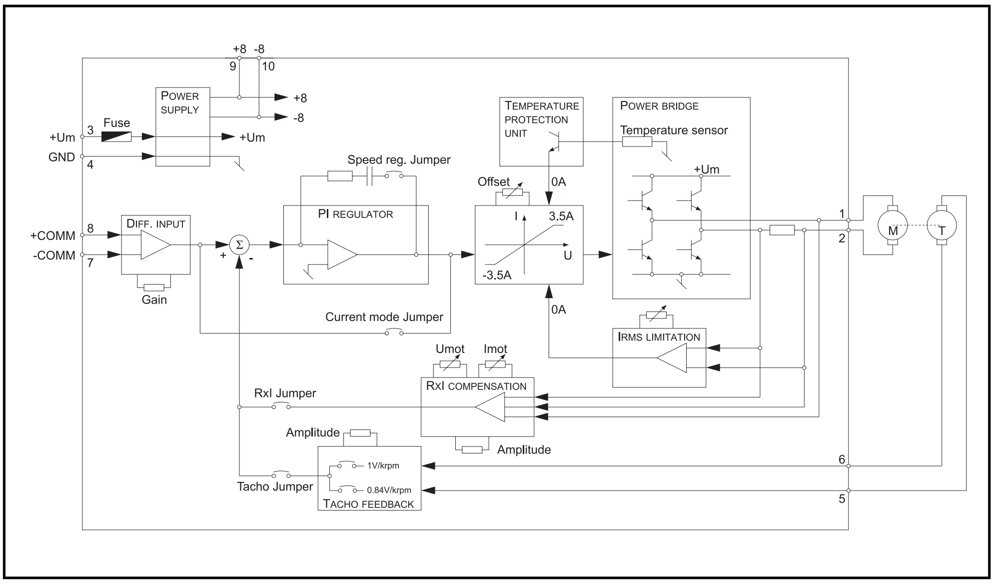
Portescap

DRIVERS FOR
IRONLESS ROTOR D.C. MOTORS



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1. Introduction

Ironless rotor DC motors present characteristics highly desirable for speed and position servos. Due to the absence of iron there is no cogging, and the rotor has no preferred positions, phenomena which tend to deteriorate precision of a servo. Furthermore, their low rotor inertia improves the dynamics of the entire drive system.

Control of a D.C. motor requires the variation of its supply voltage, or its current, or both. In other words, the electrical power supplied to the motor needs to be adapted to the work to be done. The driver amplifier takes care of this. The current determines the motor torque. The mean terminal voltage determines the angular velocity at which this torque is being supplied.

Depending on the type of application, control of only one of the two parameters is sufficient.

For controlling speed and/or position of the load "L" (figure 1), the motor "M" requires a speed sensor "T" and/or a position sensor "P". Their information is processed in the speed/position controller "C" which generates the error signal for piloting the drive amplifier "A". The driver is an important part of the servo loop, particularly so with the dynamics ironless rotor motors are capable of, and which require a driver of adequate performance.

In this paper we consider amplifiers for DC motors up to 1 kW output power. Speed or position servo loops are not dealt with.

2. The ironless rotor motor technology

Because of the relatively long air gap, permeability of the magnetic circuit of such motors is quite low. Therefore they have a lower inductance than iron rotor motors and a very short electrical time constant. Furthermore, electro-erosion of the brush gear is lower, and higher peak currents, or peak torques for that matter, are possible without the danger of demagnetising the motor. Their precious metal commutation system generates very little friction, and assures high reliability and long life.

All these factors present advantages for applications requiring high efficiency and also very high dynamic perfor-mance.

For applications needing high torque, motors using a copper commutator of larger diameter and copper-graphite brushes are required. The “Rotafente™” and “HPR” series of escap® ironless rotor DC motors utilise this type of commutation system. The rotor of these motors tolerates temperatures up to 155 °C, allowing for high continuous torque and peak torques which are very high for the motor size.

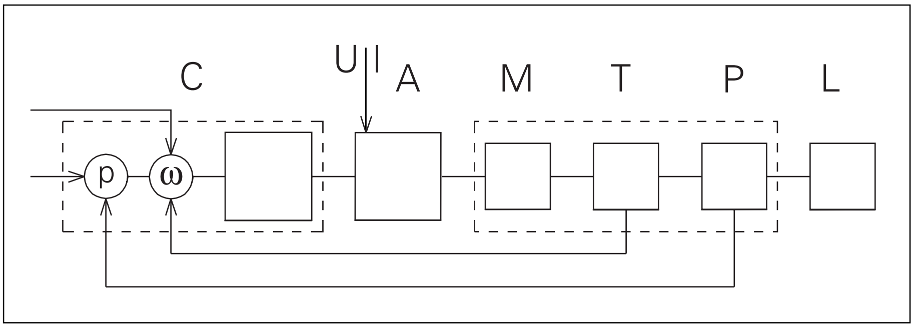

Fig. 1: Block diagram of a «Motion Control» chain

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3. Control modes

In practice there are two types of control:

	input reference	⇒	controlled output
• voltage control using a voltage amplifier:	input voltage	⇒	output voltage
• current control (using a current amplifier):	input voltage	⇒	output current

Each one of these power stages may utilise two different driver types:

- a linear amplifier
- a chopper amplifier

The usual means of current control, where current is measured using the voltage drop it causes on a small series resistance, is done with a transconductance amplifier. This amplifier is an adjustable current source where the user sets the gain between the input voltage and the output current. This gain is called the transconductance factor.

There is always some voltage loss with-
in any drive amplifier, so the motor terminal voltage is always lower than the power supply voltage.

3.1 Voltage control

Figure 2 shows the model of a voltage controlled DC motor. Remember that the driver may be either a linear amp or a chopper.

We have the power supply voltage U_{cont} into the amplifier, the load torque M, rotor inertia J, and motor speed ω at the output. The factor s is the Laplace variable.

One can see that the instantaneous velocity not only depends on the torque constant k_T and on inertia J, but also on motor resistance R and inductance L as well as on the back-EMF constant k_e, that means on speed.

The current rises freely to a value determined by the formula:

$$I = \frac{U - k\omega}{R}$$

Protection of the motor and the driver against overcurrents is a necessity, particularly with low impedance motors.

Furthermore, the delay in current rise due to the inductance L introduces a phase error which is unfavourable for servo loop stability. As a consequence, in addition to the pole due to its mechanical behaviour the voltage controlled motor presents an electrical pole depending on the parameters R and L, making it a second order system.

For a motor running at constant speed this is of no consequence. However, inductance opposes a fast change of speed, but with ironless rotor motors and their low inductance, its influence is often practically negligible.

3.2 Current control

With a current controlled amplifier the instantaneous motor speed depends only on k_T and on J. The amplifier includes an internal current loop using for instance a measuring resistance R_{mes}; it imposes the required motor current and, therefore, eliminates the delay caused by L. The suppression of the electrical pole of the motor makes it a simpler, first order system.

The direct control of torque, through current control, improves response time as long as the voltage available also covers the factor L × di/dt.

Voltage control does not need any feed-back, whereas current control of a D.C. motor only works together with a speed or position loop. Current control does, however, not need additional overcurrent protection.

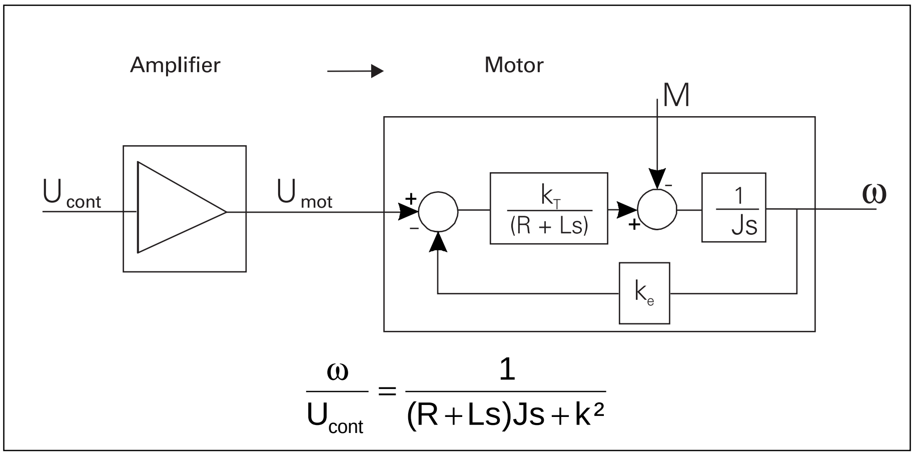

Fig. 2: Model of a voltage controlled D.C. motor and its transfer function

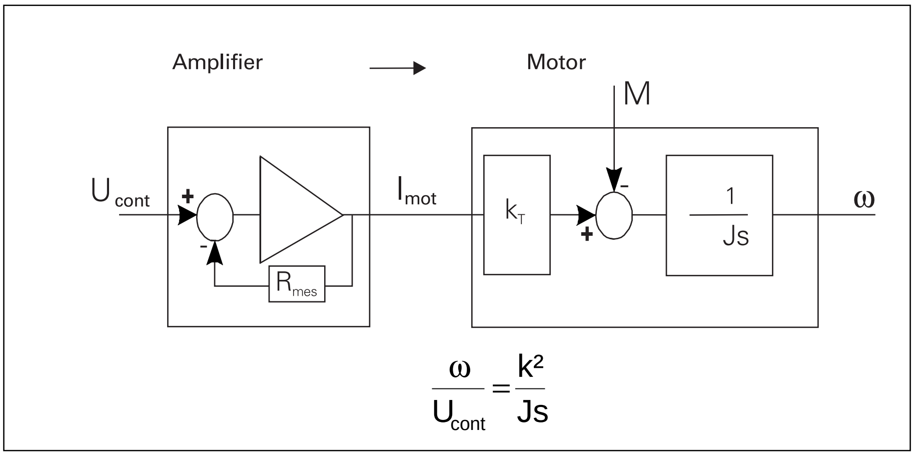

Fig. 3: Model of a current controlled D.C. motor and transfer function

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4. The linear amplifier

4.1 Principle of operation

Linear amplifiers adapt the power delivered to the motor by linearly changing the voltage or current. Supply power not delivered to the motor is dissipated in the driver:

$$P_{mot} = P_{supply} - P_{d(amp)}$$

While the supply voltage is constant, the motor voltage is not. The dissipation in the driver will be limited by using linear amps mainly with motors running at fairly constant conditions, and by choosing the supply voltage close to the value required according to the formula:

$$\begin{aligned} U_{supply} &= U_{amp} + U_{mot} \\ &= U_{amp} + R \times I(t) \\ &\quad + L \times di/dt + k \times \omega(t) \end{aligned}$$

U_{amp} is the voltage drop in the driver, the factor $L \times di/dt$ depends on dynamic conditions.

The linear driver could be an operation-al amplifier, whose simple concept provides a cost advantage, in particular for low power motors.

Let's consider the simple case of a motor operating under different conditions which require a maximum voltage of 12 V and a current of 1A. This means the driver must be capable of supplying that current at the maximum voltage (see figure 4).

If the voltage drop in the power stage is estimated at 5V, the power supply must deliver at least 17V to the driver (illustration A) .

The driver efficiency equals the ratio of output power to input power:

$$\eta = (12V \times 1A) / (17V \times 1A) = 0.71.$$

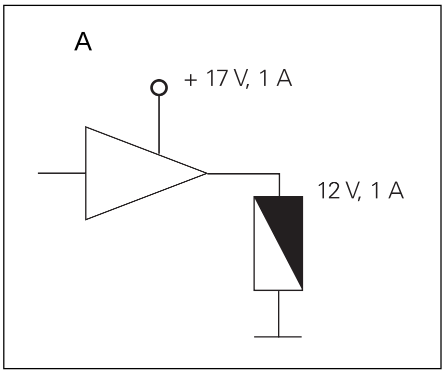
The driver must dissipate:

$$5V \times 1A = 5W.$$

If the motor has to work at different speeds, this situation changes. Consid-er first a case of lower speed and torque, where the motor needs only 6 V and 0.5A. Driver efficiency is (6V × 0.5A) / (17V × 0.5A) = 0.35, and it must dissipate:

$$11V \times 0.5A = 5.5 W.$$

Now let's consider the case where the motor needs only 6 V at 1A. Efficiency is (6V × 1A) / (17V × 1A) = 0.35 again. But with the amplifier receiving 17V it has to absorb a voltage drop of 11V (fig-ure 4B). It must therefore be capable of dissipating 11W without overheating in the existing ambient temperature.



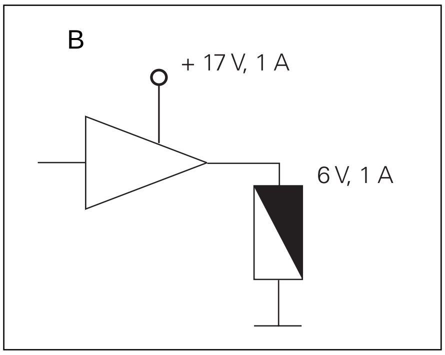


Fig. 4: Load driven from a linear amplifier, two operating conditions

The mean power dissipated in the driver follows the formula:

$$P_d = \int_0^T \left[U \cdot i_{(t)} - R \cdot i^2_{(t)} - \omega_{(t)} \cdot M_{(t)} \right] dt$$

4.2 Practical example

4.2.1 Optimised supply voltage

Let's consider an example to illustrate the procedure for choosing an optimum power supply voltage. The procedure is the following:

- determine maximum values of speed and torque in the application
- determine the minimum supply voltage
- for the amplifier, such as the ELD-3503 card, find out whether the power dissipation always remains below the allowed value.

The linear power amplifier ELD-3503 is described in section 7. The main figures are : power supply voltage 12 to 35V, voltage drop in the driver bridge approximately 6V, continuous current 2.5A, peak current 3.5A, continuous power dissipation 10W.

The application requires a continuous torque of M = 15 mNm at speeds from 1000 rpm to 4000 rpm but with low dynamics.

Step 1: Motor selection

A look at available API Portescap D.C. motors suggests as a first choice the motor 23V58-216E. It has a torque constant of k = 23.5 mNm/A and a resistance of 9.7 Ω at 22 °C, or 12.7 Ω at the maximum allowed temperature of 100 °C.

Step 2: Determine supply voltage.

The motor needs a current of I = M/k = 15/23.5 = 0.64 A. Establishing this current in the cold rotor (at 22 °C) requires a voltage of $R_{22} \times I = 9.7 \times 0.64 = 6.2$ V. With the rotor at 100 °C we need $R_{100} \times I = 12.7 \times 0.64 = 8.1$ V.

Back-EMF k × ω is 2.5V at 1000 rpm and 9.8 V at 4000 rpm. Under the different conditions the motor terminal voltage then varies from 6.2 + 2.5 = 8.7 V to
8.1 + 9.8 = 17.9 V.

Step 3: Determine dissipation in the amplifier

The card needs a power supply voltage of 17.9 + 6 = 24V. In the best case dissipation is (24 - 17.9) × 0.64 = 3.9 W; in the worst case (except for the starting current) it is (24 - 8.7) × 0.64 = 9.8 W. There is therefore no risk of overheating of the card. Motor stall current is (24 - 6)/9.7 = 1.86A, which is acceptable.

4.2.2 Imposed supply voltage

Now consider the case where, for some reason, the card has to run at 30V.

Then the stall current is (30 - 6)/9.7 = 2.47A and, in case of frequent starts, it would be wise to limit the peak current in order to improve the motor life.

Worst case power dissipation would be (30 - 6) × 0.64 = 15.4 W, which is above the 10 W limit, but it can be tolerated during short periods.

A simpler solution would be to choose a higher impedance winding such as the -213E in place of the -216E.

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5. The chopper amplifier

5.1 Principle of operation

In a switching amplifier power is supplied to the motor by using the power transistors as switches, which are either ON or OFF.

Actually, the power is controlled by varying the ON-OFF cycle of the switch, within one period of the chopper oscillator. At a frequency of 10 kHz this period is 100 µs. With the switch open for 10 µs and closed for 90 µs, the motor gets plenty of energy. In the opposite case it gets very little (see figure 5).

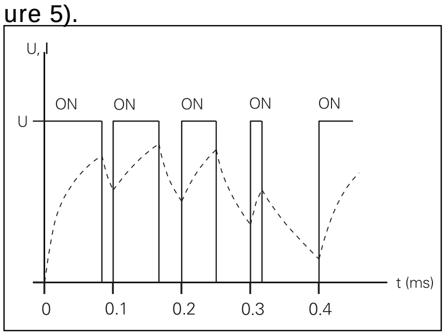

Fig. 5: Voltage (full line) and current (dotted line) with a chopper driver. The instantaneous current rises during the «VOLTAGE ON» time, its mean value depends on the ratio ON/OFF

With the switch open, losses are practically nil. With the switch closed, losses of MOS transistors are also quite low, similar to the switching losses occurring at each transition from saturation to open circuit. These are proportional to the chopper frequency but are generally negligible in MOS transistors. Therefore power dissipation in the amplifier is low.

Clearly, the amount of electrical energy required by the motor to do a job is the same whether supplied from a linear amplifier or from a chopper.

Therefore, it is interesting to measure the mean terminal voltage of a chopper driven motor, for different ratios of "t_{ON}" (bridge is active) and the chopper period "T". This mean voltage U_{av} follows the formula:

$$U_{av} = U_{supply}\,\frac{t_{on}}{T}$$

Obviously, a chopper controlled constant current driver does in fact control the average motor terminal voltage, by modifying the ratio t_{ON}/T until the desired current is obtained (the current corresponds to the average voltage U_{av}). The current value (measured across a small resistance in series with the motor) is effectively used as the feedback signal.

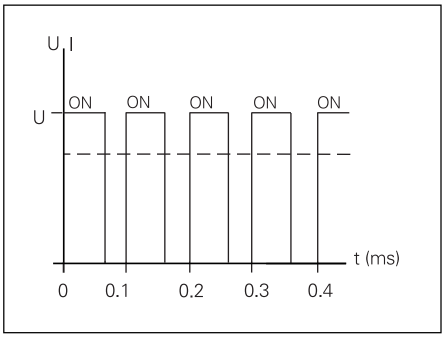

Fig. 6: With the bridge active for 2/3 of a period the motor terminal voltage U_{av} (broken line) is equal to about 2/3 of the supply voltage. When the ratio changes to 1/3 the voltage also changes to 1/3

Contrary to this, in a chopper controlled constant voltage driver it's the motor terminal voltage which serves as feedback signal, or the output of the sensor which monitors the value to be controlled.

5.2 Current circulation modes

Neglecting back-EMF, the motor can be considered as a load having resistance and inductance. With a duty cycle of D = 0.6, the transistor "Tr" is closed during 3/5 and open during 2/5 of a chopper period. Because of the energy stored in the inductance (½ LxI^2) the current cannot fall to zero at the instant where Tr opens, but is maintained in the direction established. The freewheeling diode Di (figure 8) allows this circulation, which is called non-regenerative mode.

With the more complex configurations using an H bridge with four transistors we can distinguish between three types of chopper control: non-regenerative (or non-recirculating, freewheeling) mode, the regenerative mode, and forced regenerative mode.

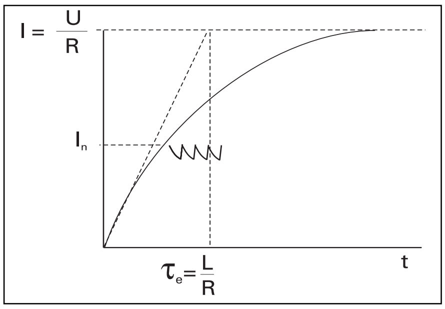

Fig. 7: In practice the voltage used is higher than necessary to obtain the nominal current I_n. Current rise depends initially only on the motor's electrical time constant (ratio L/R). Once I_n is reached, the chopper is activated and modulates the applied voltage to maintain current at I_n.

5.2.1 The non-regenerative mode

Consider the case where T1 and T4 of figure 9 conduct. Then during the OFF phase the chopper may be controlled in various ways. The first consists of deactivating one of the two transistors, say T1. The energy stored in the winding inductance is evacuated via transistor T4 and diode D2, maintaining current circulation in the direction established. Its value drops depending on the electrical time constant of this circuit, and on back-EMF, until T1 is switched ON again.

This is called non-regenerative mode, or freewheeling mode, where the current recirculates freely through a diode and the active transistor, which may be T1 or T4. The coil is either energised or practically short-circuited (neglecting the voltage drops in the transistor and the diode).

The characteristics of this mode are:

- low current ripple (during recirculation the motor is short-circuited; the motor voltage changes from +U to nearly zero)
- current rise and fall are symmetric
- no current control during its fall.

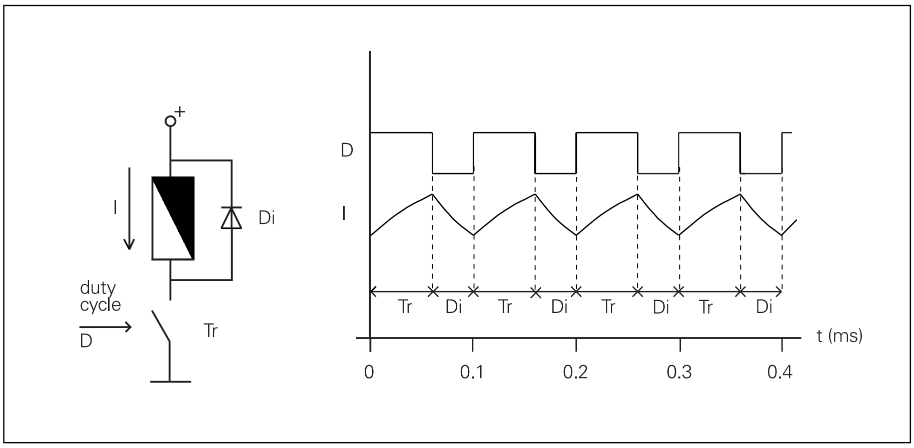

Fig. 8: Chopper principle: the transistor Tr is either open or closed; depending on the duty cycle D. The diode Di conducts each time Tr is open.

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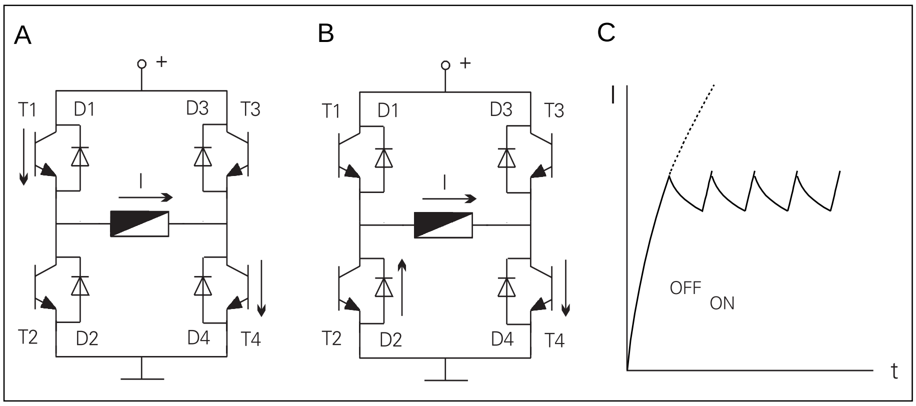

Fig. 9: A chopper H-bridge controlled in non-regenerative mode, and the current shape. Current circulation during the ON phase is shown at left and during the OFF phase, in the centre

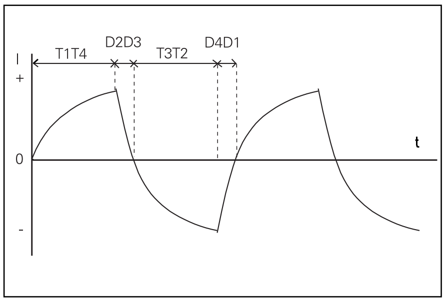

Fig. 13: The current of a chopper controlled in forced regenerative mode at a very low frequency would alternate between positive and negative values. The conductive elements are indicated above each phase of the period

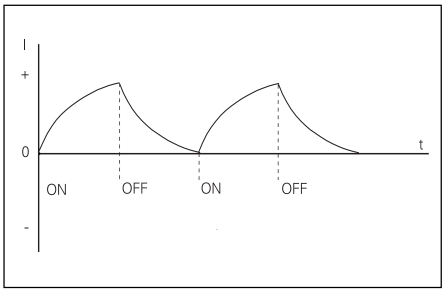

Fig. 10: A non-regenerative chopper having a very low chopper frequency would show a current alternating between a positive value and zero

5.2.2 The regenerative mode

The regenerative mode is another way of controlling a chopper, where during the OFF time all four transistors are opened. With the same direction as in the example 5.2.1 current continues to flow through D2, the coil, and D3, returning to the power supply capacitor (hence the term "regenerative", see figure 11). In that case the supply voltage is actually reversed, and the current drops more rapidly than in non-regenerative mode.

The characteristics of this mode are:

- high current ripple (during recirculation the motor voltage changes from +U to -U)
- current rise and fall are non symmetric
- during recirculation the current is under control.

5.2.3 The forced regenerative mode

With this mode, after current has decayed naturally to zero, it is forced to rise in the opposite direction by activating the other diagonal of the bridge (T2 and T3, see figure 12). Then, for the same direction as in 5.2.1, the current first recirculates through D2 and D3, then rises through T3 and T2 with reversed direction.

Similar to the regenerative mode the motor terminal voltage is reversed, leading to a very fast yet controlled decay of the current, which can then, via the transistors of the second diagonal, rise in the opposite direction (figure 13).

The motor terminal voltage follows the formula :

$$U_{av} = U_{supply}\left(2\frac{t_{on}}{T} - 1\right)$$

The characteristics of this mode are:

- current reversal during a single period, hence better control around zero current and also faster direction reversal
- simplified commanding of the bridge (diagonal by diagonal)
- high current ripple, superior to normal recirculation mode.

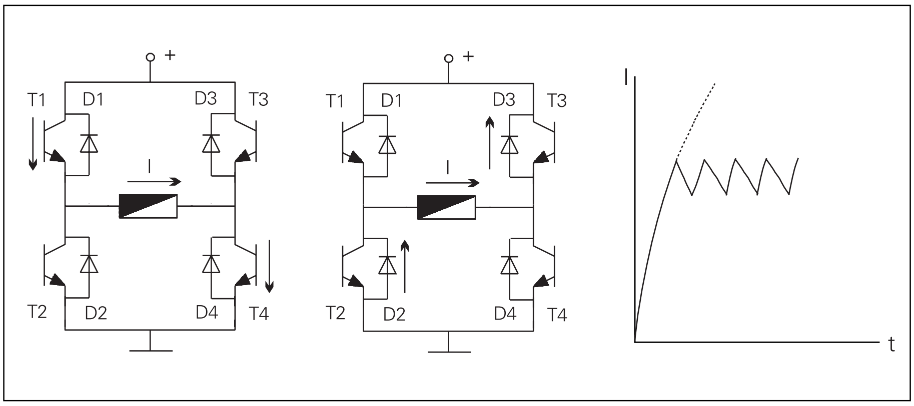

Fig. 11: A regenerative chopper and its current shape. Current circulation during the «ON» phase is shown at left and during the «OFF» phase, at the centre

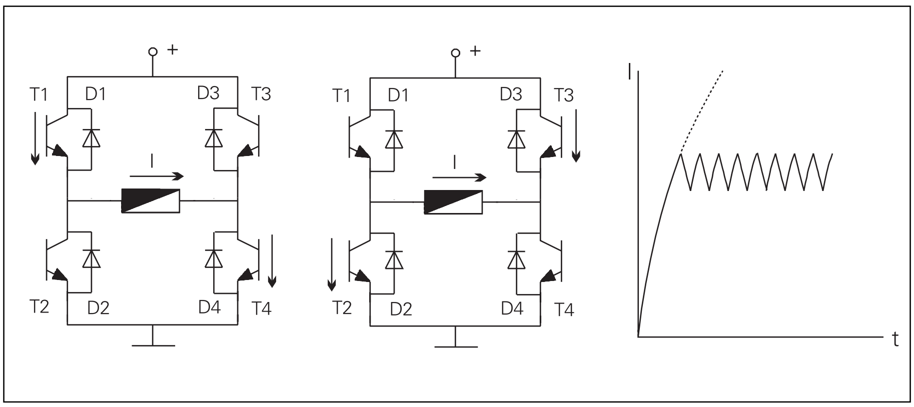

Fig. 12: The forced regenerative mode and the current shape normally obtained

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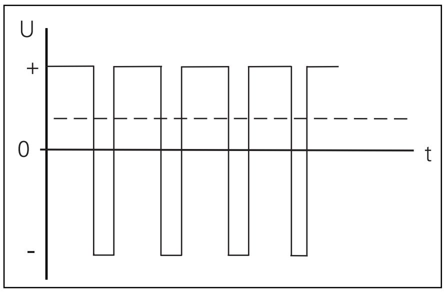

Fig. 14: Similar to the regenerative mode, in forced regenerative mode the motor terminal voltage is effectively reversed. With the bridge activated positive for 2/3 of a chopper period T and negative for 1/3, the average motor terminal voltage U_{av} (broken line) is approximately 1/3 of the positive supply voltage. If this ratio is reversed, U_{av} becomes negative. With a ratio 50:50 U_{av} is zero.

5.2.4 Practical example

The three different current circulation modes are now compared for the case of an escap® 23DT12-216P motor (k = 12.4 mNm/A, R = 2.6 Ω, L = 0.2 mH, t_e = 77 µs). It uses a 15 V driver with a chopper frequency of 20 kHz and runs at 3000 rpm and a load of 8 mNm.

At 3000 rpm, back-EMF is E = k x ω = 0.0124 x 314 = 3.9 V. When applying a short-circuit to the motor terminals, this voltage causes a braking current of 3.9V / 2.6Ω = 1.5 A.

With the chopper on, the voltage available to drive a current through the motor is U_{mot} = U - E = 15 - 3.9 = 11.1 V. The current rises towards a value of I = U / R = 11.1V / 2.6Ω = 4.27 A (for all recirculation modes). The speed of current rise corresponds to the electrical time constant of 77 µs.

The load torque requires in fact a current of I = M / k = 8mNm / 12.4mNm/A = 0.65 A. To obtain this mean value the actual current shape depends on the recirculation mode.

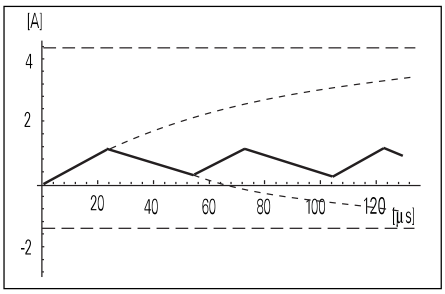

Fig. 15: Non-regenerative mode: The chopper acts at 1 A. When almost short-circuited, the back-EMF of 3.9 V wants to establish a current of -1.5A.

In non-regenerative mode (fig. 15), the negative current rises towards the value determined by back-EMF. In regenerative mode however, it corresponds to the sum of the reversed supply voltage and back-EMF, i.e. (-15V - 3.9V) / 2.6Ω = -7.3 A (fig. 16 and 17).

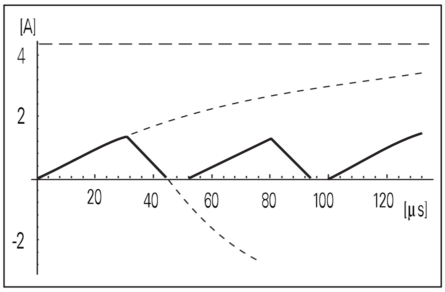

Figure 16: Regenerative mode: To obtain a mean current of 0.65 A the chopper acts only at about 1.5 A. The voltage then appears to be reversed; the current rises towards -7.3A.

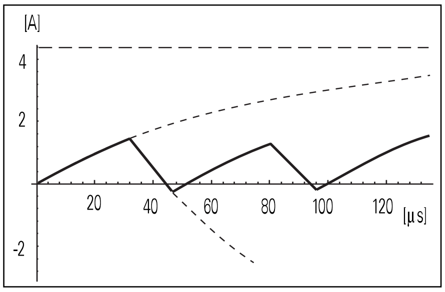

Figure 17: Forced regenerative mode: With the opposite bridge active the current rises again towards a value of -7.3A.

This example illustrates the bad influence of too low a chopper frequency, whose period lasts 50 µs compared to the motor's electrical time constant of 77 µs.

A higher chopper frequency having a shorter period would result in a lower current ripple and eliminate altogether the parts of zero or negative current. This is demonstrated in figures 18 and 19, which show the same motor under the same operating conditions but using a chopper frequency of 100 kHz.

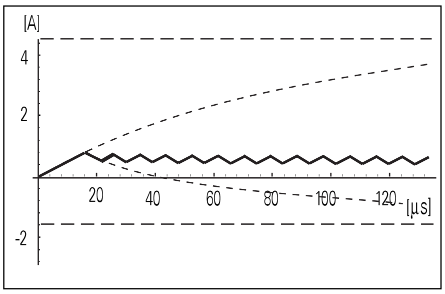

Fig. 18: Non-regenerative mode, chopper frequency of 100 kHz: The chopper acts at about 0.8 A.

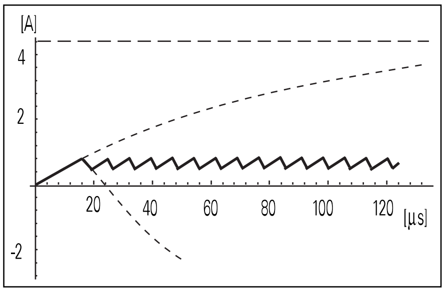

Fig. 19: Regenerative mode (forced or not): The chopper acts at about 0.8 A. The reverse current rises towards -7.3 A.

5.3 The effects of current ripple (form factor)

One inconvenience of the chopper amplifier is current ripple, its magnitude depending on the circulation mode. To quantify ripple, the form factor "F" is used, which is the ratio of rms current to mean current.
If current is separated into its D.C. and A.C. components, one gets the following relation : $F^2 = 1 + r^2$ where the ripple content r is the ratio between the rms value of the ripple and the average value of the current.

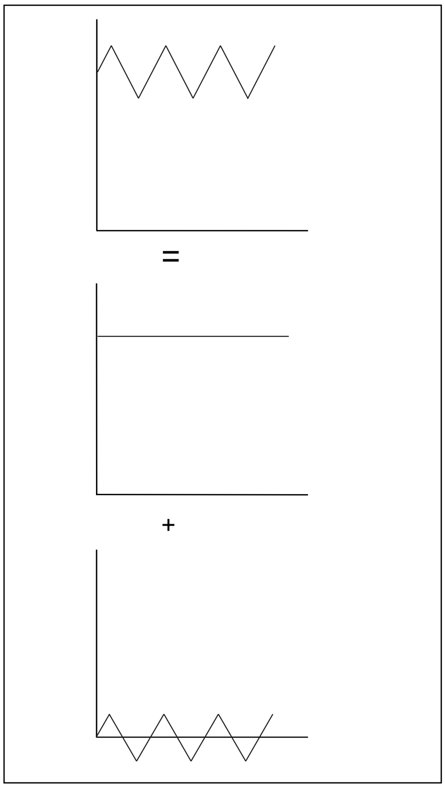

Fig. 20: Separation of the current's D.C. and A.C. components

The smaller r, the more F approaches 1 (as with a linear amplifier). The torque of a D.C. motor is the product of its torque constant "k" and its average current "I_{av}" : M = I_{av} k.

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The value of I_{av} is:

$$I_{av} = \frac{1}{T}\int_0^T i_{(t)} \cdot dt$$

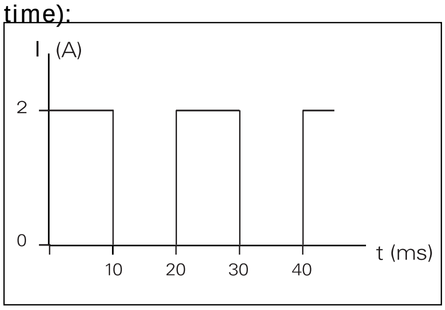
The rms current is:

$$I_{rms} = \sqrt{\frac{1}{T}\int_0^T i^2_{(t)} \cdot dt}$$

It's the rms current which determines the motor's power dissipation (Joule losses):

$$P_d = R_{mot} \times I_{rms}^2$$

It's apparent that the form factor should be close to 1 in order to limit motor heating. This is best illustrated by calculating an extreme case with a cycle of 2 A and 0 A during equal 10 ms pe-riods (neglecting current rise time):



$$I_{av} = \frac{\sum i_i t_i}{\sum t} = \frac{2 \cdot 10 + 0 \cdot 10}{10 + 10} = 1A$$

$$I_{rms} = \sqrt{\frac{\sum i_i^2 t_i}{\sum t}} = \sqrt{\frac{2^2 \cdot 10 + 0^2 \cdot 10}{10 + 10}} = 1.41A$$

Therefore, the motor torque corresponds to a current of 1 A and motor heating to a current of 1.41 A.

Similar to temperature, erosion of the commutation system also rises with the square of the rms current, and with the inductance of the commutated coil:

$$EE = L \cdot I^2_{rms}$$

The current rise due to using a coil of lower inductance, within a given motor series, does however not influence electroerosion. Within a motor series there is inverse proportionality between inductances and currents.

This is illustrated by comparing the current necessary for the same torque, and the inductance, of two motors of the 23V series:

23V58-216E:
L = 0.8 mH, I = 0.5 A, L×I² = 0.2
23V58-216P:
L = 0.2 mH, I = 1.0 A, L×I² = 0.2

To summarise the inconveniences of a chopper:

- the switching transitions can generate electromagnetic interference and parasites
- acoustic noise at the chopper frequency is generated in the driver, the motor, and even in the connecting wires. It is particularly disturbing if the frequency is in the 2 to 5 kHz range where the human ear is most sensitive
- motor current ripple can increase electroerosion and cause additional iron losses.

5.4 Criteria for selecting a chopper driver for ironless rotor motors

5.4.1 Influence of the inductance and chopper frequency

Ironless rotor motors have a relatively large air gap for the rotation of the cylindrical winding. Because of the low permeability of such a magnetic circuit motor inductance is of the order of only 1 mH if measured at 1 kHz.

Approximating the current ripple shape Δi with a triangle, we can use the following formula:

$$\Delta i = \frac{U_{alim}}{2 f_c \cdot L}$$

where f_c is the chopper frequency. The lower the inductance, the larger the ripple is. It could be lowered by increasing the chopper frequency, but due to its construction, the inductance of an ironless rotor motor varies with frequency. This lower inductance is due to the appearance of eddy currents in the motor tube and to the phenomena of hysteresis, which tend to diminish the permeability of the iron circuit. Induct-ance values of the 28DT12-222P motor measured at different frequencies are given in figure 21. The catalogue value measured at 1 kHz is 0.2 mH.

At usual chopper frequencies between 5 and 25 kHz, the low inductance of such motors may lead to excessive current ripple of 10% and more.

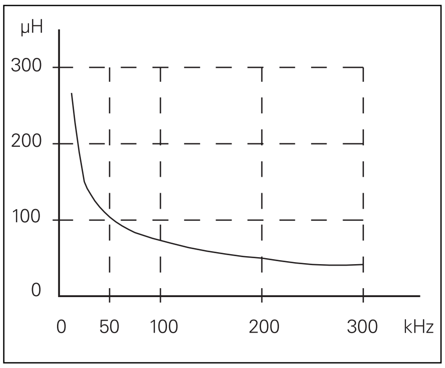


Fig. 21: Value of the inductance measured at different frequencies at the terminals of a 28DT12-222P motor

Simply raising the frequency does not always reduce ripple because of the simultaneous reduction of motor inductance. The problem can be overcome by adding an inductance externally to the motor. With chopper frequencies in the 5 to 20 kHz range, however, these additional coils are rather big and costly. Neglecting motor inductance, a coil of 1 mH gives, at 48 V and for a chopper frequency of 300 kHz, a ripple of 80 mA for any current value. Under the same conditions but with 20 kHz, the ripple is 1.2 A. The advantage of obtaining a lower motor temperature by raising the chopper frequency is paid for by external inductance, adapted to high frequencies and necessary to compensate the drop in motor inductance.

The voltage "L×di/dt" can, in applications of high dynamics, require an increase of the supply voltage, which is split between the motor and the added inductance. Furthermore this inductance may cause important arcing during micro-interruptions of the brush contact.

Besides lower iron losses, the benefit of low ripple for thermal conditions must be clearly understood for a motor that is stalled but energised, that means the controller requires zero current. In the preceding example, the actual motor current for "zero current" was ± 20 mA. Without the additional coil or with a lower switching frequency, ripple rises and so does the current actually obtain-ed for a "zero" reference.

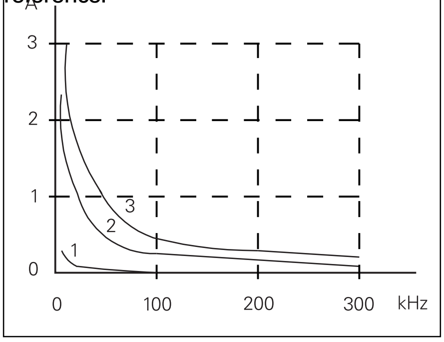


Fig. 22: Current ripple as a function of chopper frequency for three different inductances:
1 = 10 mH, 2 = 1 mH, 3 = 0.5 mH

The following formula can be used to calculate the inductance L_c of the external coil as a function of the inductance L_m of the motor, the chopper frequency f_c, the supply voltage U and the ripple tolerated:

$$\Delta i = \frac{U}{2 f_c (L_m + L_c)}$$

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There are also disadvantages linked to high chopper frequencies:

- complexity of the circuit commanding the transistors
- higher switching losses in the transistors but which remain low with MOS FET
- radio frequency emissions and electromagnetic interference (RFI and EMI).

Chopper frequencies of 200 or 300 kHz do not by themselves generate emissions. It is the rapid commutations necessary for switching at these frequencies (i.e. a high value of "dU/dt") that generate high frequency noise that needs to be suppressed by adequate filters. As the wire length between driver and motor increases beyond a few meters, these filters become a necess-ity. Note that the higher the frequency, the smaller the coil required for the filter. For reasons of dimensions and symmetry the inductance is usually split between two identical coils, one in each motor lead.

5.4.2 Precision of the current around its zero value

A position loop requires precise current control and a perfect linearity around the zero value, assuring zero current if the controller demands it. With a chopper driver this linearity depends on three interdependent factors:

- the current circulation mode
- the chopper frequency
- the motor inductance.

It was shown that the circulation mode determines current ripple, but it also influences the precision of current control. In non-regenerative mode ripple is low, but imposing zero current is not possible. In forced regenerative mode ripple is higher, but zero current is well under control. Therefore this is the preferred control mode, with ripple being limited by the procedures described in 5.4.1.

Data sheets of chopper amplifiers rarely specify the circulation mode. The form factor (if it is given at all) is only a measure of the ripple amplitude. In order to find out about the behaviour around zero current the current loop must be tested using a sinusoidal reference of low frequency. The response obtained allows to determine the circulation mode and, possibly, the type of action which could be taken to improve the behaviour.

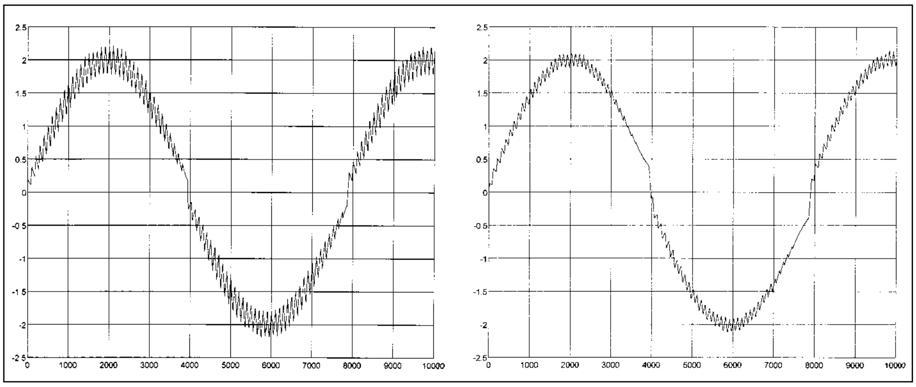

Fig. 23 :Two examples of current behaviour around zero

If the response is similar to the left half of figure 23, linearity around zero can be improved by adding an inductance and possibly raising the chopper frequency. If the response is as shown in the right half, nothing can be done. The distortion around zero is inherent to the non-regenerative mode, and that type of driver is not really adequate for precise positioning.

5.4.3 The supply voltage

As the basic function of a chopper driver is to reduce the motor terminal voltage, it can easily work on a higher supply voltage than a linear amplifier. This voltage depends mainly on the maximum duty cycle "D_{max}". A value of 100 % means that the chopper does not chop anymore because the voltage is insufficient. Neglecting the losses in the bridge, the voltage available for the load is:

$$U_{mot} = U_{supply} (2\ D_{max} - 1)$$

If we take into account the bridge dead time necessary between transitions to avoid simultaneous conduction, the available voltage becomes:

$$U_{mot} = U_{supply} (2\ D'_{max} - 1), \text{ with}$$

$$D'_{max} = D_{max} - 2\ t_m/T$$

t_m is the dead time and T, the chopper period.

As an example let's calculate the ripple for an inductance of 1 mH, a supply voltage of 24 V, a maximum duty cycle of 95 %, a dead time of 100 ns, and two chopper frequencies of 20 kHz, commonly used, and 100 kHz used in a modern card:

f	20 kHz	100 kHz
Δi	600 mA	120 mA
U_{mot}	21.4 V	20 V

The difference in ripple is impressive. The small difference in motor voltage is caused by the dead time which, at 100 kHz, has a larger influence.

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6. Conclusion

API Portescap works constantly to improve drive circuits and the implementation of new concepts. As to the choice between voltage control and current control for ironless rotor motors with low inductance there is little difference and there are few cases where the choice of the type of driver is critical. However, current control does limit peak current, and it forms a simpler first-order system with the motor.

The choice between linear amplifier and chopper is more delicate, and depends mainly on the application. For speed control a linear amplifier and a motor with precious metal commutation may be the preferred solution.

For position control with high dynamics, current control from a chopper amplifier and a motor with graphite brushes should be a better solution.

The example speed-torque curve of figure 24 illustrates this choice; our catalogue shows one for each motor size (left illustration). The maximum recommended speed is indicated by the hor-izontal line on top, maximum recommended torque by the vertical line to the right. The inclined hatching represents the continuous working range. The horizontal hatching represents the temporary working range. Between both, a vertical line indicates the continuous torque corresponding to the maximum continuous current before the motor overheats. With a chopper driver the entire area of this curve can be used.

For an application using a limited range of speeds and torques similar to the two examples at right, a linear amplifier can be the right choice, and it should not be associated with «low efficiency» all the time. By adapting the supply voltage to a particular working point, efficiency may be as good as with a chopper driver.

The power dissipation in a linear amplifier is approximately:

$$P_d = (U_{supply} - U_{mot}) \times I_{mot}$$

Efficiency of a chopper is, of course, good at any working point. Power dissipation in a chopper is approximately:

$$P_d = R_{equivalent} \times I_{mot}^2$$

The essential points to consider with each driver type for ironless rotor motors are listed below. For a linear amplifier these are:

- monitoring of thermal dissipation; the desired working range must be within the thermal limits
- adaptation of the supply voltage to the actual requirements to optimise efficiency.

For a chopper amplifier:

- choice of the chopper frequency, which is often rather low (from 10 to 30 kHz) and, if necessary, reduction of current ripple by adding external inductance
- consideration of the current circulation mode and of the form factor, which are important for motor heating and for precise current control
- avoiding limitations of the dynamic performance by choosing a high enough output voltage, to account not only for losses in the driver bridge but also for duty cycle effects.

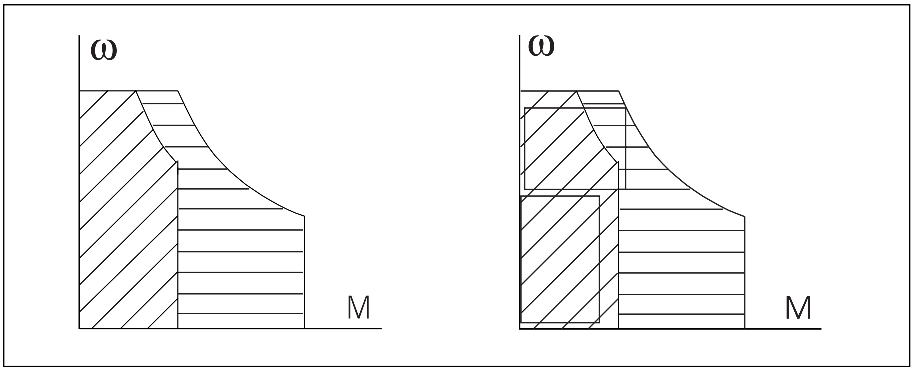

Fig. 24: Range of continuous and temporary motor operation (left), and two examples of actual working ranges in applications (right)

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7. Example driver

7.1 The ELD-3503 card

This API Portescap drive circuit for D.C. motors accepts supply voltages from 12 V to 35 V and delivers a continuous current up to 2.5 A, with 3.5 A peak. The internal voltage drop is about 6 V for any current. With a 12 V supply, for example, the motor terminal voltage is limited to 6 V.

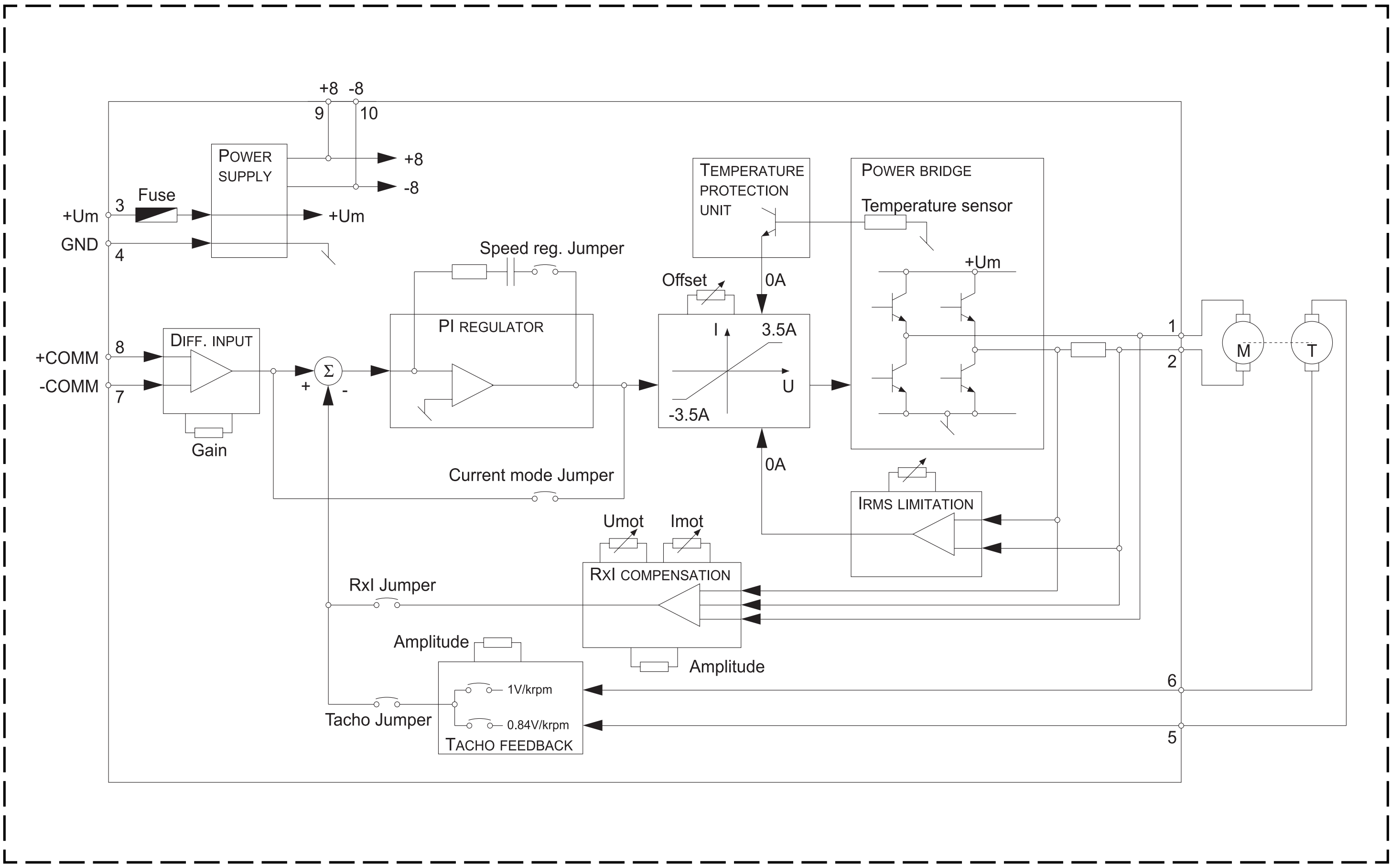
The card is intended for PI type speed control with a DC tacho or by R x I compensation, the bandwidth being 20 Hz. With a current loop activated by a jumper, position control is also possible after adding a sensor and a microprocessor control.

The linear transconductance amplifier uses an internal P type current loop with a bandwidth of 2000 Hz.

The card's dimensions are 80 x 100 x 30 mm, it is protected against short-circuits, overcurrent and overtemperature.

A "Suitability Test" program running under Windows™ is available to facilitate use of the ELD-3503. Depending on load conditions and on the selected motor, it calculates the values of voltage, current, power dissipation and temperature rise. It also includes data bases of available API Portescap D.C. motor lines.

Block diagram of the ELD-3503:



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