Application Note

I C s   f o r   M o t o r C o n t r o l

Evaluation board for the TDA5143/TDA5144

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Summary:

This paper describes the evaluation board for the Motor Control ICs TDA5143/5144. These ICs have been packaged in two different packages, viz.: SO20 and SO28.
The TDA5143T and TDA5144AT are packaged in an SO20 package and the TDA5144T in an SO28. There is one evaluation board for the SO20 package and one for the SO28 package.

Both boards have an option for two different adjustable current limiters. However, the components needed for the currents limiters have not been mounted. The users themselves have to add the components for one of the two current limiters.

In this report the calculation for the different key components is given as well.

Table of Contents:

1. Introduction.

In this report an application circuit for the TDA5144 is discussed, including the calculation of some key components. The circuit also has an option for a current limiter.
The current limiter option has been realized in two different ways: a simple one-transistor controlled current limiter and an op-amp controlled current limiter.

The complete circuit is built on a pc-board and contains both current limiter options. The selection is made by either mounting the components for the one-transistor current limiter option or the components for the op-amp controlled current limiter on the pc-board. The pc-board will be supplied without the components for any current limiter. In the parts list has been indicated which components for a certain current limiter have to be inserted.
The place for the different components has been indicated on the pc-board and in figure 9 and 10.

The Motor Control IC needs some additional discrete components for timing and speed control . In this report guidelines and formulas have been given to determine these components.

2. Circuit description.

In figure 1 the circuit has been given for the motor drive circuit. It contains all the components to drive a brushless DC motor up to a supply voltage of 16 V.

The board has two supply voltage terminals. The V_p only supplies the control circuitry of the TDA5143/5144 and the supply voltage V_s supplies the power for the motor.
Furthermore the board has an input that is called V_{in}. With the voltage on this input the motor voltage V_{mot} can be controlled. The control of this input will be discussed later on in this report.
The Fg pin (frequency generator) outputs a signal with a frequency that is half the motor commutation frequency.

The pin "TEST" is not used for normal operation and is only meant for test purposes.

Because not every motor has been provided with an external centre tap connection (MOT0) the evaluation board has been equipped with 3 sets of solder pins. Between these solder points three resistors (R17, R18, R19) of 1 KΩ can be mounted to create an artificial centre tap.

In figure 2 the circuit has been depicted with a one-transistor controlled current limiter and in figure 3 the circuit with an op-amp controlled current limiter.
The current limiter uses a resistor to measure the motor current. The one-transistor current needs a voltage drop of about 1 V before the limiter becomes active. The op-amp controlled current limiter can become active at a voltage drop of about 0.1 V. A low voltage drop increases the total efficiency of the motor driver.

For the pcb lay-out the three circuits have been merged to obtain one lay-out for the three circuit options. This gives the opportunity to use one pc-board design for an evaluation board without a current limiter or an evaluation board with a current limiter.

Both current limiters have been equipped with a current sense resistor R7. Because the circuit of figure 1 needs no current sense resistor R7 a wire jumper has been used mounted instead.

The diagrams have been drawn for the TDA5144AT (SO20), but the design can of course also be used for the TDA5143T in SO20. The TDA5143T has the same pin-out as the TDA5144AT; is functionally compatible, but has lower output capabilities.
The TDA5144T is functionally compatible to the TDA5144AT but has a different package and pin lay-out. It was not possible to use one pc-board lay-out for both types of IC. This means a separate pc-board design for the TDA5144T.

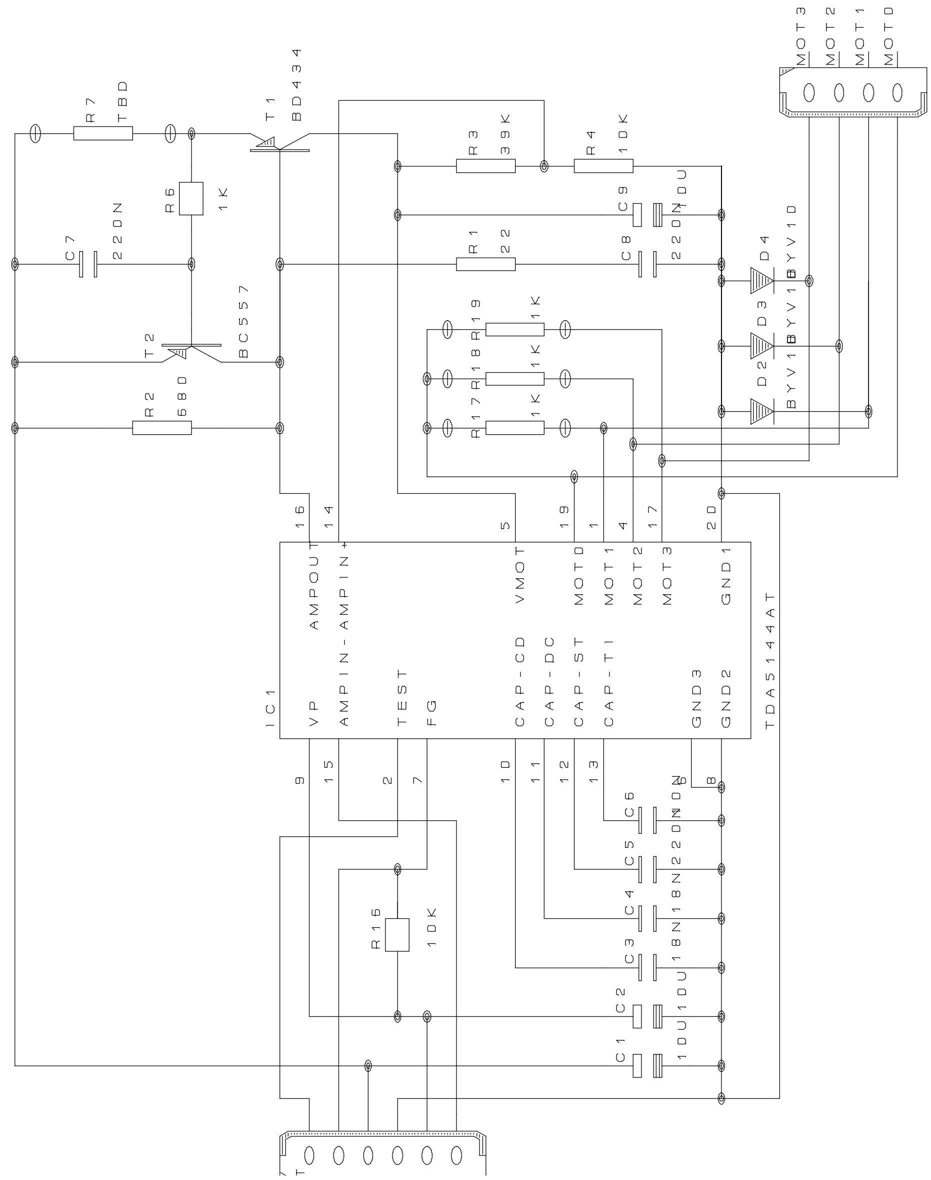


Figure 1

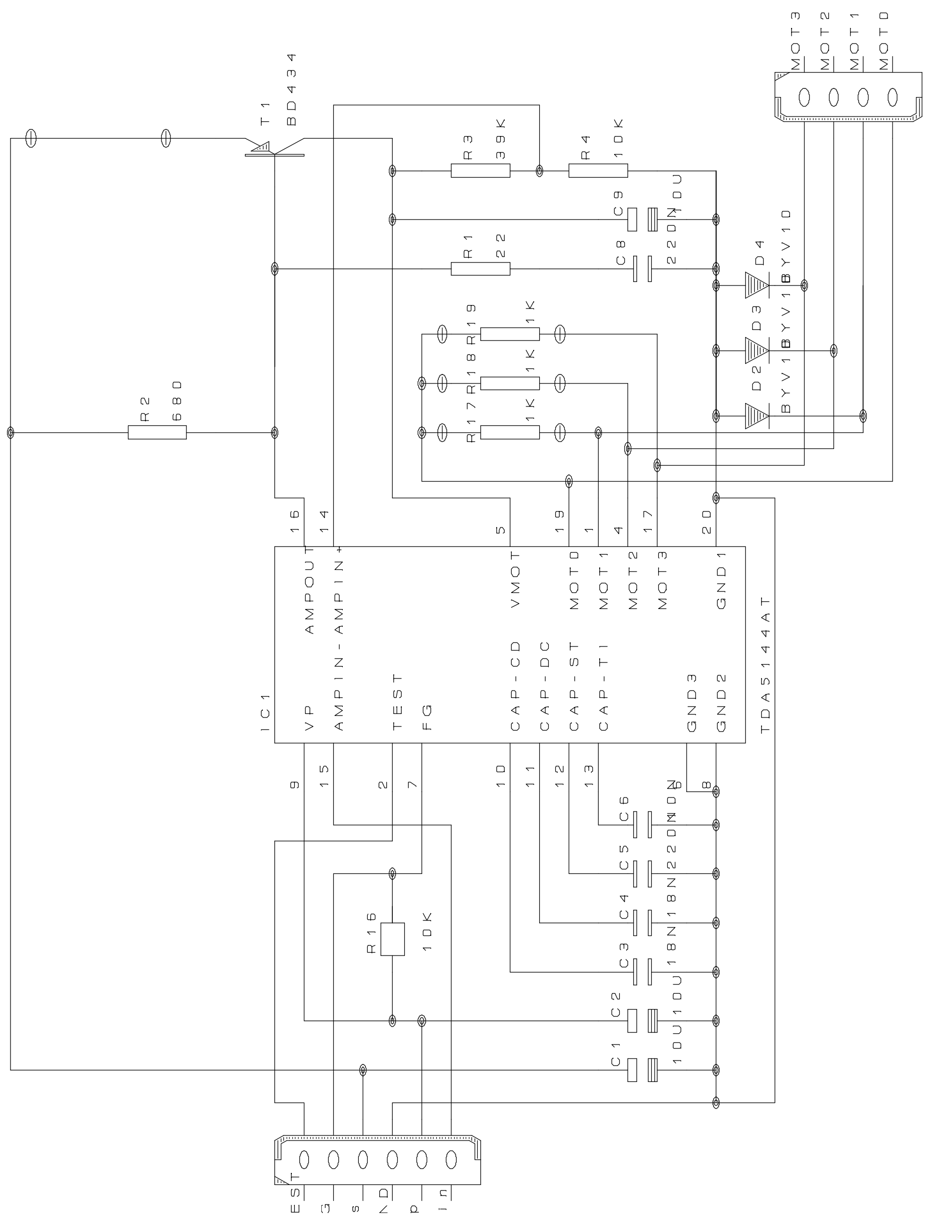
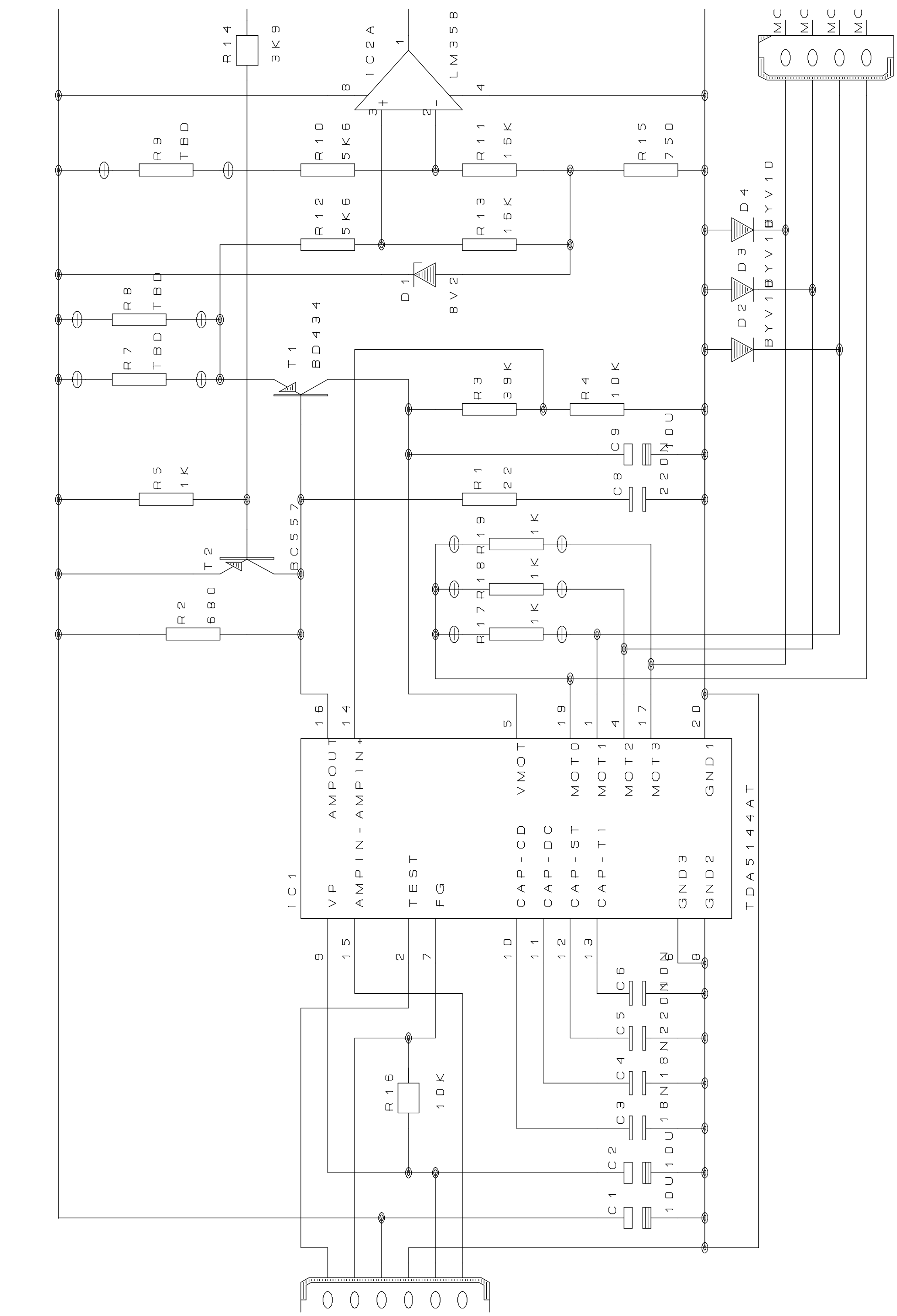

T1
BD434
R3
39K
R4
10K
C9
10U
C8
220N
R1
22
D4
D3
D2
BYV10
MOT3
MOT2
MOT1
MOT0
R19
R18
R17
1K
R2
680
IC1
TDA5144AT
AMPOUT
AMPIN+
AMPIN-
VMOT
MOT0
MOT1
MOT2
MOT3
GND1
GND2
GND3
VP
TEST
FG
CAP-CD
CAP-DC
CAP-ST
CAP-TI
C6
C5
C4
C3
C2
C1
R16
10K

Figure 2



2.1 Timing Capacitors.

For the motor control IC four timing capacitors are needed:

- The start capacitor CAP-ST (C5); this capacitor determines the

Figure 3

frequency of start oscillator.

- The two capacitors CAP-CD (C3) and CAP-DC (C4) for the adaptive commutation delay circuit; these capacitors are important to determine the optimal point of time for commutation.

- The timing capacitor CAP-TI (C6); this provides the system with necessary timing signals and the watch-dog function.

2.1.1 Start capacitor CAP-ST.

The start capacitor CAP-ST determines the frequency of the start oscillator. This capacitor is charged and discharged, with a current of 2 µA, from 0.05 V to 2.2 V and back to 0.05 V. The time to complete one cycle is given by:

$$t_{start} = 2.15 \ x \ CAP\text{-}ST \quad (with\ CAP\text{-}ST\ in\ \mu F) \tag{1}$$

The start oscillator is reset by a commutation pulse and so is only active when the system is in start-up mode. A pulse from the start oscillator will cause the outputs to change to the next state (torque in the motor). If the movement of the motor generates enough EMF the IC will run the motor. If the magnitude of the EMF is insufficient, then the motor will move one step only and will oscillate in its new position. The amplitude of the oscillation must decrease sufficient before the arrival of the next start pulse, to prevent that this pulse arrives during the wrong phase of the oscillation. The oscillation frequency of the motor can be expressed by the following equation:

$$f_{osc} = \frac{1}{2\pi \ x \sqrt{K_t \ x \ I \ x \ \frac{p}{J}}} \tag{2}$$

where:
K_t = torque constant (N.m/A)
I = current in A
p = number of magnetic pole-pairs
J = inertia (kg.m^2)

Example: J = 72 x 10^{-6} kg.m^2, K = 25 x 10^{-3} N.m/A, p= 6 and I = 0.5 A; this gives f_{osc} = 5 Hz. If the damping is high then a start frequency of 2 Hz can be chosen or t_{start} = 500 ms. According to equation (1) CAP-ST ≈ 0.23 µF.
So, take CAP-ST = 0.22 µF.

2.1.2. Adaptive commutation delay CAP-CD and CAP-DC

The commutation timing is determined with a timer circuit. The operation of this timer is based upon a charge and a discharge of a capacitor.
In figure 4 the signals MOT1 (m1.1), MOT2 (m1.2), MOT3 (m1.3) of a brushless DC-motor and the voltage on the CAP-CD (m1.4) and CAP-DC (ch4) have been displayed.
The charging of the CAP-CD starts at a certain zero-crossing and stops at a next zero-crossing and then a discharging of CAP-CD is started. A zero-crossing occurs half-way between two successive commutations. The discharge of CAP-CD is done with a current twice the charge current and a commutation is effectuated when the capacitor voltage is again on its starting (low) level. The charge of the capacitor is interrupted during the fly-back pulse on the outputs.
With the signal on CAD-CD only three of the 6 commutations per cycle can be done. The other three commutations are realized with an extra timer that the uses the same start and stop events for charge and discharge as for the CAP-CD timer. This extra timer operates as a slave of the CAP-CD timer.
A commutation is effectuated every time the voltage on CAP-CD or CAP-DC reaches the lowest voltage level.

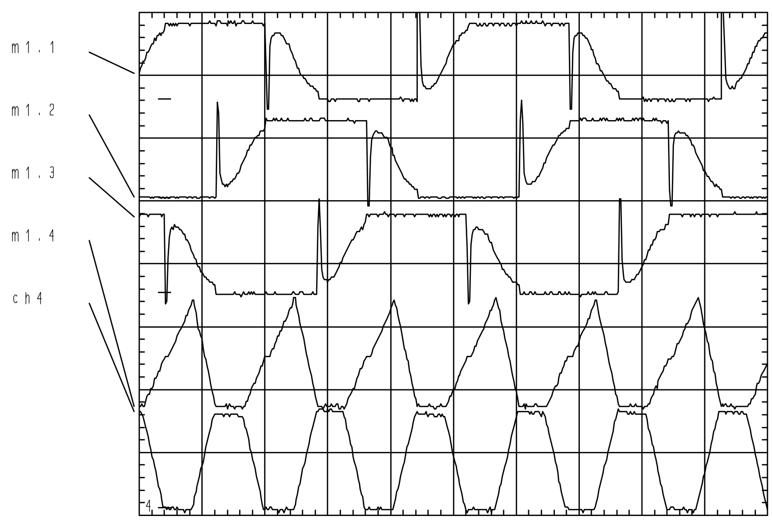


Figure 4, voltage wave forms of $V_{MOT1,2,3}$, V_{CAP-CD} and V_{CAP-DC}.

Below the calculation is given of the CAP-CD and CAP-DC. The calculation of these capacitors, for normal operation, is related to the required minimum and maximum number of revolutions per minute (RPM) of the motor and the number of magnetic pole-pairs.
The cycle time of the motor voltage is defined between two corresponding commutation events and can also be defined as 360 electrical degrees.

$$t_{cyc} = \frac{60}{RPM \times n} \;; \quad \text{where } n = \text{number of magnetic pole pairs} \qquad (3)$$

The motor voltage is commutated 6 times per cycle, so the time between two commutations is:

$$t_{com} = \frac{60}{RPM \times n \times 6} = \frac{10}{RPM \times n} \qquad (4)$$

According to the specification of the TDA5144 are for CAP-CD the charge, discharge currents and minimum and maximum voltage levels are defined as follows:

I_{ch}	min. 10.6 µA,	typ. 16.2 µA,	max. 22 µA.
I_{dis}	min. 5.3 µA,	typ. 8.1 µA,	max. 11 µA.
I_{ch}/I_{dis}	min. 0.9	typ. 1.025	max. 1.15

$V_{cap\text{-}low}$	min. 0.85 V,	typ. 0.875 V,	max. 0.9 V
$V_{cap\text{-}high}$	min. 2.3 V,	typ. 2.4 V,	max. 2.55 V

The voltage excursion on CAP-CD is the largest for the lowest value of RPM. The CAP-CD is determined at the lowest value of RPM, or at the maximum time between two successive commutations (t_{commax}). The voltage excursion on CAP-CD decreases with increasing RPM.

For the calculation of CAP-CD the minimum value of the maximum possible voltage excursion (ΔV_{capmax}) on CAP-CD and the maximum value of the charging current I_{chmax} has to be taken into account. CAP-CD is charged during t_{com}.
CAP-CD can be determined as follows:

$$CAP\text{-}CD = \frac{I_{ch} \times t_{com}}{\Delta V_{cap}} = \frac{I_{ch} \times 10}{RPM \times n \times \Delta V_{cap}}; \quad or \quad CAP\text{-}CD \geq \frac{I_{chmax} \times 10}{RPM_{min} \times n \times \Delta V_{capmin}} \qquad (5)$$

With ΔV_{capmin} = 2.3 - 0.9 = 1.4 V and I_{chmax} = 22 µA the CAP-CD is determined as follows:

$$CAP\text{-}CD \geq \frac{10 \times 22 \times 10^{-6}}{RPM_{min} \times n \times 1.4} \geq \frac{160}{RPM_{min} \times n} \quad (C \text{ in } \mu F) \qquad (6)$$

With decreasing t_{com} (higher RPM) the voltage excursion on CAP-CD decreases. The maximum value of the motor RPM at which a reliable commutation is obtained is mainly determined by a minimum allowable voltage excursion on CAP-CD and CAP-DC. At a small voltage excursion on CAP-CD or CAP-DC noise can disturb the proper operation of the commutation mechanism. So, if a minimum allowable V_{cap} is assumed, the ratio for a minimum and maximum value for RPM can be calculated for a given value of CAP-CD.
Below the maximum value of RPM has been calculated with a minimum allowable voltage excursion (ΔV_{capmin}) and the minimum charge current (I_{chmin}) for CAP-CD.

$$RPM_{max} = \frac{I_{chmin} \times 10}{\Delta V_{capmin} \times n \times CAP\text{-}CD} \qquad (7)$$

The capacitor CAP-CD has been determined for RPM_{min} with a voltage excursion ΔV_{capmax} and a

maximum a charge current on CAP-CD.
The ratio between RPM_{max} and RPM_{min} is given by the following equation

$$\frac{RPM_{\max}}{RPM_{\min}} = \frac{\Delta V_{capmax} \times I_{chmin}}{\Delta V_{capmin} \times I_{chmax}} \quad (8)$$

With the result of (8) the ratio of RPM_{max} and RPM_{min} for a given minimum voltage excursion ΔV_{capmin} can be determined to assure a reliable operation.

For the evaluation board a CAP-CD value has been calculated for a 6-pole pairs motor with a RPM_{min} of 1800.

CAP-CD ≥ 160 / 1800 x 6 = 0.0148 µF:
Take C3 = 18 nF.

CAD-DC has been connected to a similar timer circuit and needs the same capacitor value as used for CAD-CD.
So, C4 = 0.018 µF.

2.1.3. Timing capacitor CAP-TI

Capacitor CAP-TI is used for timing the successive steps within one commutation period; these steps include some internal delays.

The most important function, however, is the watchdog time in which the motor EMF has to recover from a negative diode fly-back pulse back to a positive EMF voltage (or vice versa). The watchdog timer is a guarding function that only becomes active when the expected event does not occur within a predetermined time.

The EMF usually recovers within a short time if the motor is running normally (<< ms). However if the motor is motionless or rotating in the reverse direction, then the time can be longer (>> ms).

In figure 5 two situations have been drawn. The upper two signals V_{MOT} and $V_{CAP\text{-}TI}$ represent a normal situation. The lower two signals represent a situation where the MOTOR runs in reverse direction.

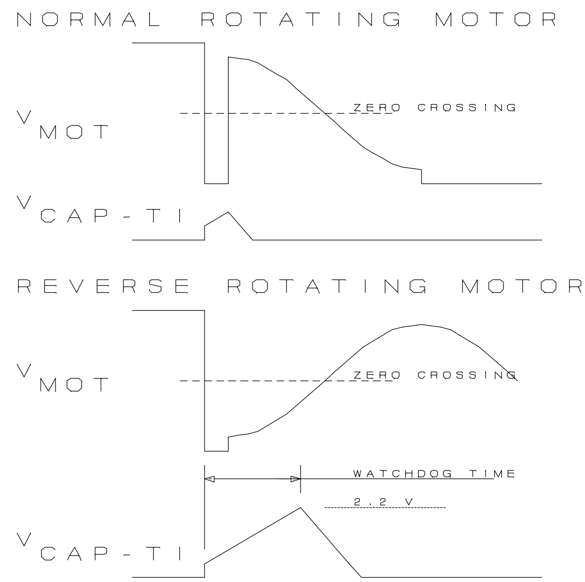


Figure 5

When $V_{CAP\text{-}TI}$ reaches a voltage of 2.2 V before the zero crossing with correct polarity after a leading edge of the diode fly-back pulse, the internal logic of the TDA5144 is triggered to adapt the sequencer that drives the outputs.

A watchdog time must chosen so that it is enough long for a motor without EMF (still) and eddy currents that may stretch the voltage in a motor winding; however, it must be short enough to detect reverse rotation. If the watchdog time is made too long, then the motor may run in the wrong direction (with little torque).

The capacitor is charged with a current of 57 µA from 0.2 to 0.3 V. This time is used for internal timing. Above this level it is charged with a current of 5 µA to 2.2 V. This level is only reached if the motor EMF remains in the wrong polarity (watchdog function). At the end, or, if the motor voltage becomes positive, the capacitor is discharged with a current of 28 µA. The watchdog time is the time taken to charge the capacitor with 5 µA from 0.3 V to 2.2 V. The value of CAP-TI is given by:

$$CAP\text{-}TI = \frac{5 \times 10^{-6} \times t_{wd}}{1.9} = 2.63\ t_{wd} \qquad (9)$$

The minimum value for CAP-TI is 2000 pF. In practice a capacitor of 0.01 µF will mostly be a good choice. For the evaluation board a value of 0.010 µF has been chosen as well.

2.2 Speed Control

The motor speed can be controlled by the voltage on the IC-pin V_{mot}. This is the supply voltage of the on-chip power stages.
On the evaluation board the voltage V_{mot} is regulated in an analogue way by a series regulator transistor T1 (BD434). The IC has been provided with an uncommitted Operational Transconductance Amplifier (OTA) that is able to drive directly the series regulator. The voltage V_{mot} is coupled back, via a voltage divider R3 and R4 (see figure 1), to the input AMPIN+ of the OTA. The input voltage V_{in}, to control V_{mot}, is connected to the AMPIN- input of the OTA.
The voltage divider R3 and R4 determines the control range for V_{in} in relation to V_{mot}.

Assume a supply voltage V_s of 12 V ± 10% and a minimum allowed voltage drop across R7 and T1 of 1 V to guarantee a linear control.

With these requirements the maximum value of V_{mot} can be 0.9 x 12 - 1 V = 9.8 V.

In general applies for the OTA AMPIN+ = AMPIN-

The value of V_{in} at which, with the given resistors R3 and R4, a V_{mot} of 9.8 V is obtained can be calculated as follows:

$$V_{in} = V_{mot} \frac{R4}{R3 + R4} = 9.8 \frac{10}{39+10} = 2\ V \qquad \textbf{(10)}$$

With a voltage V_{in}, ranging to 2 V, the voltage V_{mot} can be controlled up to 9.8 V. The voltage V_{mot}, however, may not become lower than 1.7 V for running conditions, because at lower voltages the back-EMF comparators do not operate correctly at this low voltage.

The values for R3 and R4 have to be adapted when a different relation between V_{mot} and V_{in} is required.

In the example above the maximum voltage on the OTA inputs for the active control range is 2 V. Because the supply for the OTA is retrieved from V_P the maximum voltage on the OTA inputs is related to V_p and the inputs voltages may, according to the specification, not exceed V_p - 1.7 V. So, keep in mind that, when other values for R3 and R4 have to be determined the maximum input voltages on the OTA inputs are not exceeded.

The series connection of R1 and C8 is necessary to obtain a good stability of the control loop.

2.3 One-transistor controlled current limiter.

The circuit of figure 2 shows a set-up with a one-transistor option that is based upon the voltage drop of about 0.8 V across a current sense resistor R7.
At a voltage of about 0.8 V across R7 the transistor T2 turns-on and decreases the drive for the MOSFET that determines the voltage V_{mot}.
The current sensing resistor R7 has to be determined (TBD) by the user and is of course dependent on the current limiting value.

This is a simple low-cost solution, easy to implement, that can be used in applications where a relative high voltage drop across the current sensing resistor is allowed. This high voltage drop over R7 reduces the efficiency of this circuit.

Due to the fact current limiting value is mainly determined by the V_{be} of one transistor. The V_{be} has a temperature coefficient of -2 mV/°C. This gives at a V_{be} of 800 mV a tolerance on the current limiting value of -0.25 %/°C.
So, the value at which the current is limited is temperature dependent.

In figure 6 a typical relation has been given between the current sense resistor value and the current limiting value. This relationship has been determined by measuring the start-up current of the motor at different values of R7.

In this figure we see good linear relationship between the current sense resistor (R7) and the current limiting value.
The voltage drop across the current sensing resistor can be found by multiplying the resistor value and current limit value.

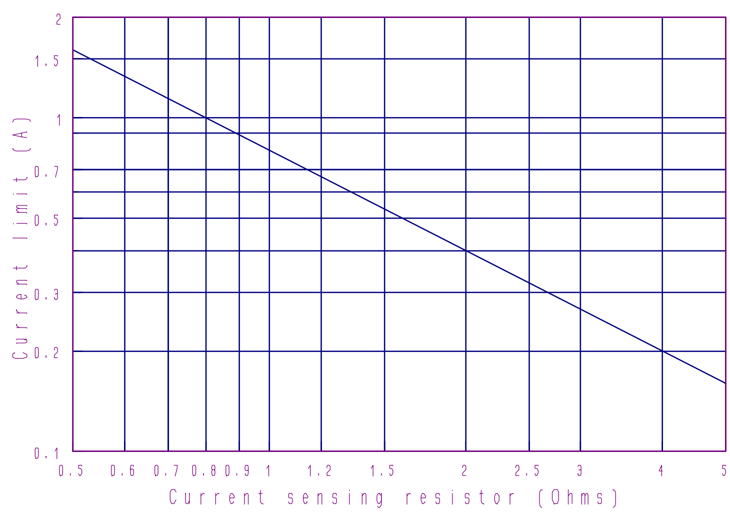


Figure 6

2.4 Op-amp controlled current limiter.

In figure 3 the option with an op-amp controlled current limiter has been depicted. By using an op-amp the voltage drop across the current sense resistors R7 and R8 can be much lower than in the one-transistor solution.
For this application a minimum current limiting sense voltage V_{cs} of 100 mV is assumed. For lower values of V_{cs} the accuracy of the current limiting value decreases.

The current limiting sense voltage V_{cs} is referenced to the positive supply rail. To measure this voltage an op-amp has to be used that allows common mode input voltages that are equal to the positive supply voltage of the op-amp. Most general purpose low-cost op-amps can not handle this kind of input voltages.

To be able to use a low-cost general purpose op-amp the input voltage has to be shifted to a level of about 2 to 3 V below the positive supply voltage. For a µA741 type op-amp this voltage shift must be at least 3 V and for an LM358 type this voltage must be 2 V. Because of its smaller DC voltage shift the LM358 has been used for further calculations.

The voltage level shift has been realized by the resistors R12 and R13. These resistor values have been determined in such a way that the voltage on the + and - terminal of the op-amp is at least 2 volts negative with respect to the positive supply V_s. The reference value for the current limiting value is set by the resistors R9, R10 and R11. To obtain a supply voltage independent DC voltage shift this circuit part has been bridged by a zener diode of 8.2 V. For the calculations a zener diode is taken with a tolerance of ± 5 %.

Since we are operating in this application with DC voltages of 100 mV the offset voltage of the op-amp has to be considered as well. An offset voltage of ± 10 mV, which is normal for general purpose low-cost op-amps, introduces a tolerance of ± 10 % on the current limiting value at a V_{cs} of 100 mV.
By shifting the DC level of the sense voltage, the amplitude of V_{cs} on the op-amp input is reduced and thus increases the influence of the offset of the op-amp. To minimize this influence as much as possible the level shift should be kept as small as possible.

This can be explained as follows:

$$V_+ = V_s - V_{D1} + (V_{D1} - V_{cs})\,\frac{R13}{R12 + R13}; \quad or \quad V_+ = V_s - V_{D1} + A \times (V_{D1} - V_{cs}) \qquad \textbf{(11)}$$

The values for R12 (5.6 KΩ) and R13 (16 KΩ), mentioned in figure 3, have been calculated for a level shift of 2 V at a V_{D1} of 8.2 V - 5% (7.79 V) and V_{cs} = 0 V.

Substituting the values for R12 and R13 in expression 11 gives:

$$\mathbf{V_+ = V_s - 7.79 + 0.741\,(7.79 - V_{cs}) = V_s - 7.79 + 5.77 - 0.741\,V_{cs} = V_s - 2.02 - 0.741\,V_{cs}}$$

From this result we see that a actual current sense voltage on the op-amp input is reduced by a factor of 0.741. This means a voltage 74.1 mV on the op-amp input at V_{cs} = 100 mV.
The LM358 has a maximum offset voltage of ± 9 mV. With this offset voltage and a V_{cs} of 100 mV the output of the op-amp will become active at input voltages varying between 90.5 and 72.5 mV, which means an initial tolerance ± 12%.

The voltage on the V- of the op-amp is determined as follows:

$$V_{-} = (V_s - V_{D1}) + V_{D1}\frac{R11}{R9 + R10 + R11}; \quad or \quad V_{-} = (V_s - V_{D1}) + B \times V_{D1} \tag{12}$$

For op-amps applies: **V+ = V-**

For V_{cs} now the following expression can be derived:

$$V_{cs} = V_{D1} - V_{D1}\frac{B}{A} \tag{13}$$

From eq. 13 one can see that V_{cs} is direct proportional to the zener voltage. So, with a tolerance on the zener voltage of ± 5% this gives a initial tolerance on the current limiting value of ± 5%.

With R9 the maximum value of the current voltage V_{cs} can be set. The actual current limiting starts when this voltage is reached.

By choosing R11 = R13 and R10 = R12 for V_{cs} the following applies:

$$V_{cs} = V_{D1}\frac{R9}{R9 + R10 + R11} \tag{14}$$

With V_{D1} = 8.2 V, R10 = R12 = 5.6 KΩ and R11 = R13 = 16KΩ the relation between V_{cs} and the resistor R9 can be determined.

This relation has been given in figure 7. With this figure the resistor R9 for a certain maximum current sense voltage can easily be determined.
This resistor can be soldered between the solder pins on the pc-board.

The current sensing resistors R7 and R8 (TBD) have to be determined by the user and the value is dependent on the current limiting value and the maximum allowable value of V_{cs}. The tolerance of this resistor, which has a very low value, is important for the overall tolerance on the current limiting value.
The initial tolerance on the current limiting value is determined by the magnitude of the op-amp offset voltage, the tolerance on the zener diode voltage and the resistors R9 up to R13.

The influence of the offset voltage can be reduced by using an op-amp with a lower offset voltage, or by allowing a higher value for V_{cs}. A current sense voltage V_{cs} of 200 mV halves the influence of the offset voltage.

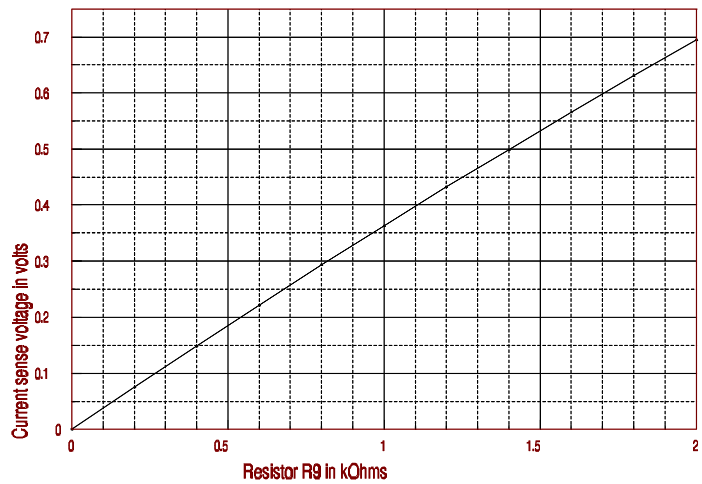


Figure 7

The second parameter that influences the current limiting value is the tolerance of zener diode voltage and its temperature drift.

The temperature drift of a BZX C 8V2 zener diode is max. 6.2 mV/°C. This means for a temperature range of 25°C ± 25°C a zener voltage variation of ± 150 mV, or related to a minimum zener voltage of 0.95 x 8.2 V = 7.79 V at 25°C an additional tolerance of ± 2%. The temperature drift has not to be taken into account for the initial tolerance on the current limiting value.

The influence of the initial zener diode tolerance can only be reduced by using a zener diode with lower tolerance. Zener diode of the BZX 79-family are available with tolerances of 1, 2, 3 and 5%.

The resistors R9, R10, R11, R12 and R13, used to realize the DC voltage shift and the current limiting value introduce an extra tolerance as well. So, if a narrow tolerance is needed on the current sense level also low-tolerance resistors should be used.

The total initial tolerance on the current limiting value, caused by the op-amp offset, zener diode voltage and resistor tolerances, can be calculated as follows:

Offset related tolerance at V_{cs} = 100 mV:	± 12 % (offset ± 9 mV)
Zener diode related tolerance:	± 5 % (5% zener diode)
Resistors tolerance:	± 5 %
Total initial tolerance :	± 23 % (1.12 x 1.05 x 1.05)

3. Grounding and Decoupling.

For the application of this IC we have to take care for a good grounding. For the grounding we must distinguish the grounding of the analogue part (timing and speed control) and the grounding for the power part.
The IC has three GND pins, where GND1 is considered to be the ground line for the power output stages and GND2 the reference for the analogue part.
GND3 is on the chip not connected to the ground line and has externally to be connected to GND2.

Most applications have only one common ground connection and therefore it is important that the routing of the ground tracks on the pcb lay-out is done in such a way that ground noise does not influence the proper operation of the circuit.

In figure 8 a possible lay-out routing for the ground tracks is given. In this lay-out also two decoupling capacitors have been drawn.
This lay-out set-up has also been used for the evaluation board and has given good results. In this lay-out set-up one can see that the current of output stages does not flow through the common line of the timing capacitors.
The place of the decoupling capacitor C2 is important. This capacitor has to connected as close as possible to the IC. The impedance of this capacitor must be lower than 2 Ω at a frequency of 2 MHz.

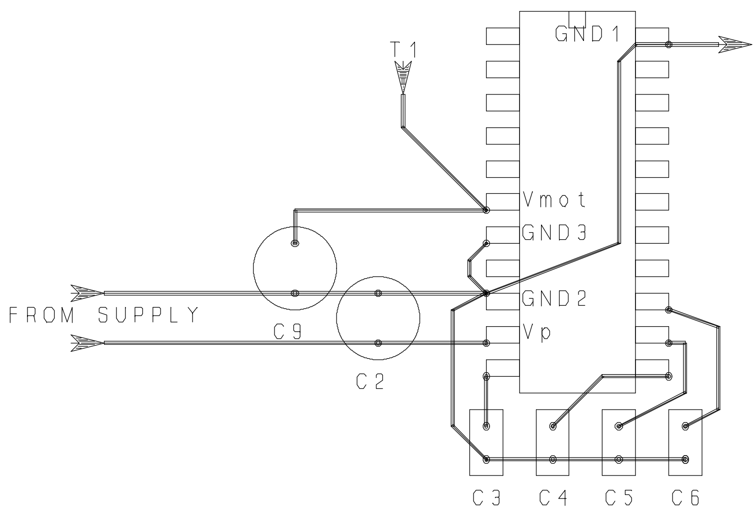


Figure 8

Most wet electrolytic capacitor of at least 10 µF can meet this specification.

In figure 9 and 10 the lay-out of both evaluation boards have been depicted. The lay-outs are almost equal; the difference is the IC package and the pin lay-out of connector X1.

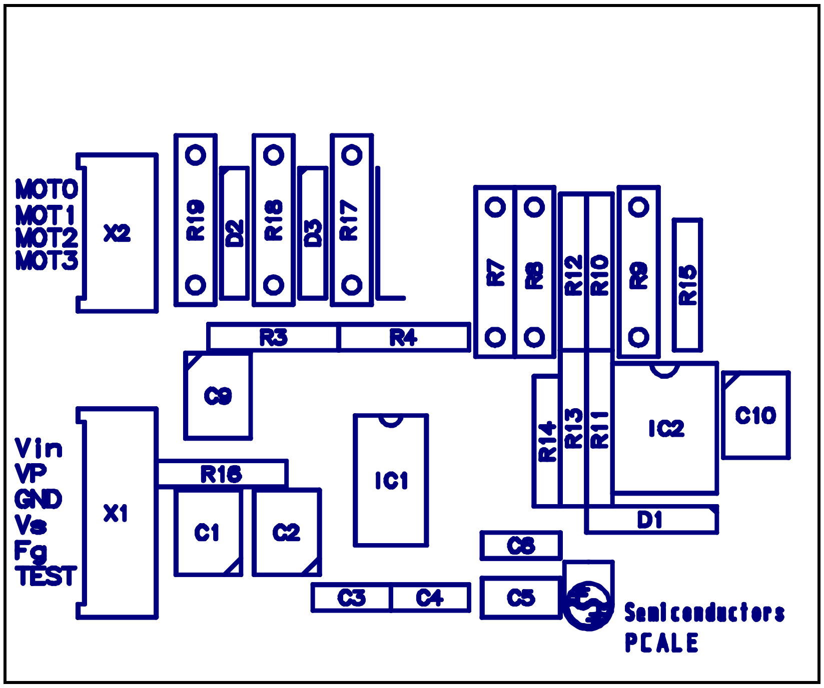


Figure 9, pc-board lay-out for the TDA5143T and TDA5144AT.

4. Parts list.

In the table below the complete parts list has been given, including the optional components for the two current limiters. The components that have only to be used for the one-transistor current limiter are marked with O1, for the op-amp controlled current limiter with O2 and for both options with O12.
The resistors R7 and R8 are low-ohmic and may have a value < 1 Ω. This type of resistor is available in the PR01 series of Philips Components.

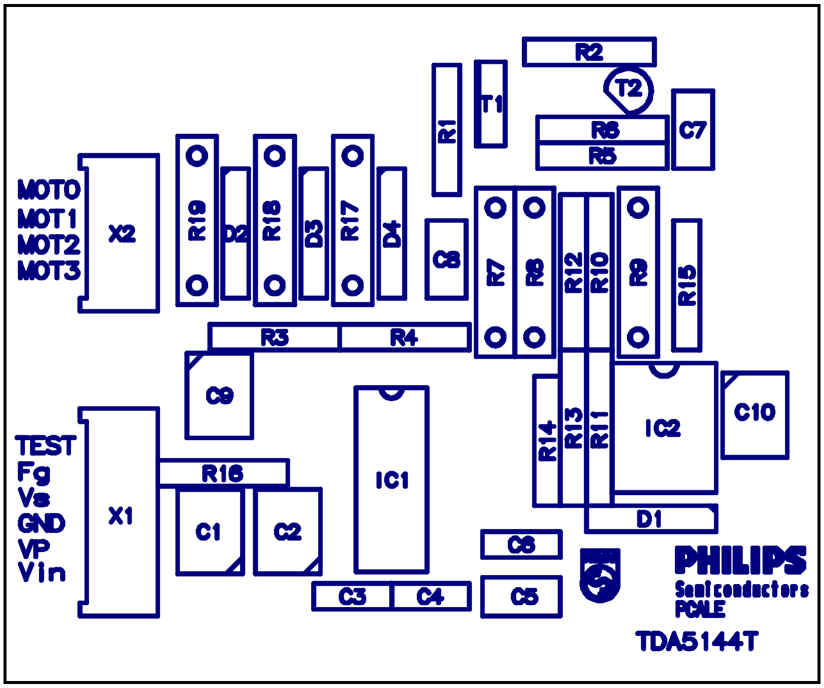


Figure 10, pc-board lay-out for the TDA5144T.

Item	Option	Value	Type	12-nc
R1		22 Ω ± 5%	SFR25	2322 181 53229
R2		680 Ω ± 5%	SFR25	2322 181 53681
R3		39 KΩ ± 5%	SFR25	2322 181 53393
R4		10 KΩ ± 5%	SFR25	2322 181 53103
R5	O2	1 KΩ ± 5%	SFR25	2322 181 53102
R6	O1	1 KΩ ± 5%	SFR25	2322 181 53102
R7	O12	TBD	PR01	2322 193 .....
R8	O2	TBD	PR01	2322 193 .....
R9	O2	TBD	SFR25	2322 181 53...
R10	O2	5.6 KΩ ± 1%	SFR25	2322 181 53562
R11	O2	16 KΩ ± 1%	SFR25	2322 181 53163
R12	O2	5.6 KΩ ± 1%	SFR25	2322 181 53562
R13	O2	16 KΩ ± 1%	SFR25	2322 181 53163
R14	O2	750 ± 5%	SFR25	2322 181 53701
R15	O2	3.9 KΩ ± 5%	SFR25	2322 181 53392

R16		10 KΩ ± 5%	SFR25	2322 181 53103
R17		1 KΩ ± 5%	SFR25	2322 181 53102
R18		1 KΩ ± 5%	SFR25	2322 181 53102
R19		1 KΩ ± 5%	SFR25	2322 181 53102
C1		10 µF	RSM 037	2222 037 68109
C2		10 µF	RSM 037	2222 037 68109
C3		18 nF	MKT 370	2222 370 21183
C4		18 nF	MKT 370	2222 370 21183
C5		220 nF	MKT 370	2222 370 17224
C6		10 nF	MKT 370	2222 370 21103
C7	O1	220 nF	MKT 370	2222 370 17224
C8		220 nF	MKT 370	2222 370 17224
C9		10 µF	RSM 037	2222 037 68109
C10	O2	10 µF	RSM 037	2222 037 68109
IC1			TDA5143/51444	
IC2	O2		LM358N	
T1			BD434	
T2	O12		BC557	

D1	O2	BZX79 C 8V2
D2		BYV10-30
D3		BYV10-30
D4		BYV10-30