DEDICATED ICs SIMPLIFY BRUSHLESS DC SERVO AMPLIFIER DESIGN

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INTRODUCTION

Brushless DC motors have gained considerable commercial success in high end four quadrant servo systems, as well as in less demanding, one and two quadrant requirements. Cost sensitive four quadrant applications thus far have not fared as well. Designs which meet cost goals often suffer from poor linearity, and cumbersome protection circuits to assure reliable operation in all four quadrants. Better performance en tails more complex circuitry and the resulting additional components quickly increase size and cost. Part of the design challenge results from the lack of control ICs tailored to four quadrant applications. The other major obstacle has been implementing a reliable and cost effective high-side switch drive. With recently introduced integrated circuits in both areas, it is now possible to design a rugged, low cost, four quadrant brushless DC servo amplifier with relatively low component count and cost.

SERVO AMPLIFIER REQUIREMENTS

First, let's quickly review general servo amplifier requirements. Figure 1 displays motor speed versus torque, depicting four possible modes of operation. While a system may be considered four quadrant by simply having the ability to operate reliably in all four modes, a servo system generally requires *controlled* operation in all four modes. In addition, a smooth, linear transition between quadrants is essential for high accuracy position and velocity control. The major performance differences between brushless DC servo amplifiers are related to accuracy, bandwidth, and quadrant transition linearity.

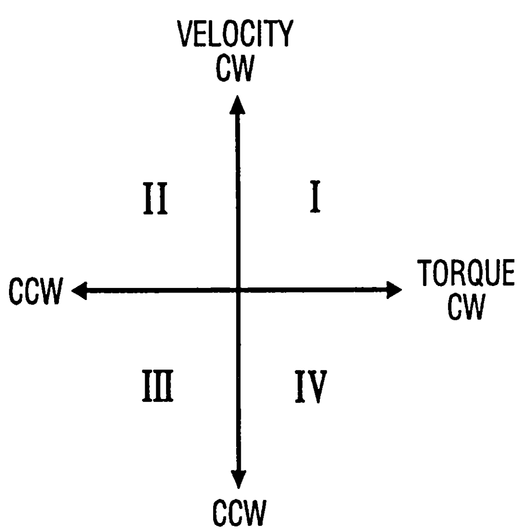

Figure 1 - Four Quadrants of Operation

Most simple brushless DC amplifiers provide two quadrant control, since even the simplest output stages (typically 3 phase bridge) allow rotation reversal. Note that this is operation in quadrants one and three where torque and rotation are in the same direction. This differs from brush motor terminology where two quadrant control normally implies unidirectional rotation with torque control in either direction. Although limited to a single rotation direction, bidirectional torque allows servo velocity control, with rapid, controlled acceleration and deceleration. These characteristics are well suited to numerous applications such as spindle and conveyer drives. With the two quadrant brushless DC amplifier, there are no provisions other than friction to decelerate the load, limiting the system to less demanding applications. Attempting to operate in quadrants two and four will result in extremely nonlinear behavior, and under many circumstances, severe damage to the output stage will follow. This occurs because the two quadrant brushless DC amplifier is unable to completely control current during torque reversal.

TWO QUADRANT VERSUS FOUR QUADRANT CONTROL

Figure 2 shows a three phase bridge output stage for driving a brushless DC motor. Current flow is shown for two quadrant control when operation is in quadrants one or three. The switches commutate based on the motor's

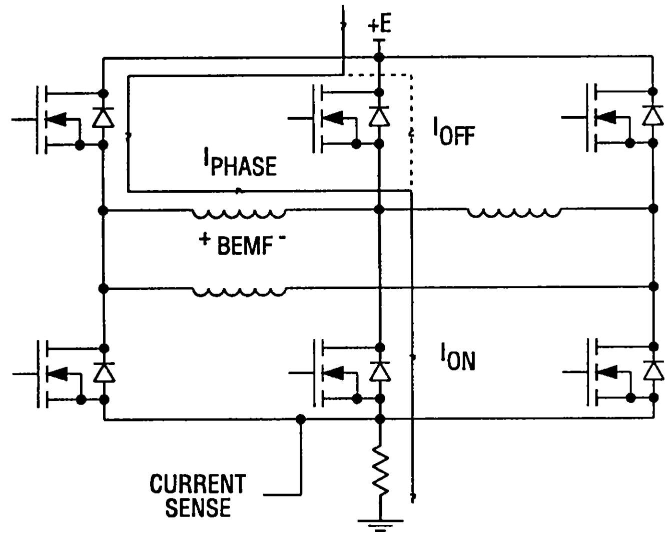

Figure 2 - Two Quadrant Chopping

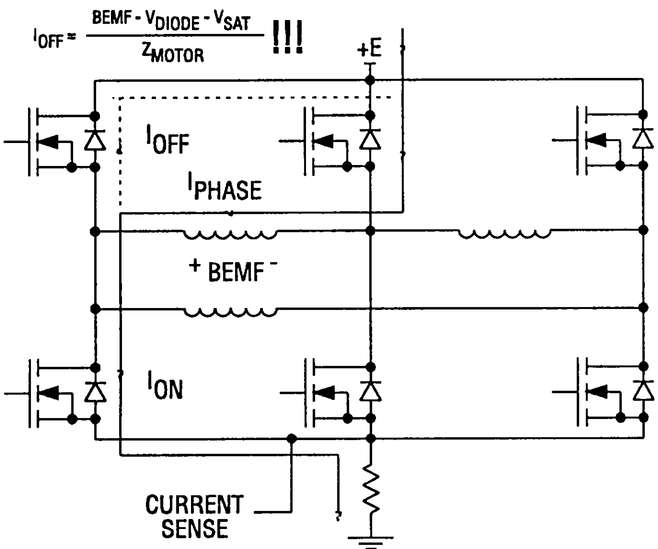

Figure 3 - Two Quadrant Reversal

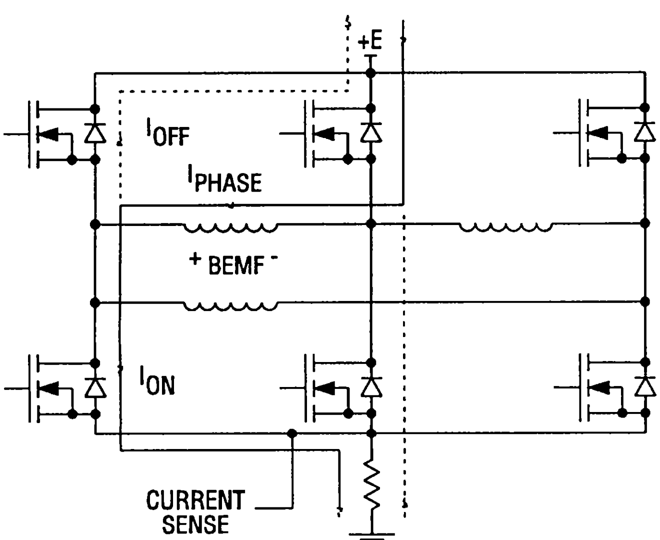

Figure 4 - Four Quadrant Reversal

rotor position, typically using Hall effect sensors for position feedback. Current is controlled by pulse width modulating (PWM) the lower switches. Figure 3 shows current flow if the direction of torque were reversed. The upper switch essentially shorts the motor's back EMF (BEMF), causing current to quickly decay and reverse direction. The current then rises to a value limited only by the motor and drive impedance, yet is undetected by supply or ground sense resistors. As the motor speed rises, its BEMF proportionally increases, quickly escalating the potential circulating current. Even if the output stage is built rugged enough to withstand this abuse, the high uncontrolled current causes high uncontrolled torque, making this technique unsuitable for most servo control applications.

By pulse width modulating the upper switches along with the lower switches, uncontrolled circulating currents are avoided. With both upper and lower switches off during during the PWM off time, motor current will always decay as shown in figure 4. Additionally, motor current always flows through the ground sense resistor, allowing easy detection for feedback. The remainder of this article will feature this mode of control, as it is well suited for a variety of demanding requirements. It should be noted however, that a penalty in the form of reduced efficiency must be paid for the improvement in control characteristics. With two switches operating at the PWM frequency, as opposed to one with two quadrant control, switching losses are nearly doubled. Ripple current is also increased which results in greater motor core loss. Although this is a small price to pay under most circumstances, extremely demanding applications may require switching between two and four quadrant operation for optimum efficiency and control.

FOUR QUADRANT CONTROLLER REQUIREMENTS

In addition to switching both upper and lower transistors, a few supplementary functions are required from the control circuit for reliable four quadrant operation. With two quadrant switching, there is inherent dead time between conduction of opposing upper and lower switches, making cross conduction virtually impossible. Four quadrant control immediately reverses the state of opposing switches at torque reversal, thus requiring a delay between turning the conducting device off and the opposing device on to avoid simultaneous conduction and possible output stage damage.

When torque is reversed, energy stored in the rotating load is transferred back to the power supply, quickly charging the bus storage capacitor. A clamp circuit is typically used to dissipate the energy and limit the maximum bus voltage. As a second line of defense, an over-voltage comparator is often employed to disable the output if the bus voltage exceeds the clamp voltage by more than a few volts.

CURRENT LOOP CONTROL TECHNIQUE

A transconductance amplifier is normally used for brushless DC servo applications, providing direct control of motor torque. Average current feedback is usually employed rather than the more familiar peak current

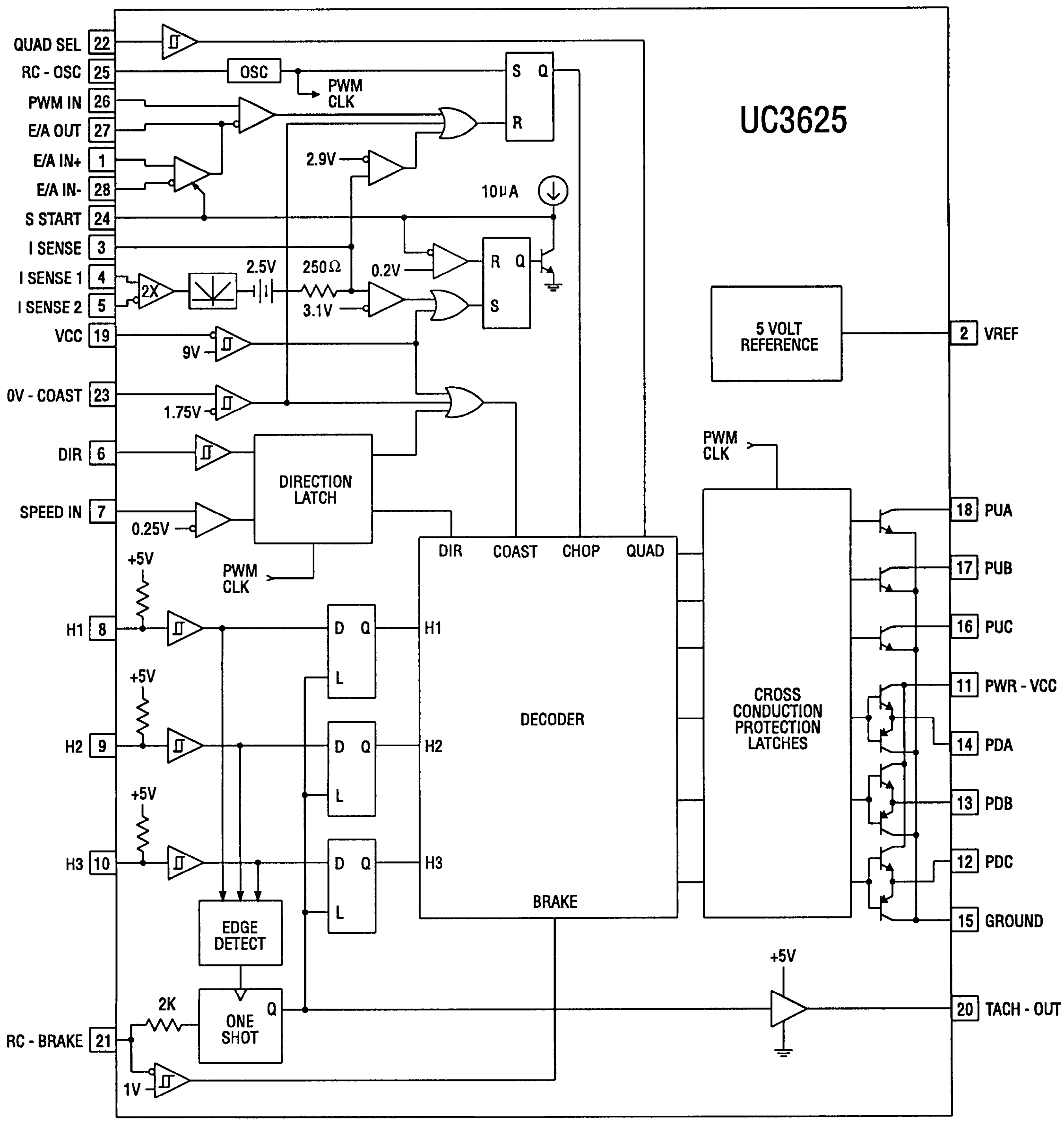

Figure 5 - UC3625 Block Diagram

control for several reasons. Peak current control is subject to subharmonic oscillation at the switching frequency for duty cycles above 50%. This condition is easily circumvented in power supply applications by summing an appropriately scaled ramp signal derived from the PWM oscillator with the current sense signal. This technique is commonly refered to as slope compensation. It can also be shown [3] that for a given inductor current decay rate, which is essentially fixed in a power supply application, there is an optimal compensation level which will produce an output current independent of duty cycle. Unfortunately, the inductor current decay rate in a four quadrant motor control system varies with both speed and supply voltage, making an optimal slope compensation circuit fairly complex. Simpler circuits which provide overcompensation assure stability but will degrade accuracy. Furthermore, severe gain degradation occurs when inductor current becomes discontinuous regardless of slope compensation, causing large nonlinearity at light load. This effect can be particularly troublesome for a position control servo. Average current feedback avoids these problems, and is therefore the preferred current control technique for servo applications.

UC3625 BRUSHLESS DC CONTROLLER

Figure 5 shows the UC3625 block diagram. Designed specifically for four quadrant operation, it minimizes the external circuitry required to implement a brushless DC servo amplifier. Flexible architecture and supplementary features make the UC3625 well suited to less demanding applications as well. The UC3625 is described in detail in references [4] and [7], however a few features critical for reliable four quadrant operation should be noted.

Cross conduction protection latches eliminate the possibility of simultaneous conduction of upper and lower switches due to driver and switch turn-off delays. Additional analog delay circuits normally associated with this function are eliminated allowing direct switch interface and reduced component count. An absolute value buffer following the current sense amplifier provides an average winding current signal suitable for feedback as well as protection. An over-voltage comparator disables the outputs if the bus voltage becomes excessive.

Although not absolutely necessary for four quadrant systems, a few additional features enhance two quadrant operation and simplify implementation of switched two / four quadrant control for optimized systems. A direction latch with analog speed input prevents reversal until an acceptably low speed is reached, preventing output stage damage. Two or four quadrant switching can be selected during operation with the Quad Select input. A brake input provides current limited dynamic braking, suitable for applications which require rapid deceleration, but do not need tight servo control.

A SIMPLE BRUSHLESS DC SERVO AMPLIFIER

To demonstrate the relative simplicity with which a brushless DC servo amplifier can be implemented, a 6 amp, off-line 115 VAC amplifier was designed and constructed. Note that current and voltage rather than horsepower are specified. Although theoretically capable of in excess of one horsepower, simultaneous high speed and torque are typically not required in servo applications, reducing the actual output power, and the corresponding power supply requirement. Average current feedback is employed, providing good bandwidth and power supply rejection, thus making the amplifier suitable for many demanding requirements. A complete amplifier schematic is shown in figure 6.

A high performance brushless servo motor from MFM Technology, Inc. was used to evaluate the amplifier. While most of the design is independent of motor parameters, several functions should be optimized for a particular motor and operating conditions. The motor used has the following electrical specifications:

<u>Model M - 178</u>

K_T	79 oz.in./Amp
R_M	1.3 ohms
L_M	5.5 mH
Poles	18

OUTPUT STAGE DESIGN

Having selected a four quadrant control strategy, we proceed to the output stage design, and work back to the controller. High voltage MOSFETs are well suited to this power level, however IGBTs may also be incorporated. MOSFETs were selected to minimize size and complexity, since the body diodes can be used for the flyback rectifiers. Unfortunately, this places greater demands on the MOSFET, and increases the device dissipation. The MOSFETs body diode is typically slower and stores more charge than a discrete high speed rectifier, which necessitates a slower turn-on and a corresponding increase in switching losses. These losses are partially offset by choosing a MOSFET with sufficiently low conduction losses which offers the secondary benefits of greater peak current capability and reduced thermal

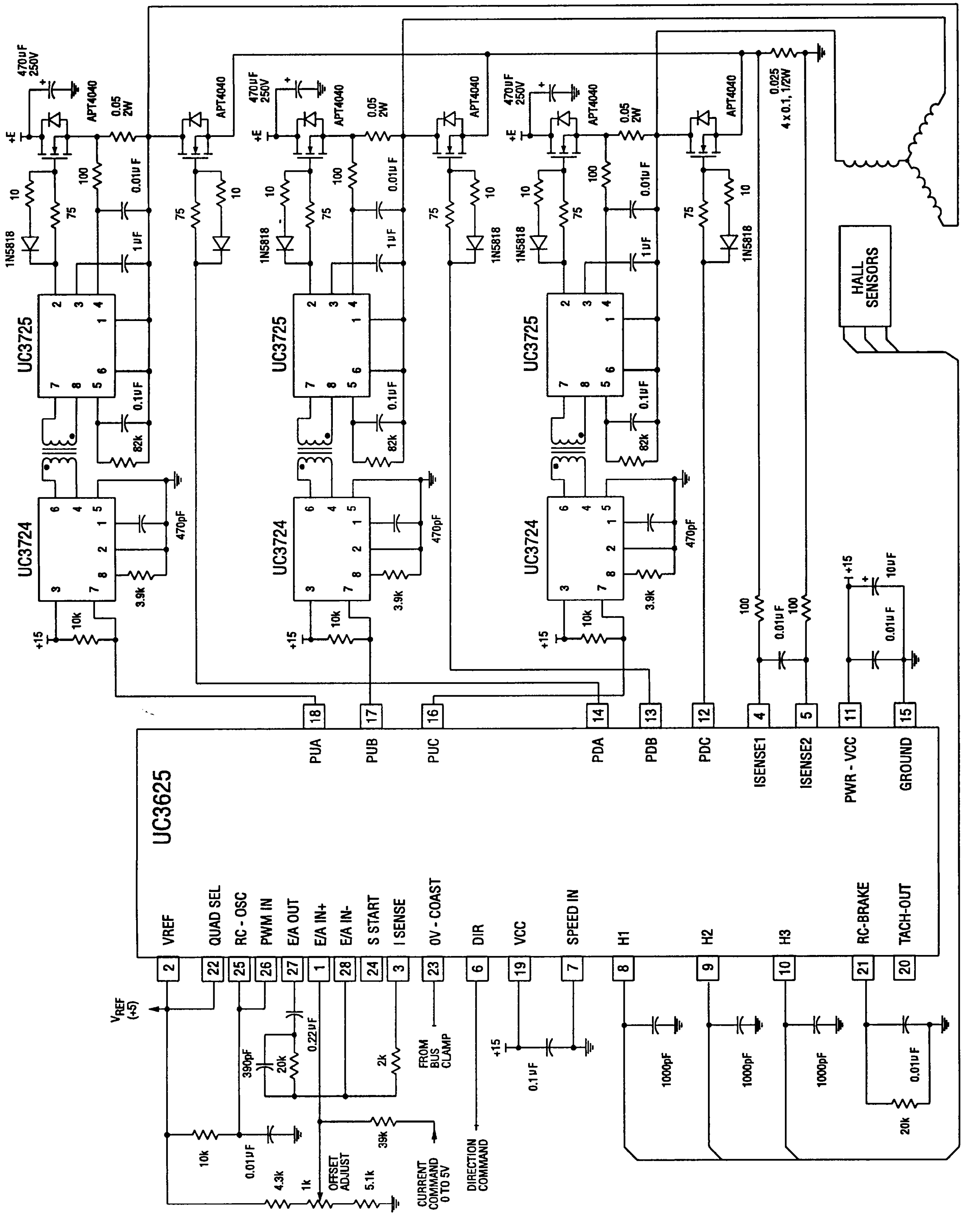

Figure 6 - Brushless DC Servo Amplifier Schematic

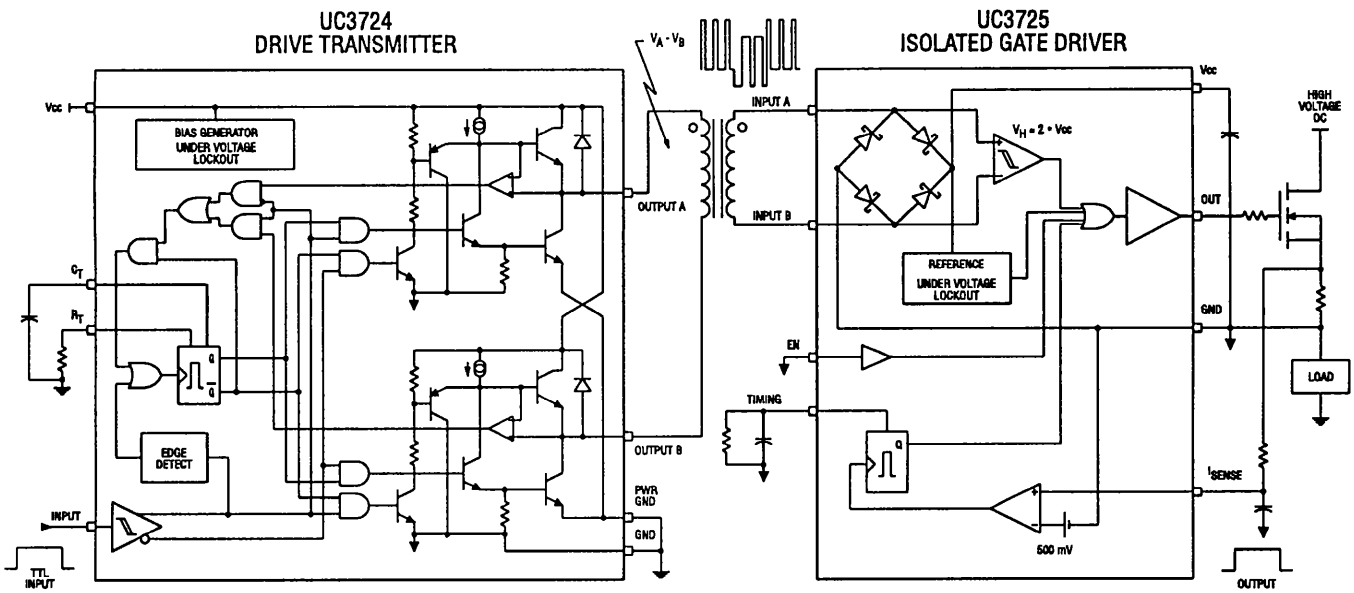

Figure 7 - UC3724/UC3725 Isolated MOSFET Driver

resistance. APT4030BN MOSFETs were selected for the output stage to handle the 6 amp load currents while providing good supply voltage transient immunity. Rated at 400 volts and 0.30 ohms, they allow high efficiency operation and have sufficient breakdown voltage for reliable off-line operation.

While the lower three FETs require simple ground referenced drive, and are easily driven directly from the UC3625 the design of the drive circuit for the upper three FETs has traditionally been challenging. Discrete implementation of the required power supply and signal transmission is often bulky and expensive. In an effort to reduce size and cost, critical functions are often omitted, opening the door to potential reliability problems. Specifically designed for high-side MOSFET drive in motor control systems, the UC3724 / UC3725 IC pair shown in figure 7, offers a compact, low cost solution. A high frequency carrier transmits both power and signal across a single pulse transformer, eliminating separate DC/DC converters, charge pump circuits, and opto-couplers. Signal and power transmission function down to DC, imposing no duty cycle or on-time limitations typical of commonly used charge pump techniques. Under-voltage lockout, gate voltage clamp, and over current protection assure reliable operation.

Design of the upper driver is a straight forward procedure, and is described in detail in reference [5].
For this application, the driver is designed with the following specifications:

500 V minimum isolation
300 kHz carrier frequency
10 Amp over-current fault
10 ms over-current off time

The pulse transformer uses a 1/2 inch O.D. toroid core (Philips 204T250-3E2A) with a 15 turn primary and 17 turn secondary. For high voltage isolation, Teflon insulated wire is used for both primary and secondary.

To provide rapid turn-off for minimal switching losses, with slower turn-on for di/dt control, a resistor/resistor-diode network is used in place of a single gate resistor. Although present generation MOSFETs can reliably commutate current from an opposing FETs body diode at high di/dt, the resulting high peak current and diode snap limit practical circuits to a more moderate rate. This increases dissipation, but significantly eases RFI filtering and shielding, as well as relaxing layout constraints. Additionally, a low impedance is maintained in the off state while turn-on dv/dt is decreased, dramatically reducing the tendency for dv/dt induced turn on. The same gate network is used for both upper and lower MOSFETs.

A sense resistor in series with the bridge ground return provides a current signal for both feedback and current limiting. This resistor, as well as the upper driver current sense resistors should be non-inductive to minimize ringing from high di/dt. Any inductance in the power circuit represents potential problems in the form of additional voltage stress and ringing, as well as increasing switching times. While impossible to eliminate, careful

APPLICATION NOTE

layout and bypassing will minimize these effects. The output stage should be as compact as heat sinking will allow, with wide, short traces carrying all pulsed currents. Each half-bridge should be separately bypassed with a low ESR/ESL capacitor, decoupling it from the rest of the circuit. Some layouts will allow the input filter capacitor to be split into three smaller values, and serve double duty as the half-bridge bypass capacitors.

CONTROLLER SETUP

The UC3625 switching frequency is programmed with a timing resistor and capacitor. Unless the motor's inductance is particularly low, 20 kHz will provide acceptable ripple current and switching losses while minimizing audible noise.

(1) $F = 2 / R_{OSC}C_{OSC}$

The relatively small oscillator signal amplitude requires careful timing capacitor interconnect for maximum frequency stability. Circuit board traces should to be as short as possible, directly connecting the capacitor between pins 25 and 15, with no other circuits sharing the board trace to pin 15 (ground).

When tight oscillator stability is required, or multiple systems must be synchronized to a master clock, the circuit shown in figure 8 can be used. As shown, the circuit buffers, and then differentiates the falling edge of the master oscillator. The last stage provides the necessary current gain to drive the 47 ohm resistor in series with the timing capacitor. If the master clock is from a digital source, the first two stages are omitted, and the clock signal is interfaced directly to the final stage through a restive divider as shown. The slaves are programmed to oscillate at a lower frequency than the master. The pulse injected across the 47 ohm resistor causes the oscillator to terminate its cycle prematurely, and thus synchronize to the master clock.

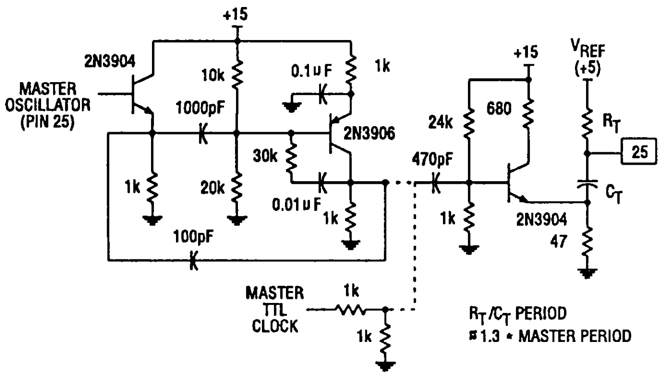

Figure 8 - External Synchronization Circuit

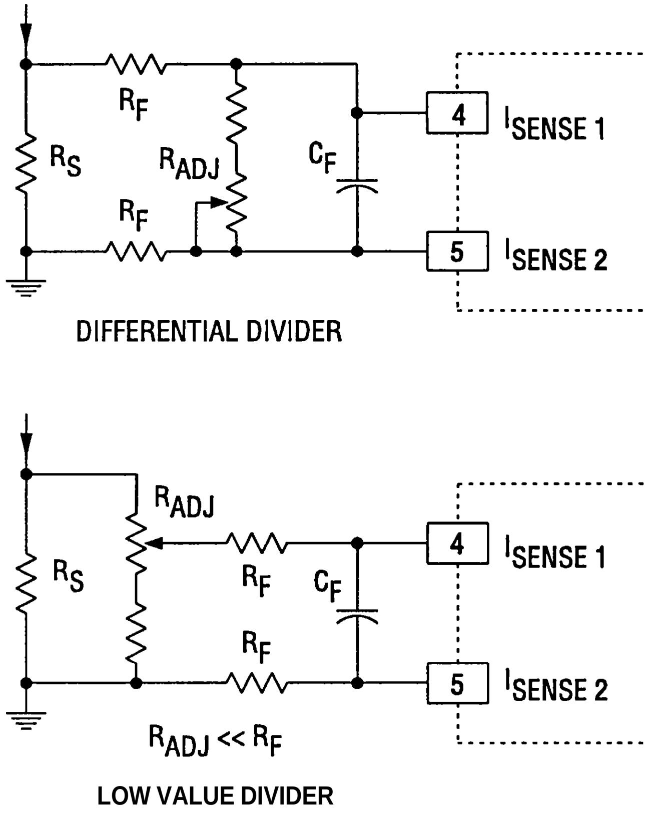

Figure 9 - Balance Impedance Current Sense Input Circuits

The RC-Brake pin serves two functions: Brake command input (not used in this design), and tachometer / digital commutation filter one-shot programming. Whenever the commutation state changes, the one-shot is triggered, outputting a tach pulse and inhibiting another commutation state change until the one-shot terminates. The one-shot pulse width is programmed for approximately 1/2 the shortest commutation period.

(2) $T_{PW} = 0.70\ R_T C_T$

where the shortest commutation period = $20 / (RPM_{MAX} N_{POLES})$

CURRENT SENSING AND FEEDBACK

For optimum current sense amplifier performance, the input impedance must be balanced. Low value resistors (100 to 500 ohm) are used to minimize bias current errors and noise sensitivity. Additionally, if the sense voltage must be trimmed, a low value input divider or a differential divider should be used to maintain impedance matching, as shown in figure 9.

An average current feedback loop is implemented by the circuit shown in figure 10. With four quadrant chopping, motor current always flows through the sense resistor. When PWM is off however, the flyback diodes conduct,

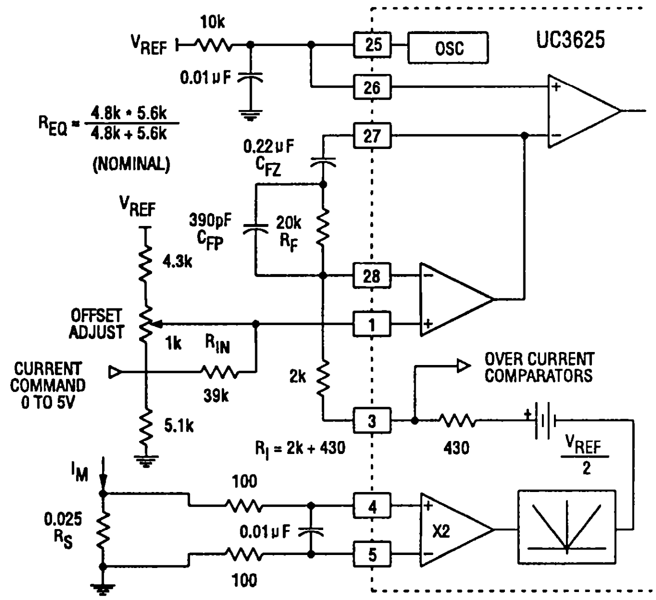

Figure 10 - Average Current Feedback Circuit Configuration

causing the current to reverse polarity through the sense resistor. The absolute value amplifier cancels the current polarity reversal by inverting the negative current sense signal during the flyback period. The output of the absolute value amplifier therefore is a reconstructed analog of the motor current, suitable for protection as well as feedback loop closure.

When the current sense output is used to drive a summing resistor as in this application example, the current sense output impedance adds to the summing resistor value. The internal output resistor and the amplifier output impedance can both significantly effect current sense accuracy if the external resistance is too low. Although not specified, the total output impedance is typically 430 ohms at 25 degrees C. Over the military temperature range of -55 to +125 degrees C, the impedance ranges from approximately 350 to 600 ohms. An external 2 k resistor will result in an actual 2.43 k summing resistance with reasonable tolerance. A higher value external resistor and trim pot will be required if high closed current loop accuracy is required.

The current sense output offset voltage is derived from the +5 V reference voltage. By developing the command offset from the +5 V reference, current sense drift over temperature is minimized. The offset divider must be trimmed initially to accommodate the current sense amplifier offset tolerance.

POWER SUPPLY AND BUS CLAMP

Input power is filtered to reduce conducted EMI, and transient protected using MOVs. Power-up current surge is suppressed using a NTC thermistor, while a bridge rectifier and capacitive filter complete the high voltage supply. A small 60 Hz. transformer supplies 15 Volts through a three pin regulator to power the control and drive circuits.

A bus clamp is easily designed around a UC3725 MOSFET driver, as shown in figure 11. As in the high-side switch drive, the UC3725 assures reliable operation, particularly during power-up and power-down. The divider current is set to 1 mA at the threshold, which is a reasonable compromise between input bias current error and dissipation. An additional tap programs the over-voltage coast a few volts above the bus clamp, saving a resistor and some dissipation while reducing the tolerance between the bus clamp and the over-voltage coast. Setting the bus clamp discharge current equivalent to the maximum motor current will assure effective clamping under all conditions. The load resistor value is therefore:

(3) $$R_L = \frac{V_{CLAMP}}{I_{MAX}}$$

The load resistor dissipation is dependant on the energy removed from the load inertia, and the frequency with which the energy is removed.

(4) $$P_{LOAD} = 1/2\, f J\,(\omega_1^2 - \omega_2^2)$$

where J = inertia in Nm sec^2
ω_1 = initial velocity in rad/sec
ω_2 = final velocity in rad/sec

Note that if the deceleration time approaches the load resistor's thermal time constant, a higher power resistor will be required to maintain reliability.

CURRENT LOOP OPTIMIZATION

The block diagram of the current control loop is shown in figure 12. The current sense input filter has minimal affect on the loop and can be ignored, since the filter pole must be much higher than the system bandwidth to maintain waveform integrity for over-current protection. The current sense resistor R_S, is chosen to establish the peak current limit threshold, which is typically set 20% higher than the maximum current command level to provide over-current protection during abnormal conditions. Under normal circumstances with a properly compensated current loop, peak current limit will not be exercised. The input divider network provides both offset adjustment and attenuation, with R_{IN} selected to accomodate the current command signal range.

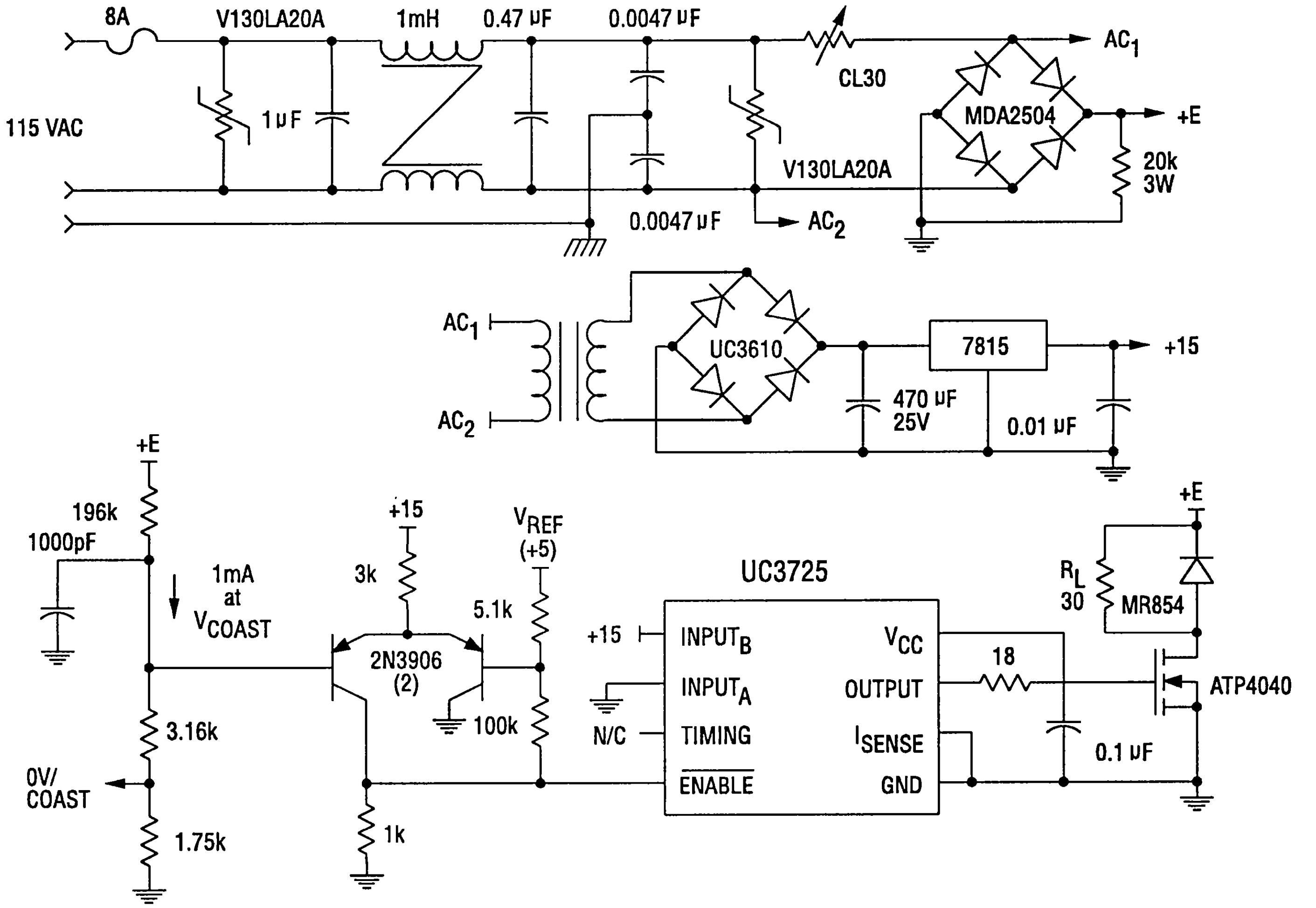

Figure 11 - Power Supply and Bus Clamp

All PWM circuits are prone to subharmonic oscillation if the modulation comparator's two input waveform slopes are inappropriately related. This behavior is most common in peak current feedback schemes, where slope compensation is typically required to achieve stability. Average current feedback systems will exhibit similar behavior if the current amplifier gain is excessively high at the switching frequency. As described by Dixon [2] to avoid subharmonic oscillation for a single pole system: *The amplified inductor current downslope at one input of the PWM comparator must not exceed the oscillator ramp slope at the other comparator input.* This criterion sets the maximum current amplifier gain at the switching frequency, and indirectly establishes the maximum current loop gain crossover frequency.

A voltage proportional to motor current, which is the inductor current, is generated by the current sense resistor and the current sense amplifier circuitry internal to the UC3625 This waveform is amplified and inverted by the current amplifier and applied to the PWM comparator input. Due to the signal inversion, the motor

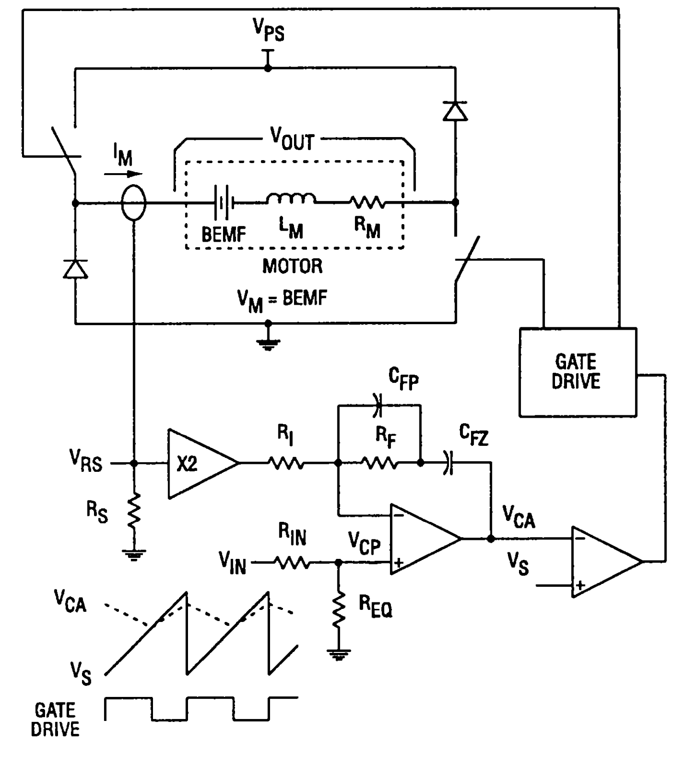

Figure 12 - Current Loop Block Diagram

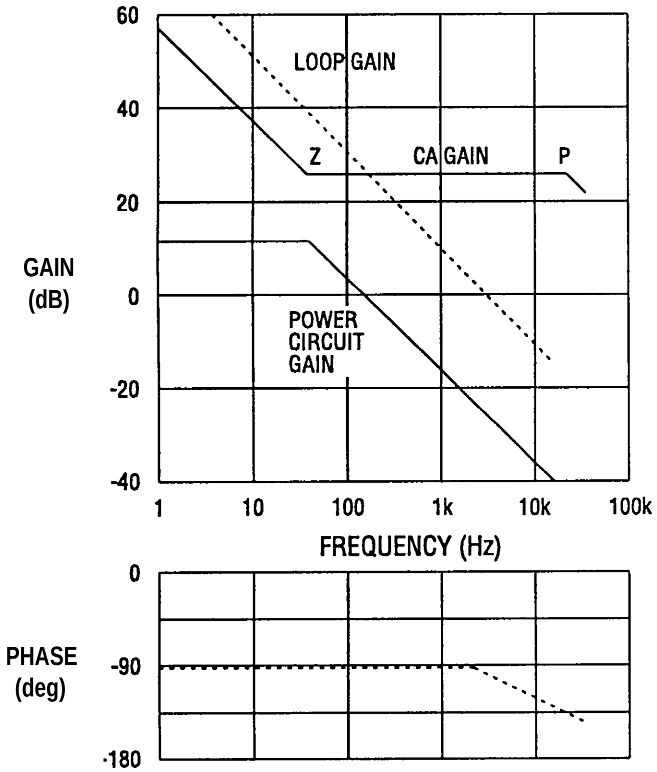

Figure 13 - Open Loop Gain and Phase Versus Frequency

current downslope appears as an upslope as shown in figure 12. To avoid subharmonic oscillation, the current amplifier output slope must not exceed the oscillator ramp slope. A motor control system typically operates over a wide range of output voltages, and is usually powered from an unregulated supply. The operating conditions which cause the greatest motor current downslope must be determined in order to determine the maximum current amplifier gain which will maintain stability. When four quadrant chopping is used, the inductor discharge rate is described by:

$$\text{Motor Current Downslope} = \frac{V_{PS} + V_M}{L_M}$$

The greatest discharge slope therefore occurs when the supply and BEMF voltages are maximum.

The oscillator ramp slope is simply:

$$\text{Oscillator Ramp Slope} = \frac{V_S}{T_S} = V_S f_S$$

Where: V_S is the oscillator ramp peak to peak voltage (1.2 V for the UC3625)
T_S is the switching period
f_S is the switching frequency

The maximum current amplifier gain at the switching frequency is determined by setting the amplified inductor current downslope equal to the oscillator ramp slope.

$$\frac{V_{PS} + V_M}{L_M} R_S G_{CA} = V_S f_S$$

$$(5) \qquad \therefore \max G_{CA} = \frac{\Delta V_{CA}}{\Delta V_{RS}} = \frac{V_S f_S L_M}{R_S (V_{PS} + V_M)}$$

The maximum BEMF and supply voltage for the design example are 87 and 175 Volts respectively, which translates to a motor speed of 1500 RPM, and a high-line supply voltage of 125 Volts AC. Using equation (5) with an oscillator voltage of 1.2 volts peak to peak at a frequency of 20 kHz, the maximum value for G_{CA} is 20.2, or 26 dB. The current sense amplifier's gain of two is also part of G_{CA}. With R_I equal to 2.43 k, 20 k is selected for R_F to allow for tolerances, resulting in an actual G_{CA} of 16.5, or 24 dB.

The small-signal control to output gain of the current loop power section is described by:

$$(6) \qquad \frac{\Delta V_{RS}}{\Delta V_{CA}} = \frac{R_S 2 V_{PS}}{V_S s L_M}$$

Note that the factor of two in the numerator is a result of four quadrant chopping which only utilizes one-half of the modulator's input range for a given quadrant of operation.

The overall open loop gain of the current loop is the product of the actual current amplifier gain and the control to output gain of the power circuit. The result is set equal to one to solve for the loop gain crossover frequency, f_C:

$$(7) \qquad G_{CA} \frac{R_S 2 V_{PS}}{V_S 2\pi f_C L_M} = 1$$

$$(8) \qquad f_C = \frac{G_{CA} R_S V_{PS}}{V_S \pi L_M}$$

At high line, where the supply is 175 Volts DC, f_C is 3.5 kHz. The crossover frequency drops to 2.8 kHz at low line, where the supply is approximately 140 Volts DC. If greater bandwidth is required, the current amplifier gain must be increased, requiring a corresponding increase in switching frequency to satisfy equation (5).

Up to this point the motor's resistance (R_M) has been ignored. This is valid since L_M predominates at the switching frequency. The motor's electrical time constant L_M/R_M, creates a pole, which is compensated for by placing zero $R_F C_{FZ}$ at the same frequency. Additionally, pole $R_F C_{FP} C_{FZ}/(C_{FP}+C_{FZ})$ is placed at f_S to reduce sensitivity to noise spikes generated during switching transitions. The filter pole at f_S also reduces the amplitude and slope of the amplified inductor current waveform, possibly suggesting that the current amplifier gain could be increased beyond the maximum value from equation (5). Experimentally increasing G_{CA} may incur subharmonic oscillation however, since equation (5) is only valid for a system with a single pole response at f_S. For the design example, standard values are chosen for C_{FZ} and C_{FP} of 0.22 μF and 390 pF respectively, placing the zero at 36 Hz, and the pole at 20 kHz. Figure 13 shows open loop gain and phase verses frequency.

At very light loads, the motor current will become discontinuous - motor current reaches zero before the switching period ends. At this mode boundary, the power stage gain suddenly decreases, and the single pole characteristic of continuous mode operation with its 90 degree phase lag disappears. The current loop becomes more stable, but much less responsive. Fortunately, the high gain of current amplifier is sufficient to maintain acceptable closed current loop gain and phase characteristics at typical outer velocity and/or position loop crossover frequencies.

When the current loop is closed, the output voltage of the current sense amplifier $(2V_{RS})$ is equal to the current programming voltage (V_{CP}) at frequencies below the crossover frequency. The closed current loop transconductance is simply:

$$(9) \qquad g_M = \frac{\Delta i_M}{\Delta V_{CP}} = \frac{\Delta V_{RS}/R_S}{\Delta V_{CP}} = \frac{1}{2R_S}$$

At the open loop crossover frequency, the transconductance rolls off and assumes a single pole characteristic. The input divider network attenuates the current command signal to provide compatibility with typical servo controller output voltages, and decreases the closed loop transconductance by the ratio of $R_{EQ}/(R_{EQ}+R_{IN})$. For the design example, the overall amplifier transconductance is 1.25 amps/volt, allowing full scale current (6 amps) with a 5 volt input command.

BIPOLAR TO SIGN/MAGNITUDE CONVERSION

The servo amplifier as shown in figure 6 requires a separate sign and magnitude input command. This is convenient for many microcontroller based systems which solely utilize digital signal processing for servo loop compensation. Analog compensation circuits however, usually output a bipolar signal and require conversion to sign/magnitude format to work with this amplifier. The circuit shown in figure 14 employs a differential amplifier for level shifting and ground noise rejection, and an absolute value circuit with polarity detection for conversion to sign/magnitude format. The current command signal is slightly attenuated and level shifted up 5 volts to allow single supply operation. The input divider circuit has been slightly modified from figure 9 to restore gain and provide a suitable offset adjustment range. Precision resistors (1%) should be used for both the differential amplifier and the absolute value circuit to minimize DC offset errors. Figure 15 shows approximately 2 Amp peak motor current with a 500 Hz sinwave command. Motor current follows the input command with minimal phase lag, however some crossover distortion is present. This is not crossover distortion in the traditional sense, rather it is simply a fixed off-time caused by the cross conduction protection circuitry. Since this distortion is current amplitude independent, and decreases with frequency, its effect on overall servo loop performance is minimal.

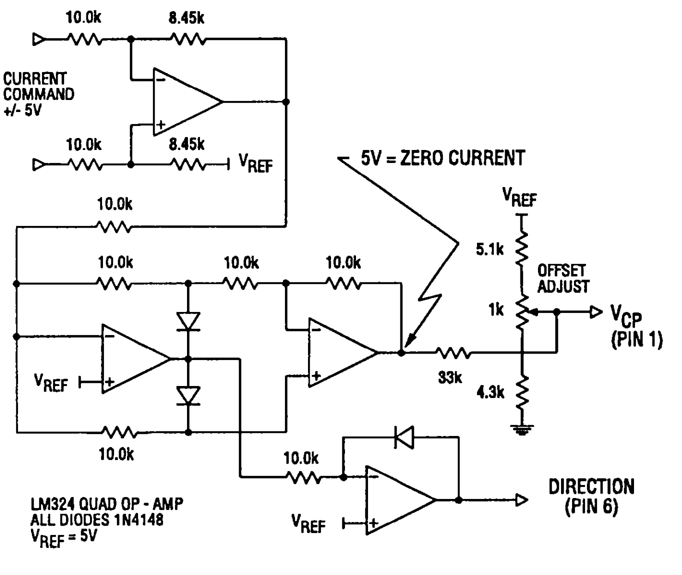

Figure 14 - Bipolar to Sign/Magnitude

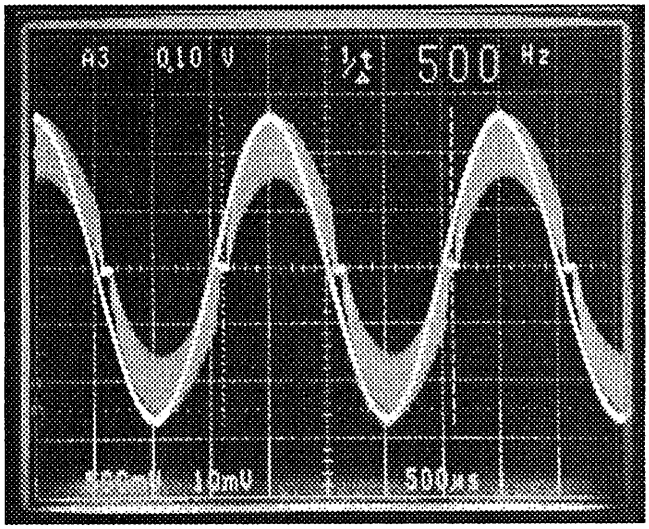

Figure 15 - 5OOHz Sine Wave Command and Output Currents

DIRECT DUTY CYCLE CONTROL

There are many less demanding brushless DC servo applications which do not need a transconductance amplifier function yet require controlled operation in all four quadrants. For these systems, direct duty cycle control, also known as voltage mode control is often employed. Note that this is not voltage feedback, which requires additional demodulation circuitry to develop a feedback signal. With direct duty cycle control the amplifier simply provides open loop voltage gain. This technique is particularly advantageous when a microcontroller is used for servo loop compensation. By outputting a PWM signal directly, a digital to analog conversion is eliminated along with the analog pulse width modulator. While the simplicity of this technique is appealing, there are two major problems which must be addressed. The first and less severe problem is the complete lack of power supply rejection. Good supply filtering will often reduce transients to acceptable levels, while the servo loop compensates for slow disturbances. The second and more troublesome predicament is the output nonlinearity which occurs when transitioning between quadrants. This is best illustrated by examining the DC equations for the two possible cases.

When operating in either quadrant one or three, rotation and torque are in the same direction. Assuming operation is above the continuous/discontinuous current mode boundary, the output voltage is described by:

(10) $V_M = 2V_{PS}D - V_{PS}$

where D = PWM duty cycle

When the direction command is reversed while the motor is rotating, operation switches to quadrant two or four, shifting the modulator's maximum output voltage point from full duty cycle to zero duty cycle.

(11) $V_M = 2V_{PS}(1\text{-}D) - V_{PS}$

Note that the gain does not change, only the reference point has shifted. This occurs because the modulator only has a single quadrant control range -four quadrant operation results from the output control logic which is after the modulator. With the transconductance amplifier previously described, the error amplifier quickly slews during quadrant transitions, providing four quadrant control with minimal disturbance. When direct duty cycle control is used however, the servo loop filter must slew to maintain control. Unfortunately, this causes an immediate loop disturbance, with the greatest severity at the duty cycle extremes. This behavior can greatly effect the performance of an analog compensated servo, and therefore limits such systems to lower performance requirements.

With a microcontroller providing the servo loop compensation, nonlinear duty cycle changes can be accommodated, restoring linearity when transitioning between quadrants. Although nonlinear behavior still occurs when motor current becomes discontinuous, the effect on overall system performance is usually minimal. By correcting for quadrant transition nonlinearities, the advantages of an all digital interface can be exploited without severely degrading system performance. The control system is fully digital right up to the output stage, where the motor's inductance finally makes the conversion to analog by integrating the output switching waveform.

The circuit shown in figure 16 uses a PWM input from a microcontroller to set the output duty cycle and synchronize the oscillator, while another input controls direction.

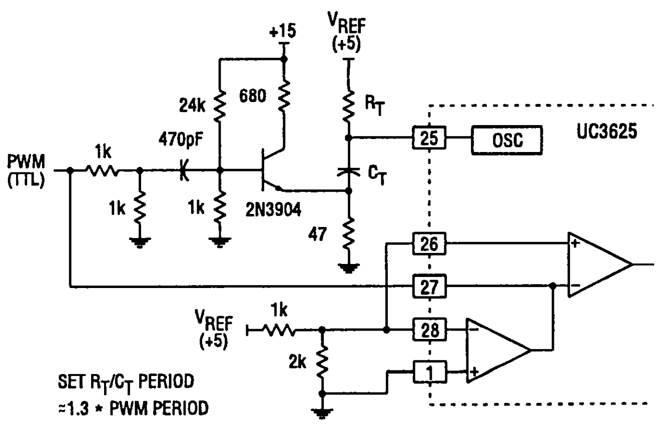

Figure 16 - Digital PWM Interface

Complete line isolation can easily be achieved by using opto-couplers. Although the performance of this technique falls short of the transconductance amplifier, the circuitry's simplicity while maintaining all of the protection features of the UC3625 make it well suited to many cost sensitive applications.

SUMMARY

The application example demonstrates the relative simplicity in implementing a brushless DC transconductance servo amplifier using the latest generation controller and driver ICs. For less demanding applications, direct duty cycle control using a dedicated controller provides size and cost reduction, without sacrificing protection features. While more and more control functions are implemented in microcontrollers today, the task of interfacing to output devices, and providing reliable protection under all conditions will remain a hardware function. Dedicated integrated circuits offer considerable improvement over the discrete solutions used in the past, reducing both size and cost, while enhancing reliability.

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