UNITRODE
APPLICATION NOTE

UC1637/2637/3637
SWITCHED MODE CONTROLLER
FOR DC MOTOR DRIVE

INTRODUCTION

There is an increasing demand today for motor control circuits, as a result of the incredible proliferation of automated position control equipment, which is itself made possible by recent developments in the field of digital computation.

The UC1637 Switched Mode Controller for DC motors is one of several integrated circuits offered by Unitrode for motor controls. This Application Note presents the general principles of its operation and the circuit details that optimize its use. As an illustration we will carry out an actual design which will involve not only the UC1637, but also a

power H-bridge using MOSFET transistors, and a modern DC motor tachometer. Using the tach output and UC1637's error amplifier, we will close the velocity control loop after a brief analysis of the factors that affect the feedback loop stability.

To achieve high efficiency power amplification, the UC1637 uses pulse width modulation, or PWM. This technique is employed today in many different circuits where power losses must be minimized, and is most suitable in applications involving inductive loads such as motors voice coils, etc.
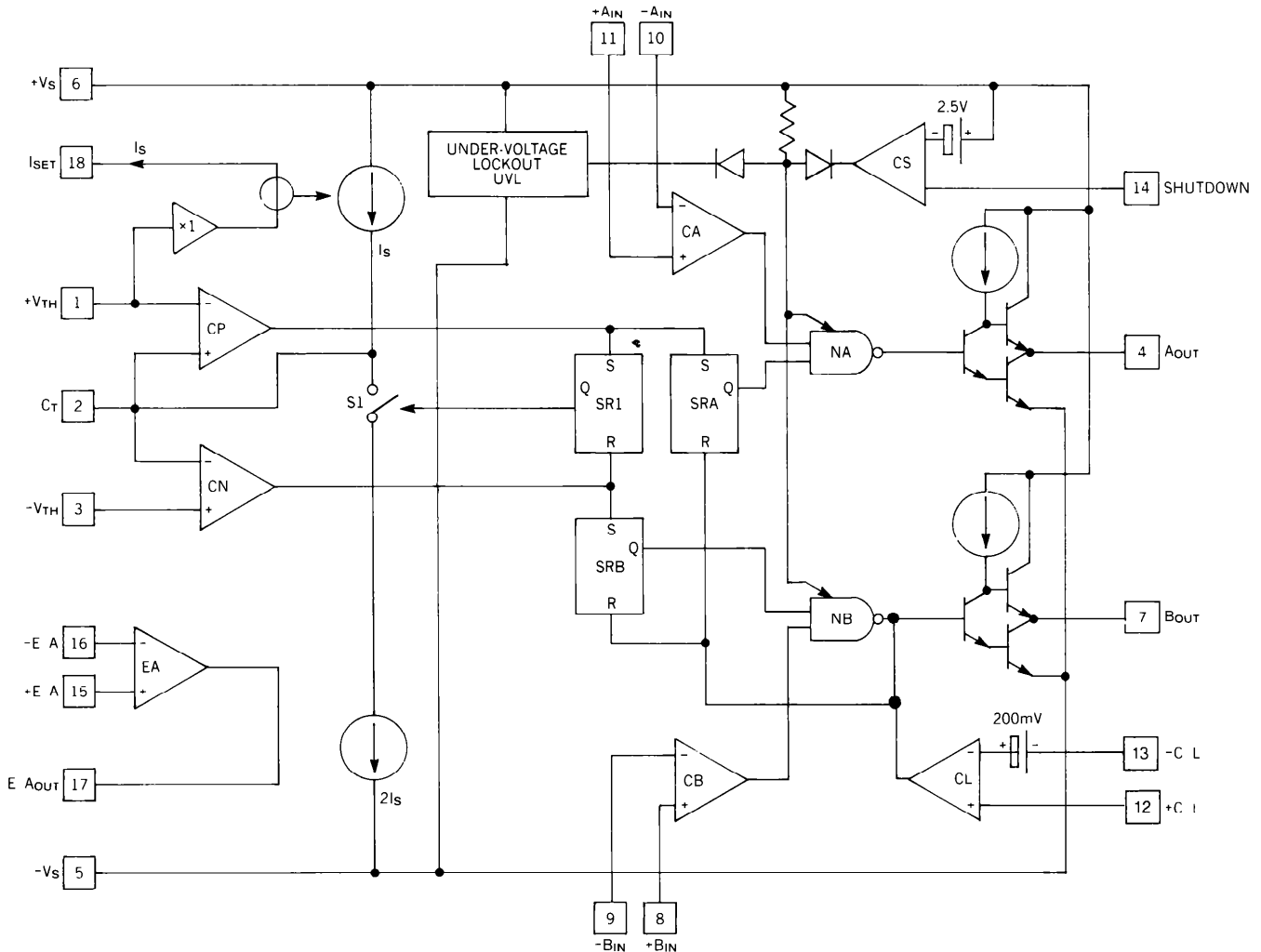

FIGURE 1. BLOCK DIAGRAM OF UC1637.

PULSE WIDTH MODULATION (PWM)

The function of a power amplifier is to regulate the flow of energy from a power supply to a load, under the control of an input signal. A linear amplifier does this by interposing a controlled voltage drop in series with the load, while carrying the full load current. The product of this voltage and current represents the amount of power that must be dissipated by the amplifier itself, and it is easy to see that the method is not very efficient. In fact, its usefulness diminishes rapidly as the amount of power to be controlled increases and, at some point, a more efficient method becomes imperative.

PWM is a switching technique in which the supply voltage is fully applied (switched) to the load and then removed, the "on" and "off" times being precisely controlled. The effect on the load is the same as if some lower voltage were continuously applied whose value depended on the duty-cycle, that is, the ratio of "on" time to the full switching period. Since supply current only flows during the "on" times, it is apparent that the efficiency should be much higher than in the linear amplifier, as in fact it is. Still, switching transistors have small but finite "on" voltages and transition times, all of which introduce losses, which limit practical PWM efficiencies to something between 75% and 90%.

THE UC1637

The diagram of Figure 1 shows in block form the internal organization of the device. The main functions are:

A) Triangular wave generator; CP, CN, S1, SR1
B) PWM comparators; CA, CB
C) Output control gates; NA, NB
D) Current limit; CL, SRA, SRB
E) Error amplifier; EA
F) Shutdown comparator: CS
G) Undervoltage lockout; UVL

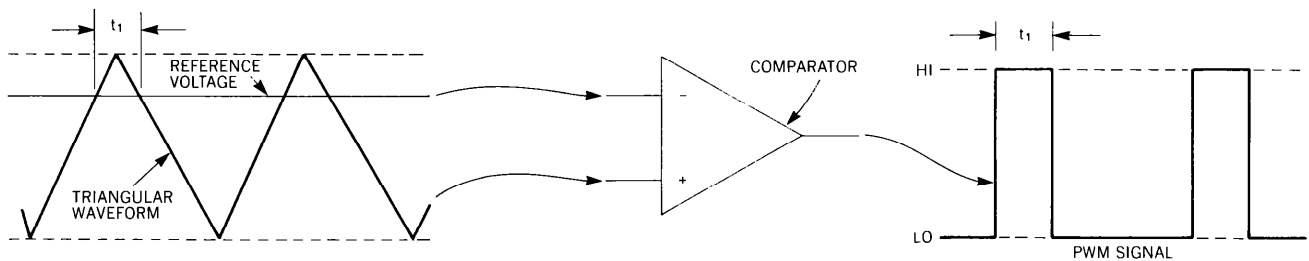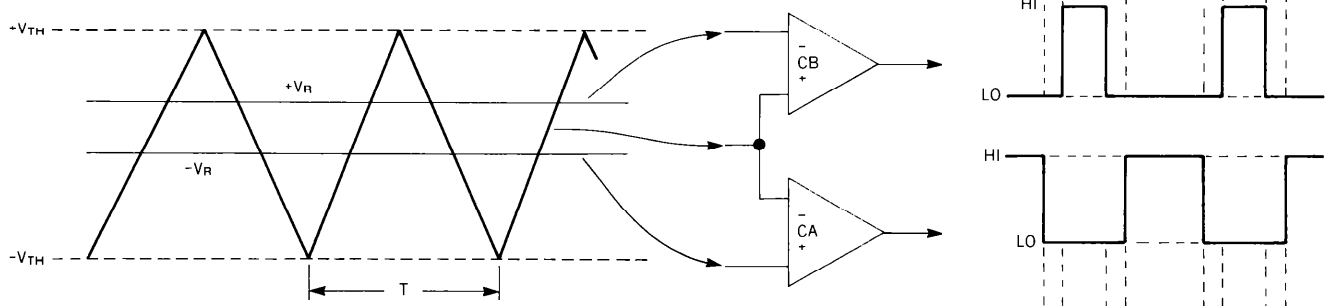The two output lines, A_{OUT} and B_{OUT}, are meant to drive the two legs of an H-bridge power amplifier, with the load driven in bipolar fashion. The A_{OUT} and B_{OUT} outputs themselves are rated at 500mA peak and 100mA continuous, which makes it easy to interface the device with most amplifiers.

In order to generate two PWM output signals, we first produce a triangular waveform, or linear ramp. This is done by charging a capacitor C_T (pin 2) with constant current I_S until the comparator CP, with a fixed threshold voltage of $+V_{TH}$, delivers a pulse to "set" the SR1 latch circuit. This forces Q high, which closes the switch S1 and adds a negative current $2 \times I_S$, to the node of pin 2. As a result, a net current equal to I_S now flows out of C_T, discharging it linearly until the comparator CN resets SR1, and the cycle restarts. Thus, the voltage at pin 2 ramps continuously between $-V_{TH}$ and $+V_{TH}$ at a frequency that depends on these two threshold voltages, on C_T, and on Is.

The current I_S is programmed by means of a resistor connected to pin 18. The voltage at this pin is equal to $+V_{TH}$ and an internal current mirror forces the charging current I_S to be equal to the current flowing out of pin 18. If a resistor R_S is connected from pin 18 to $-V_S$ (pin 5) instead of to ground, the ramp frequency becomes independent of power supply voltage variations, since Is will then change together with V_{TH}.

As Figure 2 shows a triangular waveform can be compared with a reference voltage to generate a PWM signal. The UC1637 uses two separate comparators to generate the two output signals A_{OUT} and B_{OUT}. The way the signals are handled, and the results, are shown in Figure 3 where it can be seen that the difference between V_A and V_B is the cause of the time intervals during which both outputs are low.

FIGURE 2. HOW A PWM SIGNAL IS GENERATED.

FIGURE 3. TWO PWM SIGNALS ARE GENERATED IN THE UC1637.

The two nand gates, NA and NB, will be enabled if the following two conditions are met:

A) supply voltage +V_S is greater than +4.15 volts (typ)
B) the shut-down input line (pin 14) is at least 2.5 volts (typ) negative with respect to +V_S.

If these are satisfied, the A_{OUT} output line will be high if the CA output *and* Q of SRA are both high. Since SRA is set at each positive peak of the oscillator ramp, the output A_{OUT} can be controlled by CA singly — as long as a current-limit pulse from CL does not occur. The operation of the NB gate is similar.

The timing diagrams of Fig. 4 show the sequence of events before and after a current limit pulse occurs. Before time t_1 the PWM action is smoothly controlled by the ramp comparisons with V_A and V_B. The pulse from CL at time t_1 resets both SRA and SRB; the output lines are now disabled until SRA is set (at time t_2) and SRB is set (at time t_3).

The current limit comparator CL provides a means to protect both driver and motor from the consequences of very high currents. If the current delivered by the driver to the motor is made to flow through a low value resistor (for example, see R_S in Figure 7) the voltage drop across this resistor will be a measure of motor current. This voltage is applied between pins 12 and 13 of the UC1637, with pin 12 positive. A 200mV threshold is provided internally (see Figure 1) so that when the R_S voltage is equal to 200mV the output of CA goes high, resetting both SRA and SRB and, consequently, terminating any active output pulse. This pulse-by-pulse method of current limiting is very fast and provides effective protection, not only for the driver components, but also for the motor where the possibility of demagnetization due to excessive current is a matter of serious concern.

Finally, the UC1637 contains also an operational amplifier EA, that can be used to provide gain and phase compensation as will be seen later.
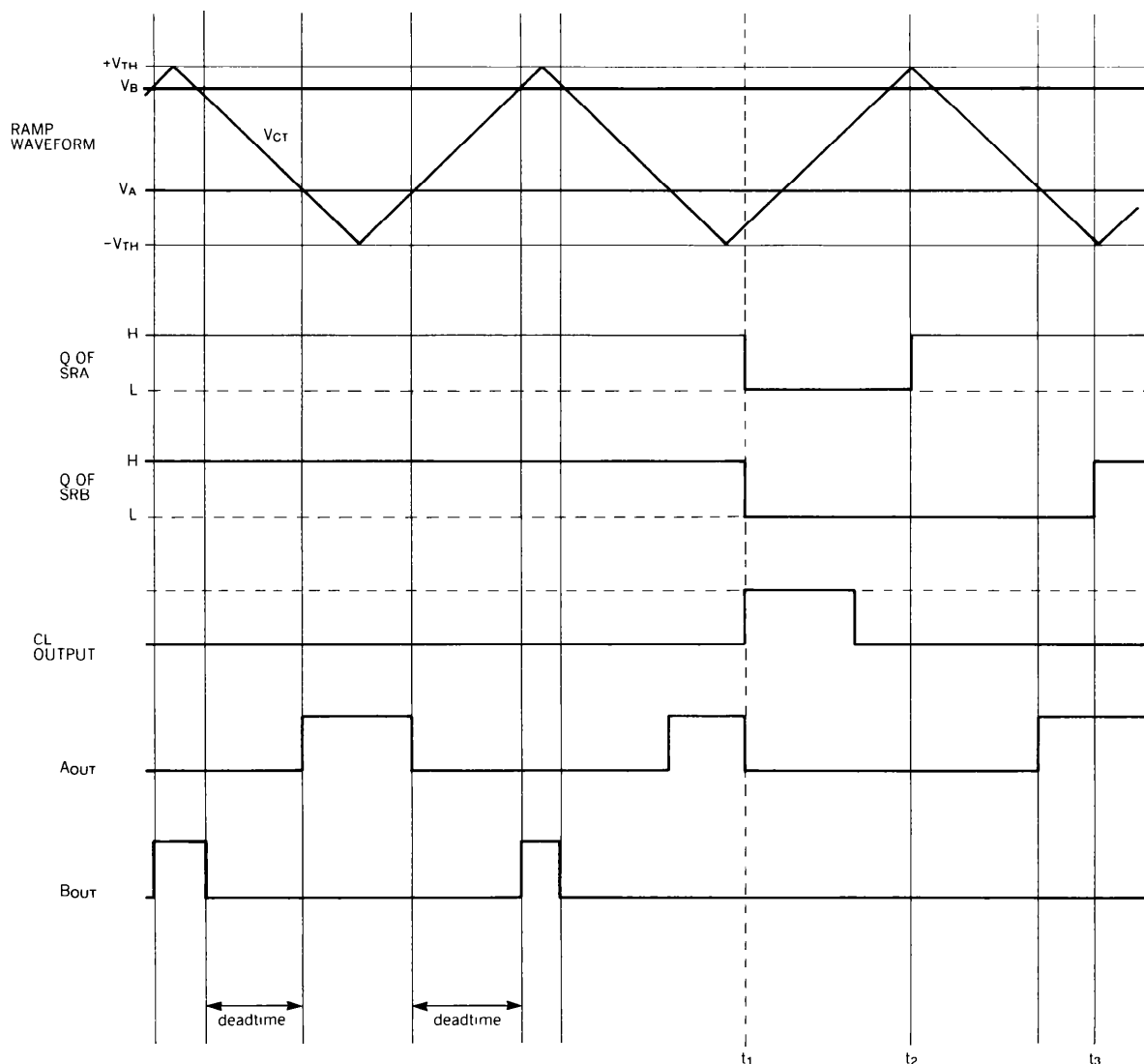

FIGURE 4. TIMING DIAGRAM SHOWING THE GENERATION OF PWM PULSES AT A_{OUT} AND B_{OUT}.
BEFORE TIME t_1, THE Q OUTPUTS OF SRA AND SRB ARE BOTH HIGH AND THE OUTPUT PULSES ARE CONTROLLED BY THE RAMP INTERSECTIONS WITH V_A AND V_B. AT TIME t_1, THE CURRENT LIMIT COMPARATOR HAS SENSED EXCESS CURRENT AND THE CL OUTPUT HAS GONE HIGH, RESETTING BOTH SRA AND SRB. THIS TERMINATES THE A_{OUT} PULSE THAT WAS ACTIVE AT THE TIME. A_{OUT} CAN RESUME ONLY AFTER SRA IS SET AT t_2; B_{OUT} CAN RESUME ONLY AFTER SRB IS SET AT t_3.

Figure 5 shows the connections needed to get the ramp generator and the two comparators ready to go. There is no great difficulty in calculating values for the various resistors, which are no more than two simple voltage dividers. Still, certain things should be considered before proceeding.

The input impedance R_{IN}, seen by the control voltage V_C will be

$$R_{IN} = \frac{R_3 + R_4}{2} \tag{1}$$

and this value may be specified or determined in advance. Also, it would be economical to have a minimum number of different values of resistors. If we make

$$R_1 = R_3 \tag{2}$$

we will have four resistors of equal value in the final circuit.

There is also the question of deciding on the separation V_G between the reference voltages $+V_R$ and $-V_R$. The voltage gain of the PWM amplifier will have one of the four characteristics depicted in Figure 6, depending on your choice of reference voltage separation. You can get a linear response by making $V_G = 0$, as in Curve #1, or by making $V_B - V_A = 2V_{TH}$, as in Curve #3. In Curve #2, there is a change in slope due to the contribution, near zero, of both V_A and V_B to the output changes, which in some systems may be undesirable, but which may be of interest due to the fact that it results in zero losses at null.
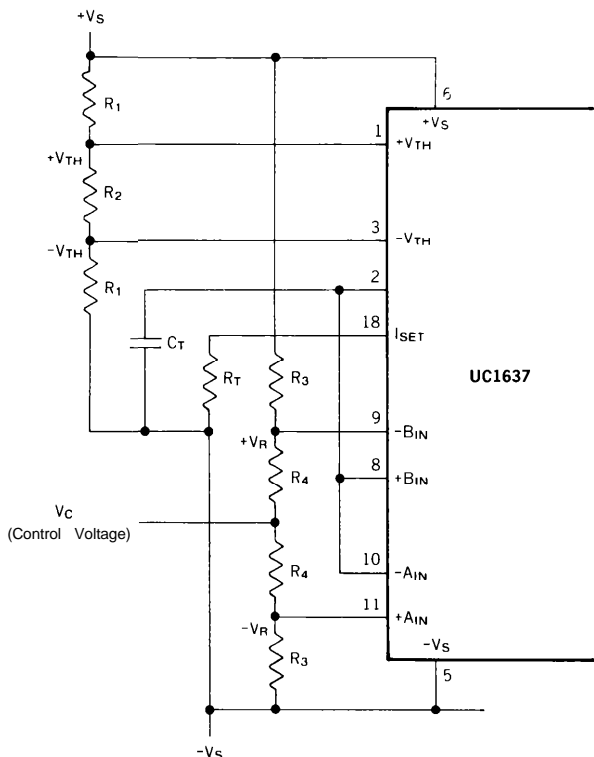

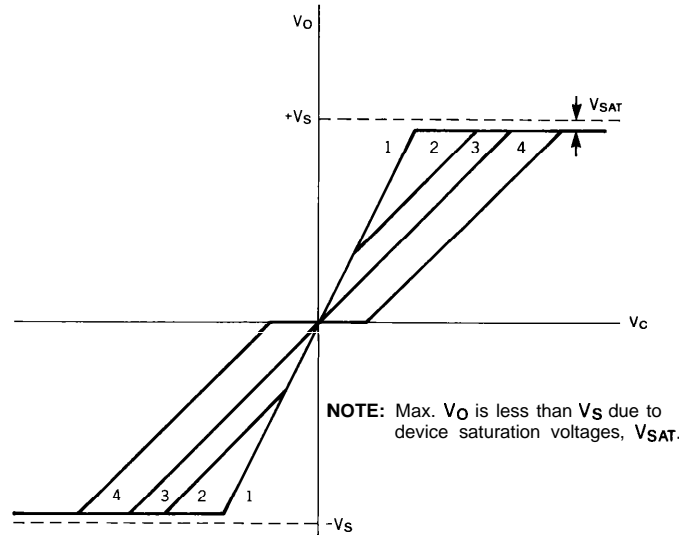**NOTE:** Max. V_O is less than V_S due to device saturation voltages, V_{SAT}.

FIGURE 6. PWM VOLTAGE GAIN CHARACTERISTICS OBTAINABLE WITH VARIOUS VALUES OF REFERENCE VOLTAGE SEPARATION, OR GAP VOLTAGE 2 V_R.
1. LINEAR GAIN WITH $V_R = 0$ (a = 0).
2. NON-LINEAR GAIN WITH V_R GREATER THAN ZERO BUT LESS THAN V_{TH} (0 < a < 1).
3. LINEAR GAIN WITH $V_R = V_{TH}$ (a = 1).
4. NON-LINEAR GAIN WITH V_R GREATER THAN V_{TH} (a > 1).
NOTE: THE SLOPE OF LINE 1 IS TWICE THAT OF LINE 3.

At this point, this choice of PWM gain characteristic amounts only to the choice of the ratio between V_R and V_{TH}:

$$a = \frac{V_R}{V_{TH}} \tag{3}$$

The values of V_{TH} and V_R, as well as R_3 and R_4, depend on the following:

$\pm V_S$: power supply voltages
R_{IN}: desired control input resistance
$V_C$max: peak value or input voltage V_C. This is the input voltage at which the output reaches 100% duty cycle
a: ratio of V_R to V_{TH}

These values being known the designer can proceed to calculate the following circuit values:

FIGURE 5. SETTING UP THE A AND B COMPARATOR INPUTS.

$$R_3 = \frac{2 R_{IN} V_S \left(1 + \dfrac{1}{a}\right)}{V_C max + V_S \left(1 + \dfrac{1}{a}\right)} \tag{4}$$

$$R_4 = 2 R_{IN} - R_3 \tag{5}$$

$$V_A = \frac{V_S R_4}{2 R_{IN}} \tag{6}$$

$$V_{TH} = \frac{V_R}{a} \tag{7}$$

$$R_2 = 2 R_3 \frac{V_{TH}}{V_S - V_{TH}} \tag{8}$$

and, from Eq. (2), $R_1 = R_3$.

Having chosen a frequency f_T for the PWM timing circuit, you can now calculate C_T and R_T. A suitable starting value for the charging current I_S is 0.5mA which gives

$$R_T = \frac{V_S + V_{TH}}{.0005} \tag{9}$$

$$C_T = \frac{.0005}{4f_T V_{TH}} \tag{10}$$

You will probably need to make an adjustment here, so as to get a standard value for capacitor C_T, and it is best to keep I_S in the range from 0.3mA to 0.5mA when you do this. It may be desirable, or even necessary in some conditions,

to bypass the $+V_{TH}$ and $-V_{TH}$ inputs to ground, and for this, ceramic capacitors of $0.1\mu f$ should be adequate. Remember also that terminal 14, the shut-down line, must be held "low" (at least 2.5V below the positive rail) in order to enable the drive. With an external switch to ground, or to $-V_S$, and a pull-up resistor to $+V_S$, this line can be used to enable (low), and disable (high), the output. Both A_{OUT} and B_{OUT} will be low when the shut-down line is high.

The next step is to connect the UC1637 to a suitable power amplifier, and the amplifier to the motor. The UC1637 has provisions for current limiting, as discussed earlier, and you must make arrangements to develop a voltage proportional to motor current at the driver side. This can be done by adding to an H-bridge a low value resistor in series with rail connections. The current limit comparator has a common mode range that reaches all the way down to the negative rail (on the positive side the limit is 3V below the positive rail). A resistor R_S is then added at the bottom of the bridge, and its value is selected so as to give a voltage drop to 200mV when the desired limit current flows.

$$R_S = \frac{.2}{I_{MAX}} \quad \text{(ohms)} \tag{11}$$

where I_{MAX} is the maximum desired motor current in amperes. In a breadboard, a twisted pair of wires should be used to make the connection from this resistor to pins 12 and 13, and an RC filter should be added, as shown in Figure 7.

On a PC board, it is a good idea to keep R_S close to the UC1637 to minimize the length of the connecting traces. The RC filter should still be used.
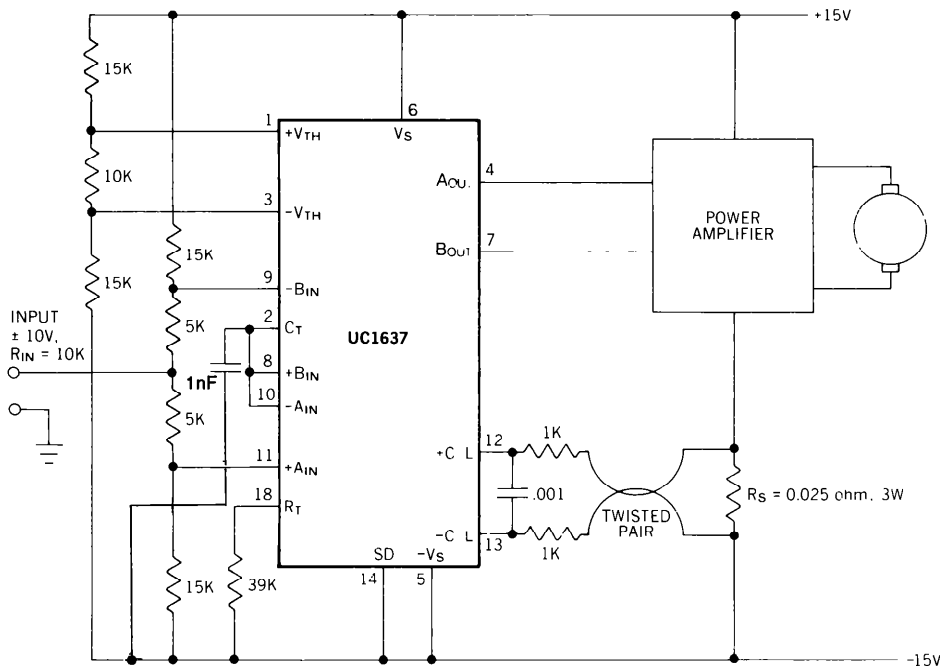

FIGURE 7. CIRCUIT DIAGRAM OF PWM VOLTAGE AMPLIFIER WITH GAIN OF 3.

AN EXAMPLE

We are ready now to design a current limited, PWM voltage amplifier to drive a small DC servomotor. Here are the requirements:

> Supply voltages: ±15V
> Input: ±10V max.; 10K input res.
> PWM frequency: 30KHz
> Motor current limited at 8A
> Minimum power losses at idle

> We have:
> $V_S = 15V$
> $V_C max = 10V$
> $R_{IN} = 10^4 ohm$
> $f_T = 3 \times 10^4$ Hz
> and $I_{MAX} = 8A$

and also, from the last requirement, a = 1.

FROM EQUATIONS

(4) $\quad R_3 = \dfrac{2 \times 10^4 \times 15 \times 2}{10 + 15 \times 2} = 15K$

(5) $\quad R_4 = 2 \times 10^4 - 15 \times 10^3 = 5K$

(6) $\quad V_R = \dfrac{15 \times 5 \times 10^3}{2 \times 10^4} = 3.75V$

(7) $\quad V_{TH} = 3.75V$

(8) $\quad R_2 = (2 \times 15 \times 10^3) \times \dfrac{3.75}{15 - 3.75} = 10K$

(9) $\quad R_T = \dfrac{15 + 3.75}{.0005} = 37.5K$

and of course, $R_1 = R_3 = 15K$.

(10) $\quad C_T = \dfrac{.0005}{4 \times 30 \times 10^3 \times 3.75} = 1.11 \times 10^{-9} fd$

If we settle for $R_T = 39K$, I_S becomes slightly less than 0.5mA and if we then pick $C_T = 1000pf$, the nominal frequency becomes 32KHz.

To limit the motor current at 8A, we need, from Eq. 11,

$$R_S = \dfrac{.2}{8} = 0.025 \text{ ohm}$$

The peak power in the resistor will be

$$P_S = 8^2 x .025 = 1.6 \text{ watts.}$$

Incidentally, the voltage gain of the amplifier can be determined from the fact that a 10V change at the input results in a 30V change at the output; therefore, the gain from input to motor terminals is 3. The above circuit is shown in Figure 7.

THE POWER AMPLIFIER

Where space is tight and motor current is less than five amperes, the Unitrode PIC900 offers a perfect solution to your power bridge design. This device comes in a DIL-18 package, requires only 5mA of input drive current, and is rated at 5A absolute maximum output current. It contains all you need for the output H-bridge - including the circulating diodes - and with only a few added parts, you are ready to go. A circuit diagram showing a velocity feedback loop using one UC1637 and one PIC900 appears in the UC1637 data sheet.

For higher currents, you will have to design your own amplifier, and for the purposes of this application note, a sample design is shown in Figure 8. Referring to that circuit, note that with $+V_S$ and $-V_S$ applied, if the inputs are left open, the power MOSFETs are all "off". If Drive A, for example, is driven to within 3.6V of either power rail, then the corresponding output is switched to that rail. Note that since the PNP and NPN junction transistors are by nature faster switching "on" than "off", while the MOSFETs are much faster than the junction transistors driving them, this connection provides a simple guarantee against cross-conduction. Also working toward this goal is the fact that the junction transistor can discharge the MOSFET's input capacitance faster than the 1K, 1W resistor can charge it. The arrangement shown in Figure 8 results in a transition time of about $1.5\mu S$ during which both MOSFETs in a given leg are off. This amount of time is a very small portion of the $33\mu S$ period toward which we are designing our example.

The power MOSFET transistors, in TO-220 package, are rated at 60V and 12A. The channel "on" resistance is quite low, 0.25 ohms at 8A, for the UFN533, resulting in low thermal losses. You can easily find other devices with even lower R_{DS} values, if needed, but as always, the price you pay is that you must pay the price.

Finally, a word about circulating diodes - conspicuous in Figure 8 by their absence. All power MOSFETs have an intrinsic rectifier, or body diode, a junction rectifier whose current rating is the same as that of the transistor. With the drive format provided by the UC1637, the two bottom MOSFETs (N-channel) are "on" during the time when motor current circulates, and as a result the reversed diode carries only a small portion of the current: most of it flows from source to drain through the channel. In fact, the diode fully conducts only during the $1.5\mu S$ when both devices in one bridge leg are off. You can add fast recovery diodes in shunt with the MOSFETs if you find that they are essential. The intrinsic MOSFET diode is not particularly fast, and as your output current requirements increase, the need for fast external diodes will become more and more apparent.
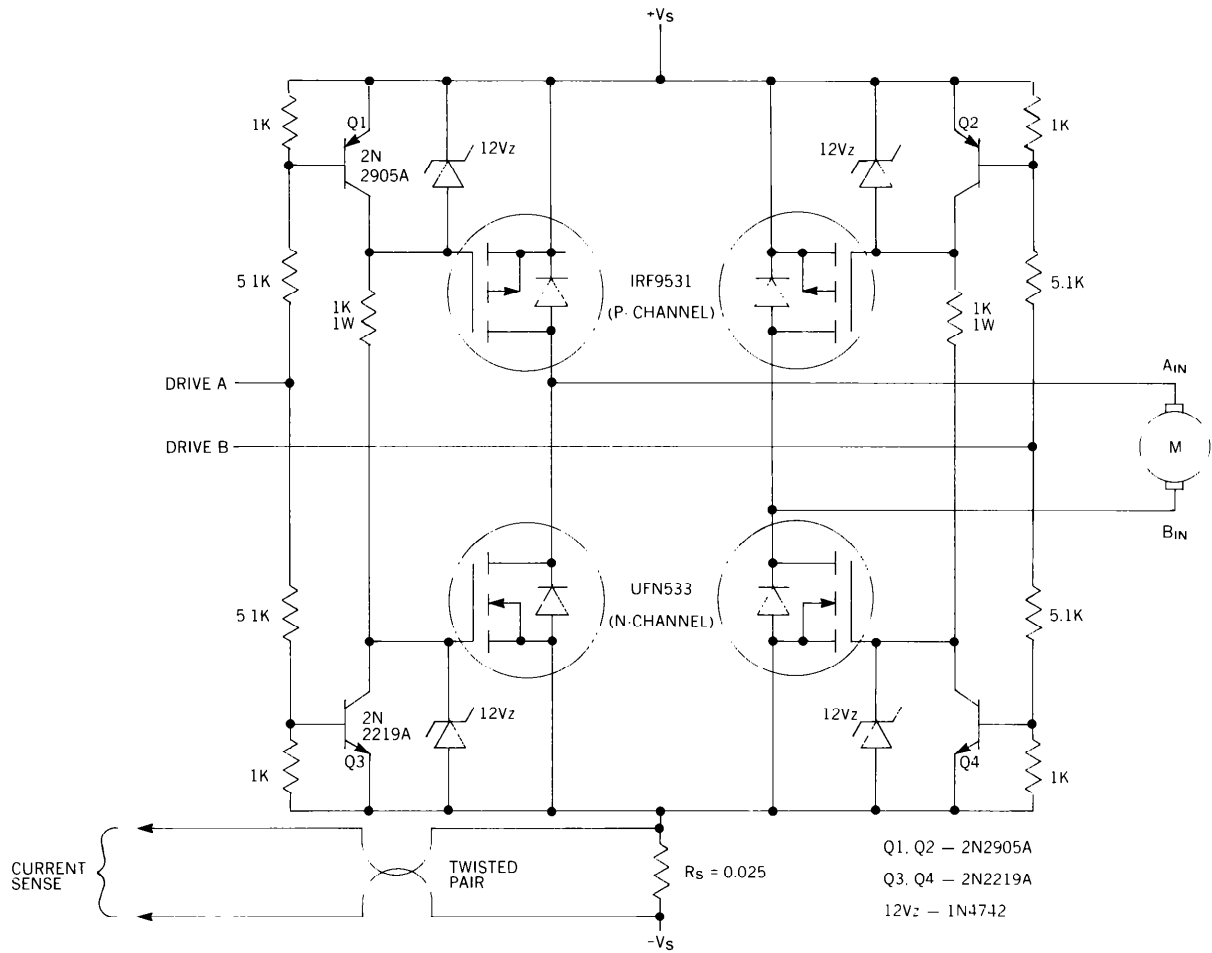
FIGURE 8. THIS 8A POWER AMPLIFIER IS SUITABLE FOR 30KHz OPERATION.

THE SERVOMOTOR

It is convenient to represent the DC servomotor by a simple equivalent circuit, and one such circuit is shown in Figure 9. Note that by expressing the moment of inertia J and the motor constant K in metric units (Nm sec^2 and Nm/A respectively), we avoid the need to include a multiplying constant in the expressions for C_M and e_O. Also, the motor constant K, in metric units, defines both the voltage con-

stant in volt-sec/rad, and the torque constant in Nm/A, as one and the same number.

The ratio J/K^2 has the dimensions of capacitance, with a value runnng to several thousand microfarads. The voltage across this capacitor is equal to $K\omega$ where ω is the angular velocity of the rotor in rad/sec. Consequently, this voltage is the analog of shaft velocity.

Our equivalent circuit, then, is a simple series connection of R_A, the armature resistance; L_A, the armature inductance; and C_M, the equivalent capacitance, equal to J/K^2. It should come as no surprise that such a circuit will have a natural resonant frequency ω_N, and a resonant Q as well. This is indeed the case, and we have for its transfer function,
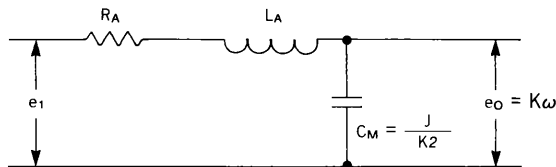

FIGURE 9. EQUIVALENT CIRCUIT OF MOTOR, WHERE J IS THE TOTAL MOMENT OF INERTIA OF ROTOR PLUS LOAD.

R_A = armature resistance; ohms.
L_A = armature inductance; henrys.
C_M = equivalent capacitance; farads.
J = total moment of inertia: Nm sec^2.
K = motor constant; volt sec/rad, or Nm/A.
ω = rotor angular velocity: rad/sec.

TO CONVERT FROM	TO	MULTIPLY BY
oz in sec^2	Nm sec^2	7.06×10^{-3}
volts/KRPM	volt sec/rad	9.55×10^{-3}

$$\frac{e_O(s)}{e_{1(s)}} = \frac{1}{(s/\omega_N)^2 + s/Q\omega_N + 1} \tag{12}$$

$$\text{where} \quad \omega_N = \sqrt{\frac{K}{L_A J}} \tag{13}$$

$$\text{and} \quad Q = \frac{K}{R_A} \sqrt{\frac{L_A}{J}} \tag{14}$$

We can now use these sample results in our sample design. Here are some of the data given by a motor manufacturer:

EG & G TORQUE SYSTEMS
MODEL NO. MT-2605-102CE
(motor - tach assembly)

MOTOR: K_T = 4.7 oz in/amp
$\quad\quad K_V$ = 3.5V/KRPM
$\quad\quad R_A$ = 0.7 ohms
$\quad\quad J_M$ = 0.0018 oz in sec^2
$\quad\quad T_M$ = 8.6 ms (mech. time const.)
$\quad\quad T_e$ = 1.6 ms (el. time const.)
TACH: J_T = 0.001 oz in sec^2
$\quad\quad K_V$ = 3V/KRPM

The several motors in this series and size have the same electrical time constant T_E, and since we know R_A,
$$L_A = T_E R_A = 0.016 \times 0.7$$
$$L_A = 1.12 \text{ mH}$$

The total moment of inertia is
$$J = J_M + J_T = 0.0018 + 0.001$$
$$J = 0.0028 \text{ oz in sec}^2$$

In metric units,

$$J = \frac{0.0028}{141.612} \text{ (Nm sec}^2)$$

Putting K_T in metric units,

$$K = \frac{4.7}{141.612} \text{ (Nm/amp)}$$

The equivalent capacitance is

$$C_M = \frac{J}{K^2} = \frac{141.612 \times 0.0028}{(4.7)^2} = 18,000\mu f$$

For the equivalent circuit, then, the values are

$$R_A = 0.7 \text{ ohms}$$
$$L_A = 1.12 \text{ mH}$$
$$C_M = 18,000\mu f$$

The angular velocity will be proportional to the voltage e_O across C_M;
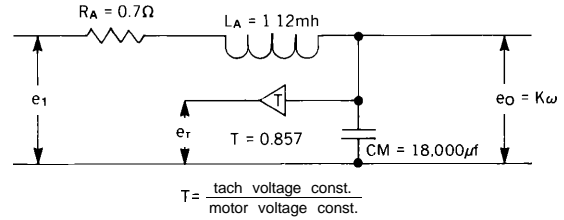
$$\omega = \frac{e_O}{K}$$

FIGURE 10. THE TACH VOLTAGE e_T IS PROPORTIONAL TO ω.

If the motor has a tachometer attached, we can include it in the equivalent circuit by deriving an equivalent tach voltage proportional to e_O. This is illustrated in Figure 10, where

$$T = \frac{\text{Tach. voltage constant}}{\text{Motor voltage constant}}$$

$$T = \frac{3V/KRPM}{3.5V/KRPM} = .857$$

From Eq. 13, $\quad\quad \omega_N$ = 222.7 rad/sec

From Eq. 14, $\quad\quad Q = 0.356$

(Note: Since $\zeta = \frac{1}{2Q}$, the damping factor here is 1.4)

From Eq. 12 and the above data, we can write the ratio of tach voltage to input as

$$\frac{e_T(s)}{e_1(s)} = \frac{.857}{\left(\dfrac{s}{222.7}\right)^2 + \dfrac{s}{79.3} + 1} \tag{15}$$

THE VELOCITY LOOP

Our objective is to put together a feedback loop using our UC1637, H-bridge, and motor: the controlled variable is ω, the motor shaft's angular velocity. For high accuracy, we need a high loop gain, so that small velocity errors are magnified and corrected. The UC1637 internal ERROR amplifier is appropriate for this purpose, and will be used as a summing amplifier. But before proceeding, let us take a look at Figure 11, where a plot of the motor-tach transfer function (Eq. 17) is shown. The plot shows that as the frequency increases, the tach output decreases and the phase lag increases towards a maximum of 180°. This means that although we can introduce plenty of gain at very low frequencies, where the phase lag is low, the added gain must be reduced at the higher frequencies, where the 180° phase lag tends to make our loop a regenerative one. If we want the closed loop response to be "snappy", that is, if we want a bandwidth of several tens of hertz, then the loop gain must be
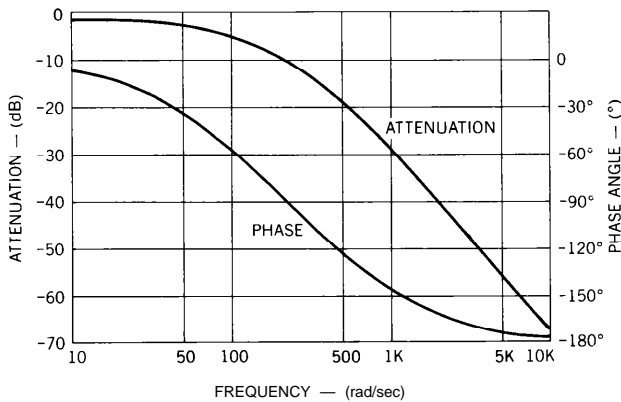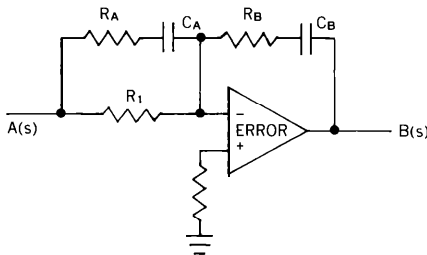
FIGURE 11. PLOT OF MAGNITUDE AND ANGLE OF EQ. 15. WHICH DESCRIBES PERFORMANCE OF OUR TEST MOTOR.

fairly high at all frequencies in the band; yet, for flat response and fast step response with no overshoot we must make certain that the overall phase shift is less than 180° at any frequency at which the gain is greater than unity.

$$\frac{A(s)}{B(s)} = \frac{(1 + sR_BC_B)\,[1 + s(R_1 + R_A)\,C_A]}{sR_1C_B\,(1 + sR_AC_A)}$$

FIGURE 12. ERROR AMPLIFIER WITH ITS FREQUENCY COMPENSATION NETWORK.
THE MAGNITUDE AND ARGUMENT OF THE TRANSFER FUNCTION CAN BE EASILY PLOTTED WITH THE AID OF A PROGRAMMABLE CALCULATOR.

The high gain ERROR amplifier of the UC1637, together with a few external components, is shown in Figure 12. Without R_A and C_A, the phase response of the circuit would go from -90° at low frequencies to 0° at high frequencies. This amount of phase correction is inadequate if we want a tight loop with good transient response. With R_A and C_A shunting R_1, it becomes possible to have a leading phase angle
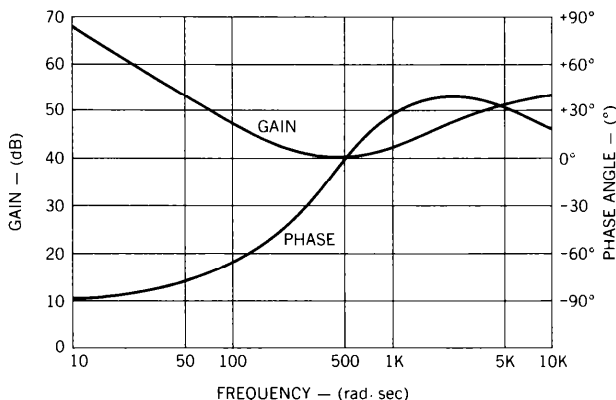

FIGURE 13. MAGNITUDE AND ANGLE OF COMPENSATION AMPLIFIER OF FIGURE 12.

somewhere at midrange, even though the high frequency asymptote is still at zero degrees (R_A and C_A introduce both a zero and a pole). The transfer function of the circuit shown in Figure 12 is plotted in Figure 13 for the following component values

$$R_1 = 9.1\,K$$
$$R_A = 1K$$
$$C_A = .22\mu f$$
$$R_B = 470K$$
$$C_B = .0047\mu f$$

The break frequencies are:

$$\frac{1}{R_BC_B} = \frac{1}{(R_1 + R_A)\,C_A} = 450\ \text{rad}/\text{sec}$$

$$\frac{1}{R_1C_B} = 23,400\ \text{rad}/\text{sec}$$

$$\frac{1}{R_AC_A} = 4,500\ \text{rad}/\text{sec}$$

The plot shown in Figure 14 shows the result of cascading the compensation amplifier, PWM amplifier, and motor-tach. All gain contributions have been simply added together, and all phase contributions have also been added. The result, shown in Figure 14, shows the open loop frequency response of the complete velocity control system.
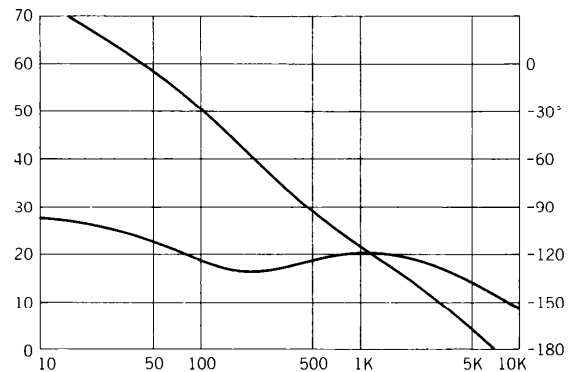

FIGURE 14. OVERALL OPEN-LOOP RESPONSE, INCLUDING +8dB DUE TO PWM AMPLIFIER GAIN AND MOTOR-TACH DC GAIN.

The inclusion of the ERROR amplifier with its compensation components has had the effect of introducing a large amount of gain at the lower frequencies, and also of reducing the phase lag at the higher frequencies. The loop gain is 0dB at about 7KHz, and the phase margin is about 40°.

Moreover, since the phase never exceeds 180°, we have the needed indication of relative stability, and can proceed to close the loop as shown in Figure 15 and make measurements. Note that a noise filter has neen added at the output of the tachometer. Such a filter is usually necessary, especially in PWM control loops of relatively wide bandwidth, because of the inevitable AC coupling between the motor signal and the tach output. In our filter, the 3dB cut-off point is at 21 KHz, which is high enough not to affect the loop behavior.
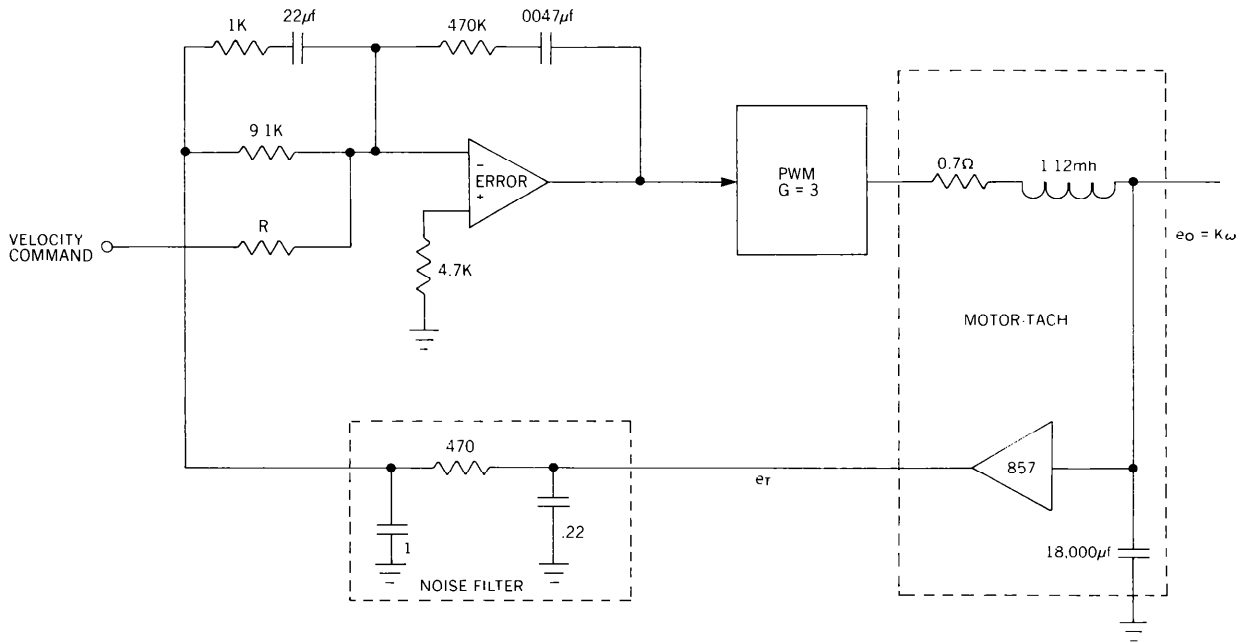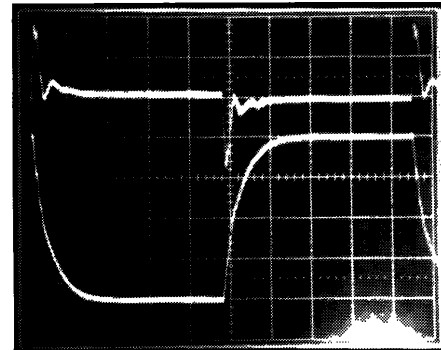
FIGURE 15. THE COMPLETE VELOCITY LOOP.

The oscilloscope trace shown in Figure 16 reveals that the step response of our loop is very well behaved. The motor shaft reaches full speed in less than 10mS, and there is no noticeable overshoot. The net velocity change in Figure 16 amounts to 133 RPM, and the current trace shows that the current does not quite reach the chosen limit of 8A. With larger input steps, the motor accelerates at constant 8A current, and the acceleration rate is approximately 100RPM per millisecond. The 3dB bandwidth of the loop measured about 80Hz.

CONCLUSIONS

We have discussed in some detail the characteristics of Unitrode's UC1637 and have presented in detail a design approach which illustrates those points. The sample design was built and tested, with the measured results as presented above. These results show that excellent performance can be obtained with few components, and that the design technique is quite simple. Our velocity loop would perform well as an inner loop in a position control system, for example, although a different response might perhaps be desirable. However that may be, using the UC1637 a sizable portion to the job is completed beforehand.

Top trace: 5A/cm
Bottom trace: 100mV/cm
Horizontal: 5msec/cm

FIGURE 16. STEP RESPONSE OF THE VELOCITY CONTROL LOOP OF FIGURE 15. THE UPPER TRACE SHOWS THE MOTOR CURRENT: THE LOWER TRACE SHOWS THE TACH OUTPUT VOLTAGE, I.E., MOTOR VELOCITY.
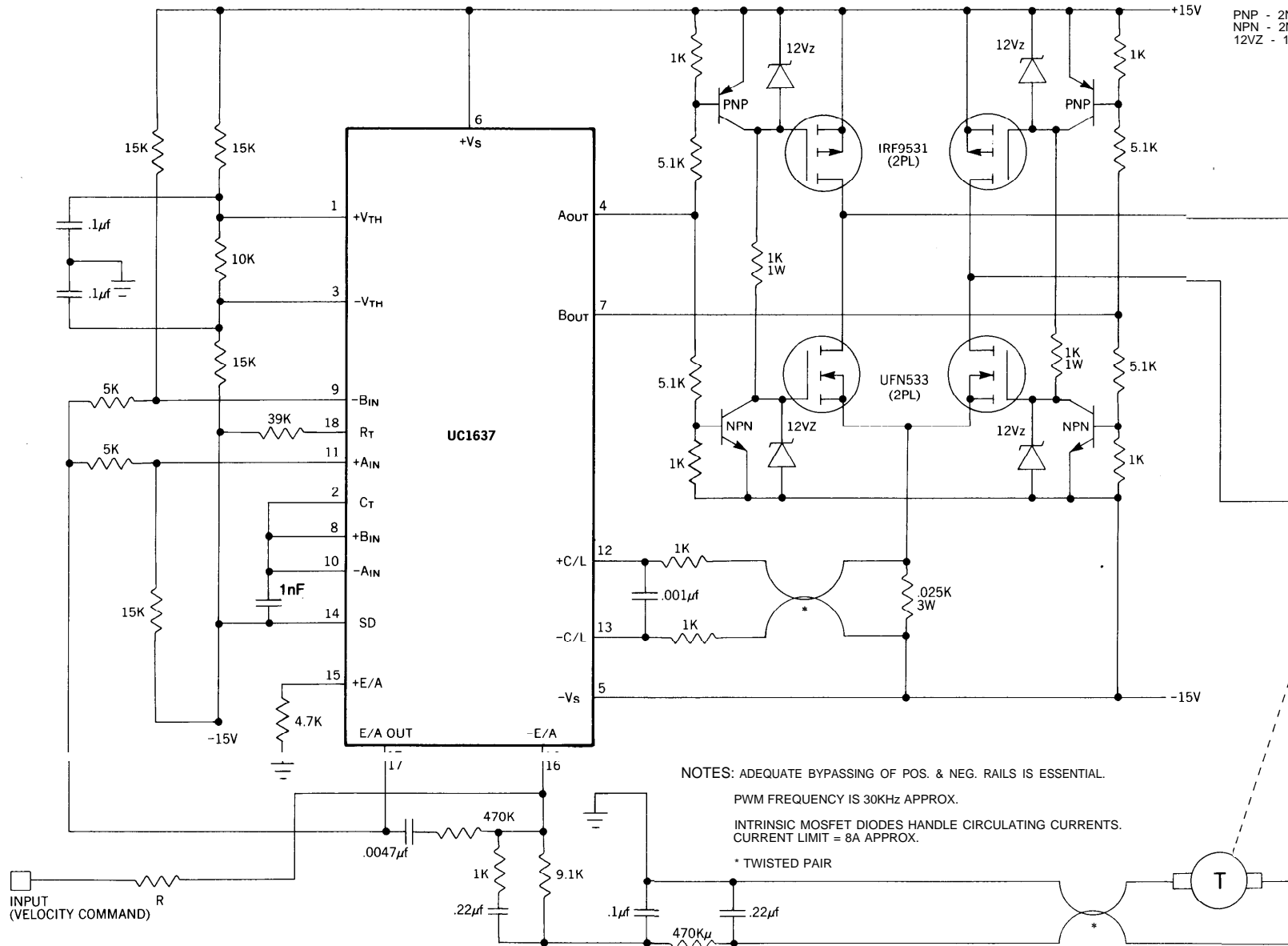
See Figure 17.

ACKNOWLEDGMENTS

PNP - 2N2905A, OR EQUIVALENT
NPN - 2N2219A, OR EQUIVALENT
12VZ - 1N4742, OR EQUIVALENT

UC1637

IRF9531
(2PL)

UFN533
(2PL)

MOTOR-TACH:
EG & G TORQUE SYSTEMS
MODEL MT-2605-102CE

1 +V$_{TH}$
3 -V$_{TH}$
9 -B$_{IN}$
18 R$_T$
11 +A$_{IN}$
2 C$_T$
8 +B$_{IN}$
10 -A$_{IN}$
14 SD
15 +E/A
E/A OUT   -E/A
6 +V$_S$
4 A$_{OUT}$
7 B$_{OUT}$
12 +C/L
13 -C/L
5 -V$_S$

INPUT
(VELOCITY COMMAND)

NOTES: ADEQUATE BYPASSING OF POS. & NEG. RAILS IS ESSENTIAL.

PWM FREQUENCY IS 30KHz APPROX.

INTRINSIC MOSFET DIODES HANDLE CIRCULATING CURRENTS.
CURRENT LIMIT = 8A APPROX.

* TWISTED PAIR

SELECT VALUE OF R FOR DESIRED GAIN.

WITH R = 9.1K, GAIN WILL BE 333.3 RPM PER VOLT.

FIGURE 17. COMPLETE VELOCITY CONTROL LOOP OF SAMPLE DESIGN.

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