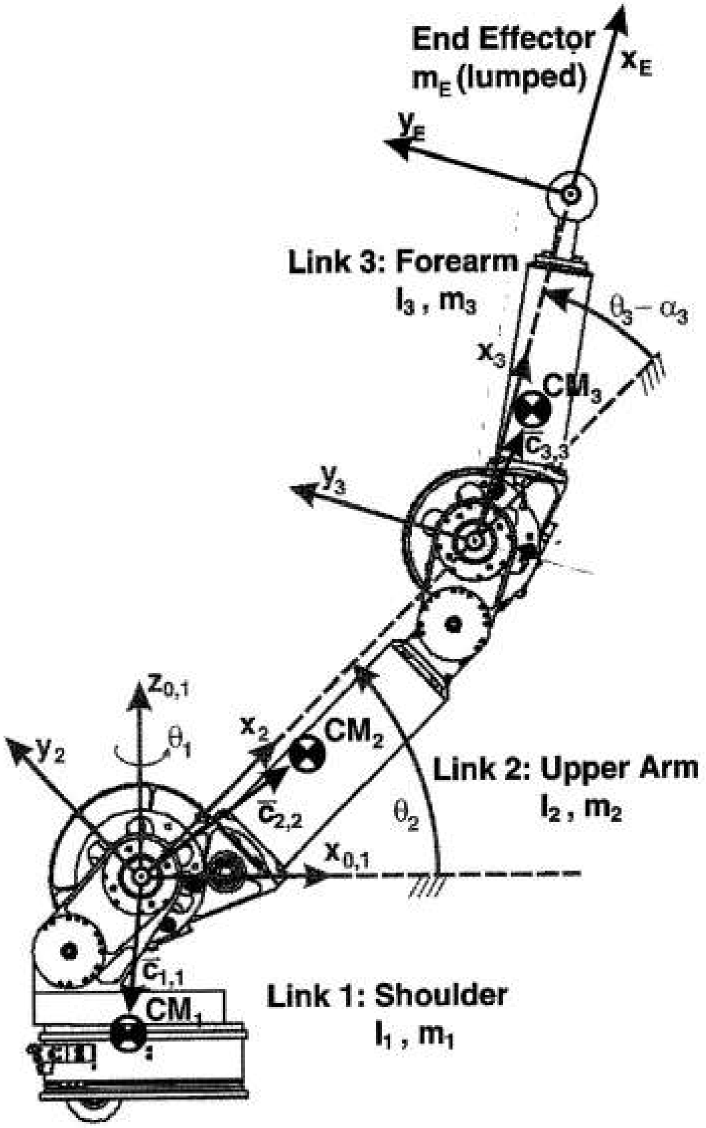


Figure 4.1: Coordinate systems for the analysis of the Stanford 3-DOF HFR. The coordinate systems are aligned following the conventions for the use of Denavit-Hartenberg parameters.

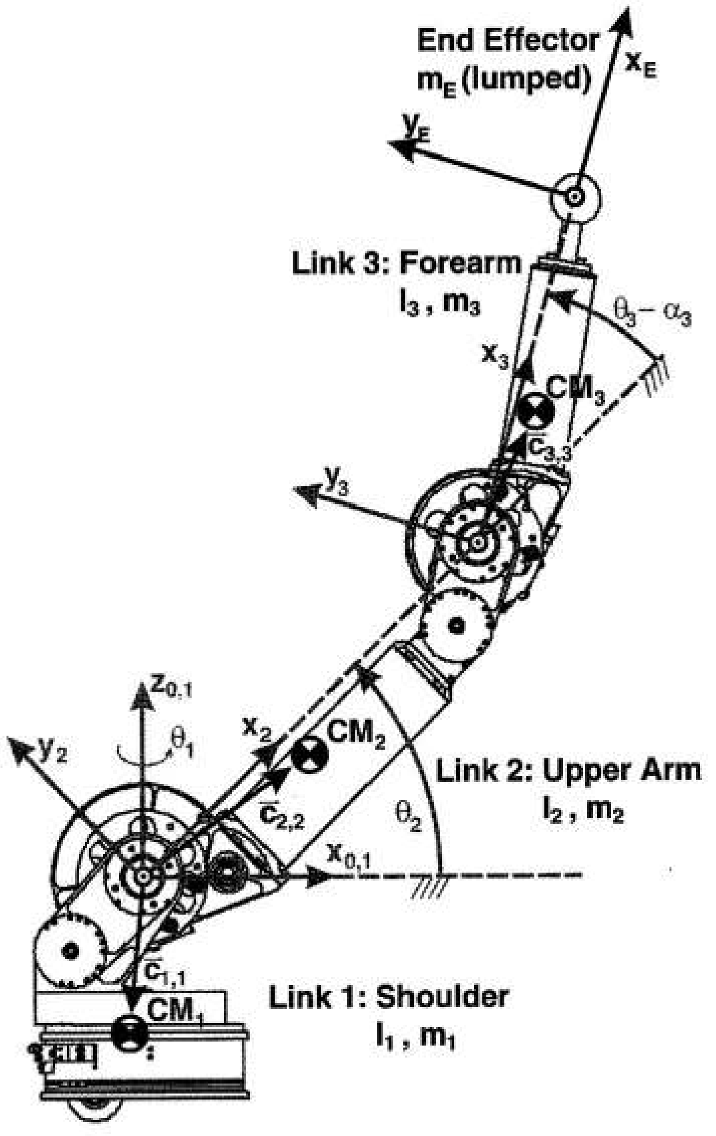


Figure 4.1: Coordinate systems for the analysis of the Stanford 3-DOF HFR. The coordinate systems are aligned following the conventions for the use of Denavit-Hartenberg parameters.

Canon custom shielded cable

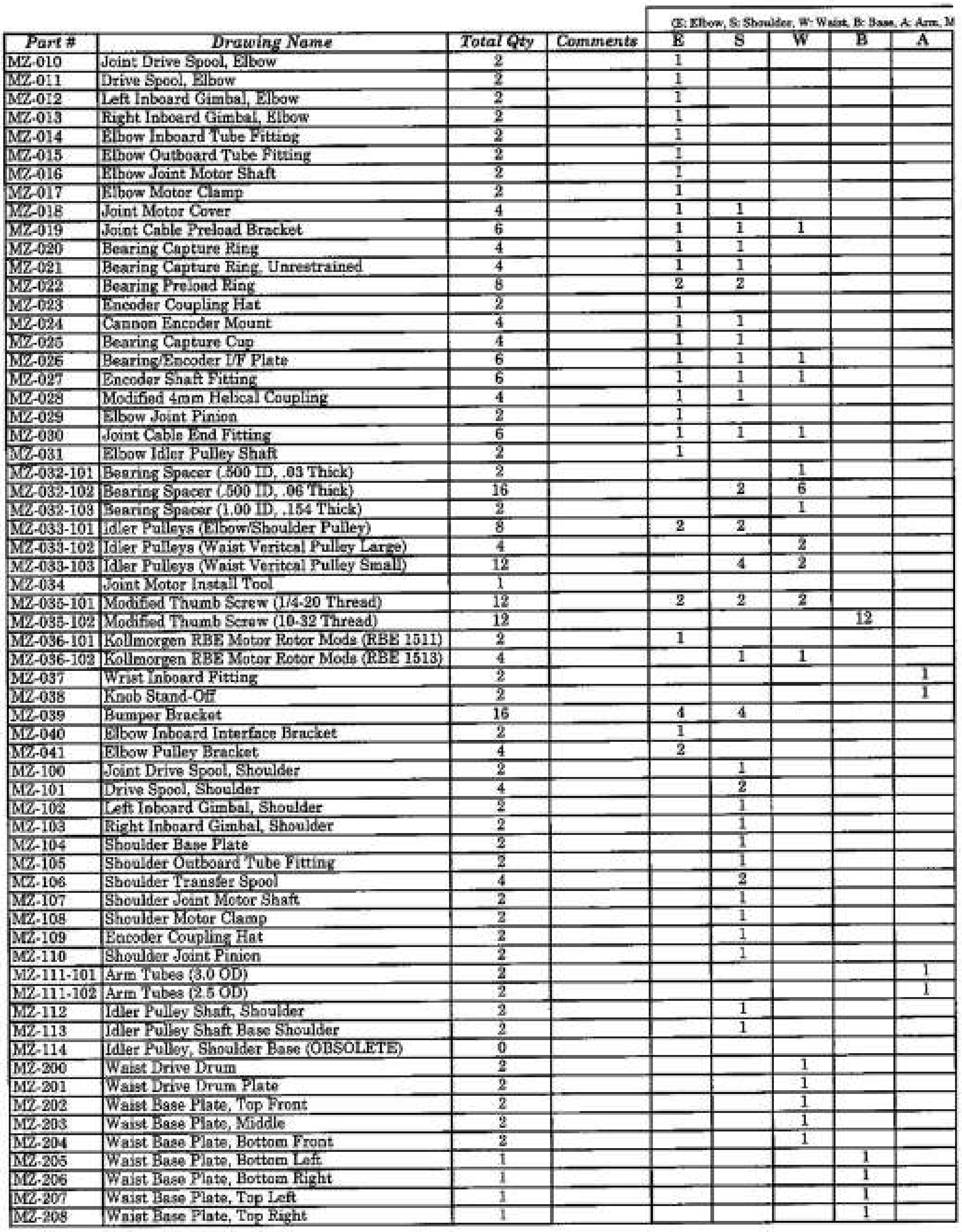
alpha color	function	cannon pin	stg pin	ribbon color
red	+5V	7		
white	-5V	9		
black	GND	8		
green	shield	12		
orange	GND	10		
blue	A+	1	6	cream
brown	A-	2	5	orange
yellow	B+	3	8	cream
gray	B-	4	7	yellow
pink	Z+	5	10	cream
sand	Z-	6	9	green

Cannon Hirose Connector Pinout

cannon pin	function
1	A+
2	A-
3	B+
4	B-
5	Z+
6	Z-
7	+5V
8	GND
9	-5V
10	GND
11	GND
12	shield

Copley Brushless connections

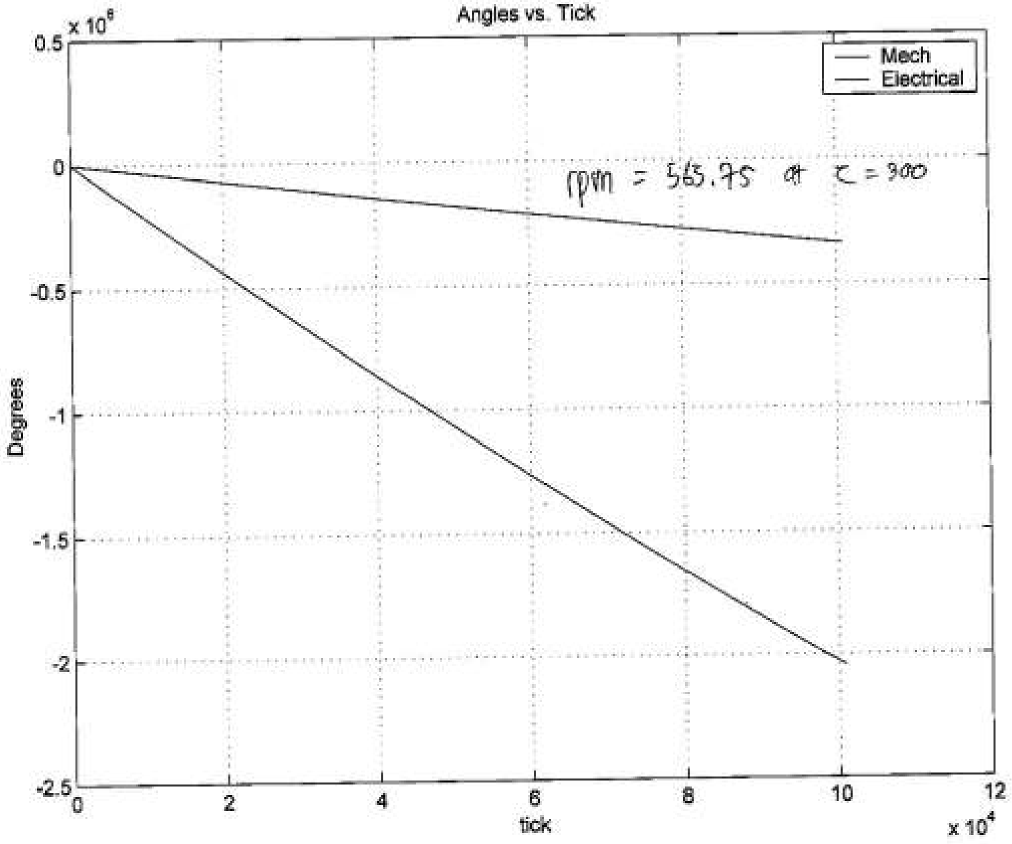
#2	#3			IDC 10	
DAC2	u	u + (11)	→	(3)	orange
		u - (12)	→	(4)	yella
DAC3	v	v + (11)	→	(5)	green
		v - (11)	→	(6)	blue
		GND (1)	→	(1) (2)	brown red.



				(E: Elbow, S: Shoulder, W: Waist, B: Base, A: Arm, M				
Part #	***Drawing Name***	***Total Qty***	***Comments***	**E**	**S**	**W**	**B**	**A**
MZ-010	Joint Drive Spool, Elbow	2		1				
MZ-011	Drive Spool, Elbow	2		1				
MZ-012	Left Inboard Gimbal, Elbow	2		1				
MZ-013	Right Inboard Gimbal, Elbow	2		1				
MZ-014	Elbow Inboard Tube Fitting	2		1				
MZ-015	Elbow Outboard Tube Fitting	2		1				
MZ-016	Elbow Joint Motor Shaft	2		1				
MZ-017	Elbow Motor Clamp	2		1				
MZ-018	Joint Motor Cover	4		1	1			
MZ-019	Joint Cable Preload Bracket	6		1	1	1		
MZ-020	Bearing Capture Ring	4		1	1			
MZ-021	Bearing Capture Ring, Unrestrained	4		1	1			
MZ-022	Bearing Preload Ring	8		2	2			
MZ-023	Encoder Coupling Hat	2		1				
MZ-024	Cannon Encoder Mount	4		1	1			
MZ-025	Bearing Capture Cup	4		1	1			
MZ-026	Bearing/Encoder I/F Plate	6		1	1	1		
MZ-027	Encoder Shaft Fitting	6		1	1	1		
MZ-028	Modified 4mm Helical Coupling	4		1	1			
MZ-029	Elbow Joint Pinion	2		1				
MZ-030	Joint Cable End Fitting	6		1	1	1		
MZ-031	Elbow Idler Pulley Shaft	2		1				
MZ-032-101	Bearing Spacer (.500 ID, .03 Thick)	2				1		
MZ-032-102	Bearing Spacer (.500 ID, .06 Thick)	16			2	6		
MZ-032-103	Bearing Spacer (1.00 ID, .154 Thick)	2				1		
MZ-033-101	Idler Pulleys (Elbow/Shoulder Pulley)	8		2	2			
MZ-033-102	Idler Pulleys (Waist Veritcal Pulley Large)	4				2		
MZ-033-103	Idler Pulleys (Waist Veritcal Pulley Small)	12			4	2		
MZ-034	Joint Motor Install Tool	1						
MZ-035-101	Modified Thumb Screw (1/4-20 Thread)	12		2	2	2		
MZ-035-102	Modified Thumb Screw (10-32 Thread)	12					12	
MZ-036-101	Kollmorgen RBE Motor Rotor Mods (RBE 1511)	2		1				
MZ-036-102	Kollmorgen RBE Motor Rotor Mods (RBE 1513)	4			1	1		
MZ-037	Wrist Inboard Fitting	2						1
MZ-038	Knob Stand-Off	2						1
MZ-039	Bumper Bracket	16		4	4			
MZ-040	Elbow Inboard Interface Bracket	2		1				
MZ-041	Elbow Pulley Bracket	4		2				
MZ-100	Joint Drive Spool, Shoulder	2			1			
MZ-101	Drive Spool, Shoulder	4			2			
MZ-102	Left Inboard Gimbal, Shoulder	2			1			
MZ-103	Right Inboard Gimbal, Shoulder	2			1			
MZ-104	Shoulder Base Plate	2			1			
MZ-105	Shoulder Outboard Tube Fitting	2			1			
MZ-106	Shoulder Transfer Spool	4			2			
MZ-107	Shoulder Joint Motor Shaft	2			1			
MZ-108	Shoulder Motor Clamp	2			1			
MZ-109	Encoder Coupling Hat	2			1			
MZ-110	Shoulder Joint Pinion	2			1			
MZ-111-101	Arm Tubes (3.0 OD)	2						1
MZ-111-102	Arm Tubes (2.5 OD)	2						1
MZ-112	Idler Pulley Shaft, Shoulder	2			1			
MZ-113	Idler Pulley Shaft Base Shoulder	2			1			
MZ-114	Idler Pulley, Shoulder Base (OBSOLETE)	0						
MZ-200	Waist Drive Drum	2				1		
MZ-201	Waist Drive Drum Plate	2				1		
MZ-202	Waist Base Plate, Top Front	2				1		
MZ-203	Waist Base Plate, Middle	2				1		
MZ-204	Waist Base Plate, Bottom Front	2				1		
MZ-205	Waist Base Plate, Bottom Left	1					1	
MZ-206	Waist Base Plate, Bottom Right	1					1	
MZ-207	Waist Base Plate, Top Left	1					1	
MZ-208	Waist Base Plate, Top Right	1					1	

Draw 1
"523"

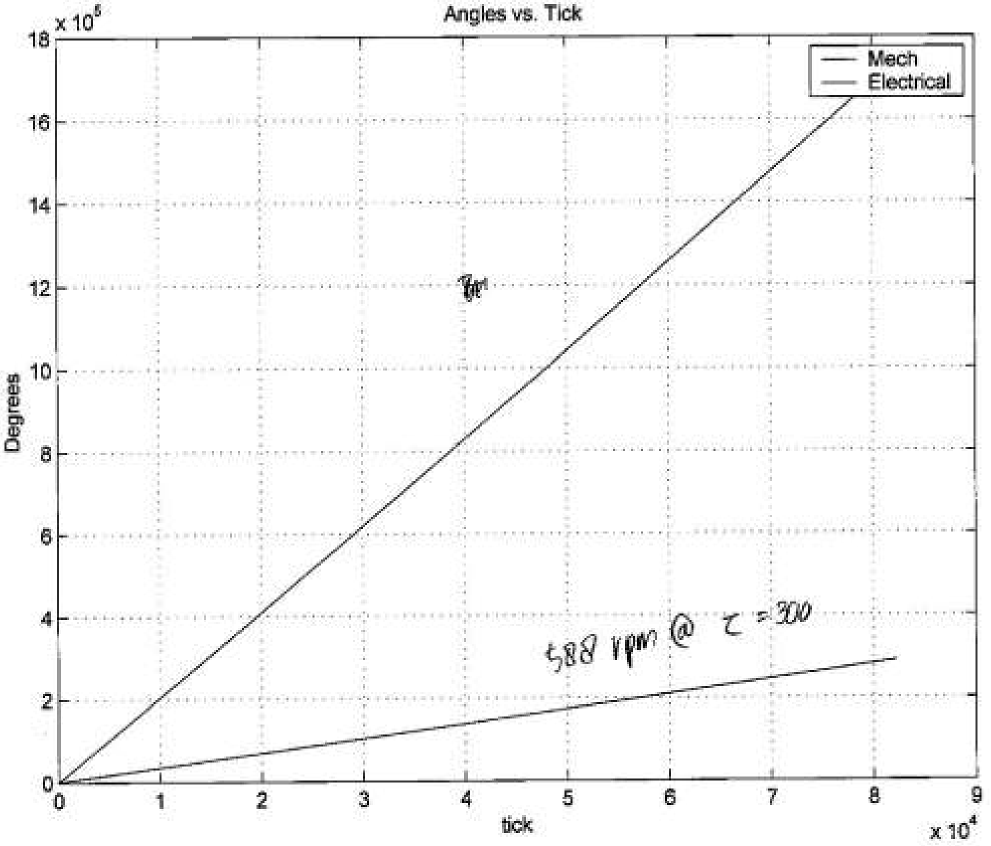
CCW



rpm = 563.75 at $\tau = 900$

elbow 1
"523"

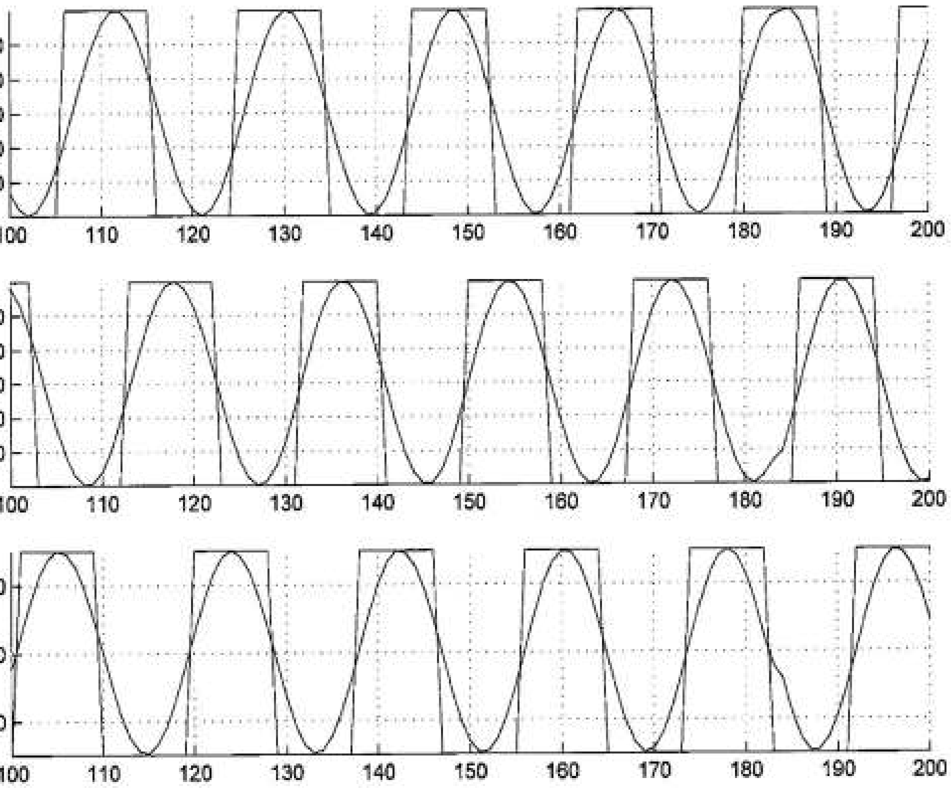
CCW

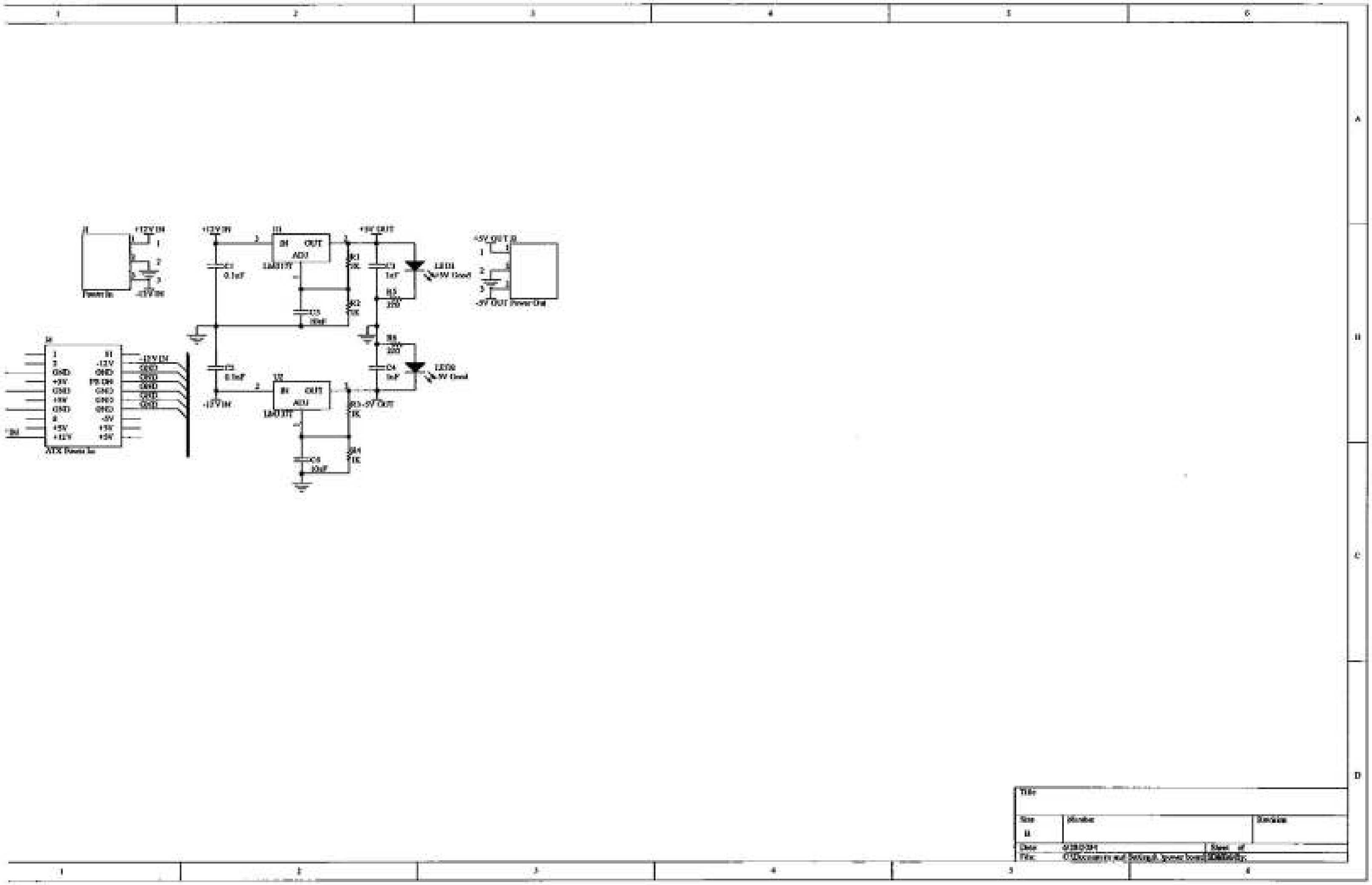


588 rpm @ $\tau = 300$

elbow 7
"523"

CW




1
2
3
4
5
6
A
B
C
D
J1
+12V IN
Power In
-12V IN
+12V IN
U1
IN
OUT
ADJ
LM317T
C1
0.1uF
R1
1K
R2
1K
C5
10uF
+5V OUT
C3
1uF
R5
220
LED1
+5V Good
+5V OUT
J2
-5V OUT
Power Out
J4
1
2
GND
+5V
GND
+5V
GND
8
+5V
+12V
11
-12V
GND
PS ON
GND
GND
GND
-5V
+5V
+5V
ATX Power In
-12V IN
GND
GND
GND
GND
GND
GND
C2
0.1uF
-12V IN
U2
IN
OUT
ADJ
LM337T
R3
-5V OUT
1K
R4
1K
C6
10uF
R6
220
C4
1uF
LED2
-5V Good
Title
Size
B
Number
Revision
Date
Sheet
of
File
By:

~~[illegible]~~ index calibration

1100 } 42 count
1142

7727 - 7685 = $\boxed{+42}$

348.8386°

6 : 360°

5 : 300°

$\boxed{-182}$ 300°

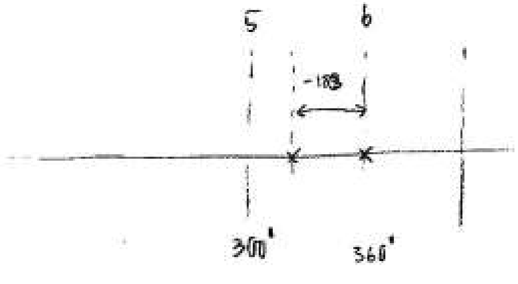
$\boxed{-42 \text{ counts from } 360^\circ}$ or $\boxed{+182 \text{ counts from } 300^\circ}$

$$42 \text{ counts} \times \frac{360}{8128} \frac{\text{deg}}{\text{counts}} = 1.86^\circ \text{ mechanical degrees}$$

$$= \underline{\underline{11.16^\circ}} \text{ electrical degrees}$$

(×6)

electrical degrees at index = 360 - 11.16° = 348.8386°

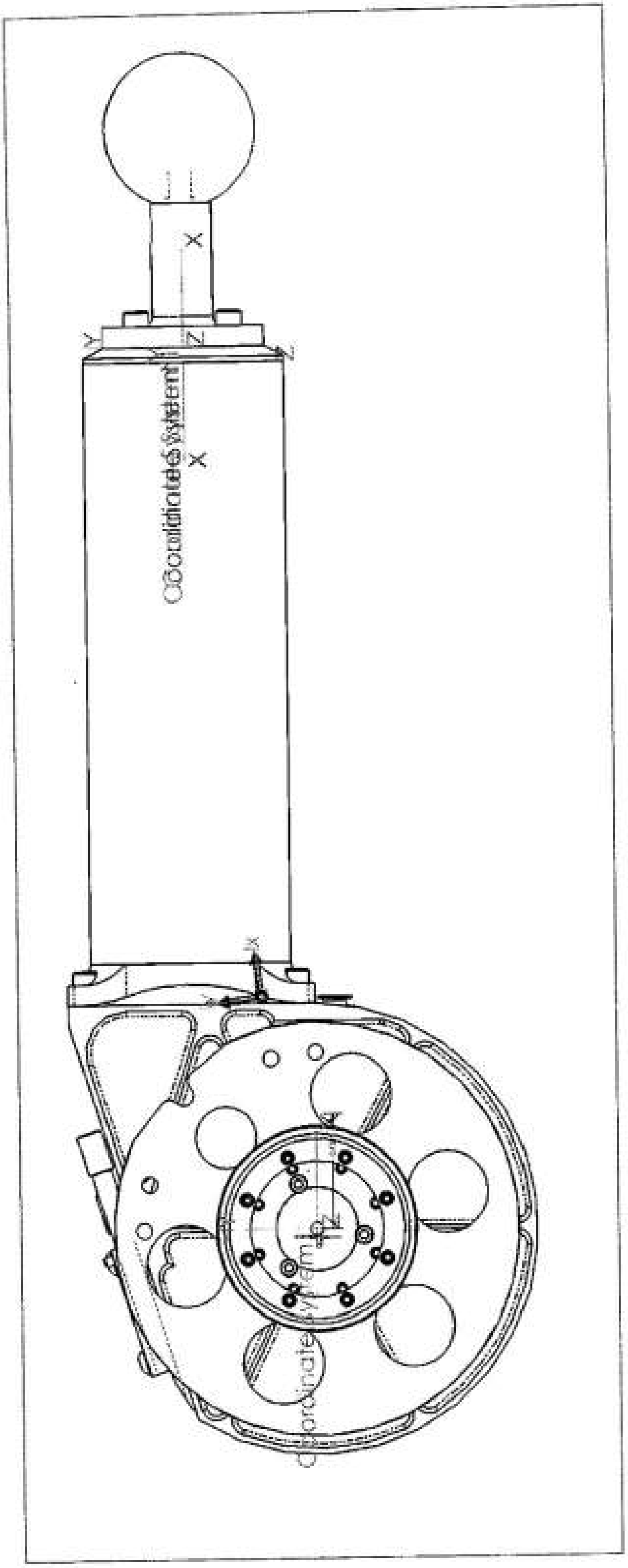


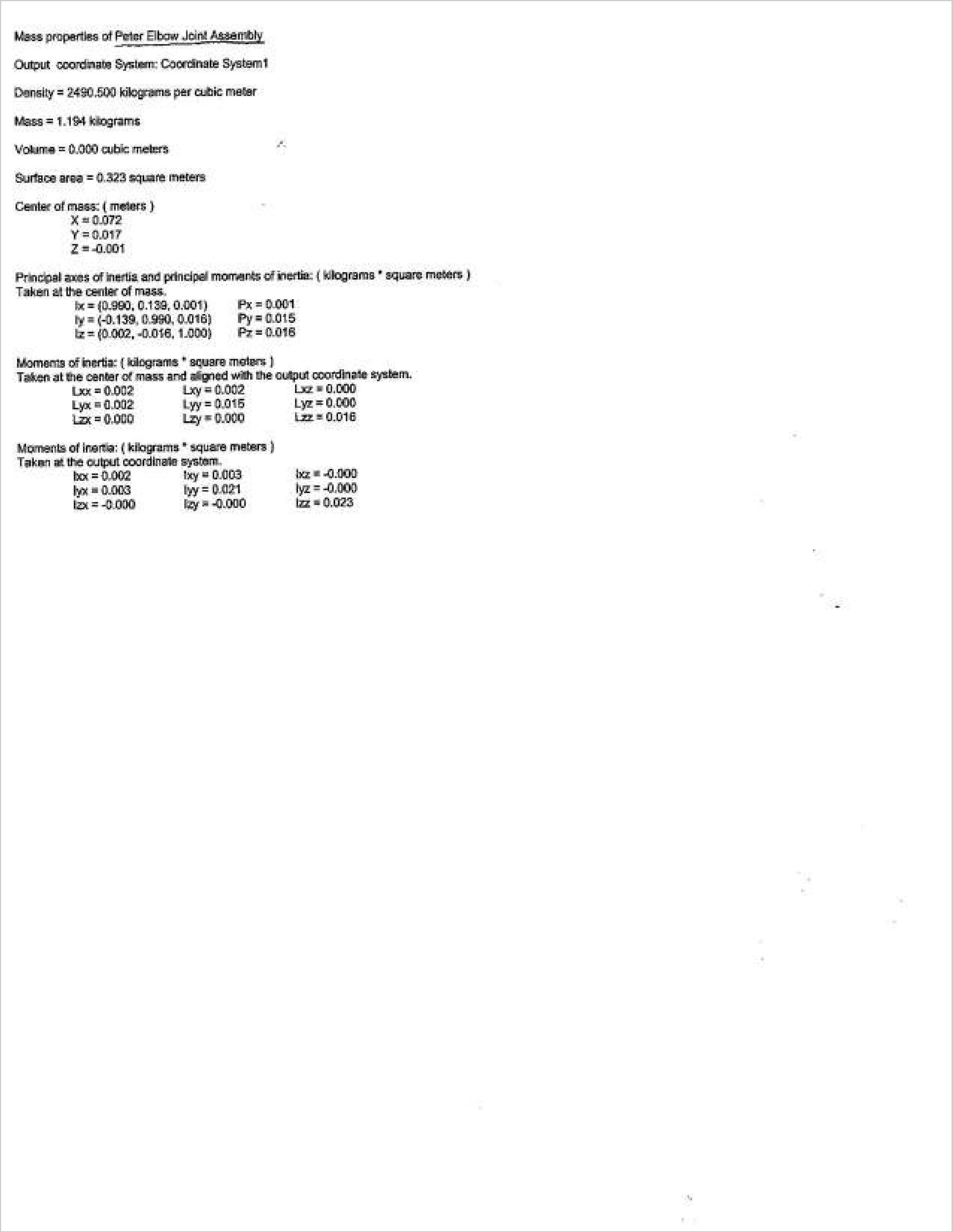
300

360 - 1

182 counts = 48.25° electrical

360


Coordinate System
X
Y
Z
X
Y
Z
X
Y
Z



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Mass properties of Peter Elbow Joint Assembly

Output  coordinate System: Coordinate System1

Density = 2490.500 kilograms per cubic meter

Mass = 1.194 kilograms

Volume = 0.000 cubic meters

Surface area = 0.323 square meters

Center of mass: ( meters )
        X = 0.072
        Y = 0.017
        Z = -0.001

Principal axes of inertia and principal moments of inertia: ( kilograms * square meters )
Taken at the center of mass.
        Ix = (0.990, 0.139, 0.001)      Px = 0.001
        Iy = (-0.139, 0.990, 0.016)     Py = 0.015
        Iz = (0.002, -0.016, 1.000)     Pz = 0.016

Moments of inertia: ( kilograms * square meters )
Taken at the center of mass and aligned with the output coordinate system.
        Lxx = 0.002         Lxy = 0.002         Lxz = 0.000
        Lyx = 0.002         Lyy = 0.015         Lyz = 0.000
        Lzx = 0.000         Lzy = 0.000         Lzz = 0.016

Moments of inertia: ( kilograms * square meters )
Taken at the output coordinate system.
        Ixx = 0.002         Ixy = 0.003         Ixz = -0.000
        Iyx = 0.003         Iyy = 0.021         Iyz = -0.000
        Izx = -0.000        Izy = -0.000        Izz = 0.023
```