Application Note

An Introduction to Low Voltage DC Motors

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Abstract

An introduction to the fundamental principles of operation of low voltage DC motors, focussing on the common "brushed" and "brushless" constructions.

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Application Note

An Introduction to Low Voltage DC Motors

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Summary

This application note will describe the physical construction and electrical characteristics of typical "brushed" and "brushless" low-voltage DC motors, with particular attention paid to speed control of both motor types. The application note will then consider the relative merits of power MOSFETs when used in typical DC motor control circuits, and will give detailed guidance on choosing power MOSFETs for use in this application.

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1. BRUSHED DC MOTORS

An introduction to brushed DC motors.

Nearly all motors exploit the force which is exerted on a current carrying conductor placed in a magnetic field. This phenomenon is often demonstrated in classroom experiments where a bar magnet is placed near a wire carrying current, and the resultant force can be seen to distort the shape of the wire. Unfortunately, the force generated in such an experiment is miniscule and could hardly be used to do any useful mechanical work. In order to make a useful electric motor we need to arrange for there to be a strong magnetic field and for it to interact with many conductors, each carrying as much current as possible. The component parts of a motor are designed so that interaction of currents and magnetic fields produces a continuous, smooth rotational movement.

A brushed DC motor is the simplest of all motor types, and typically consists of the following parts;

Stator. The stationary part of the motor in which the rotor revolves. The stator typically takes the form of a metal can, open at one end, with two or more curved magnets mounted inside it. The stator housing often doubles up as the housing for the motor as a whole. See Fig. 1a.

Rotor. The rotating part of the motor, mounted axially in the centre of the stator housing. The motor windings are wound on the rotor. See Fig. 1b.

Brushes. A series of electromechanical contacts which enable current to flow to the rotating motor windings in the correct sequence and direction. The stationary brushes make electrical contact with part of the rotor known as the commutator. This arrangement creates the correct sequence of current through the motor coils as the rotor rotates (see Fig. 2.).

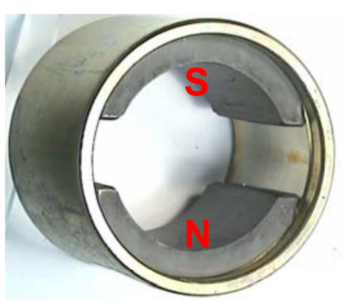

Fig. 1a. Brushed DC motor stator

Fig. 1b. Brushed DC motor rotor and brushes

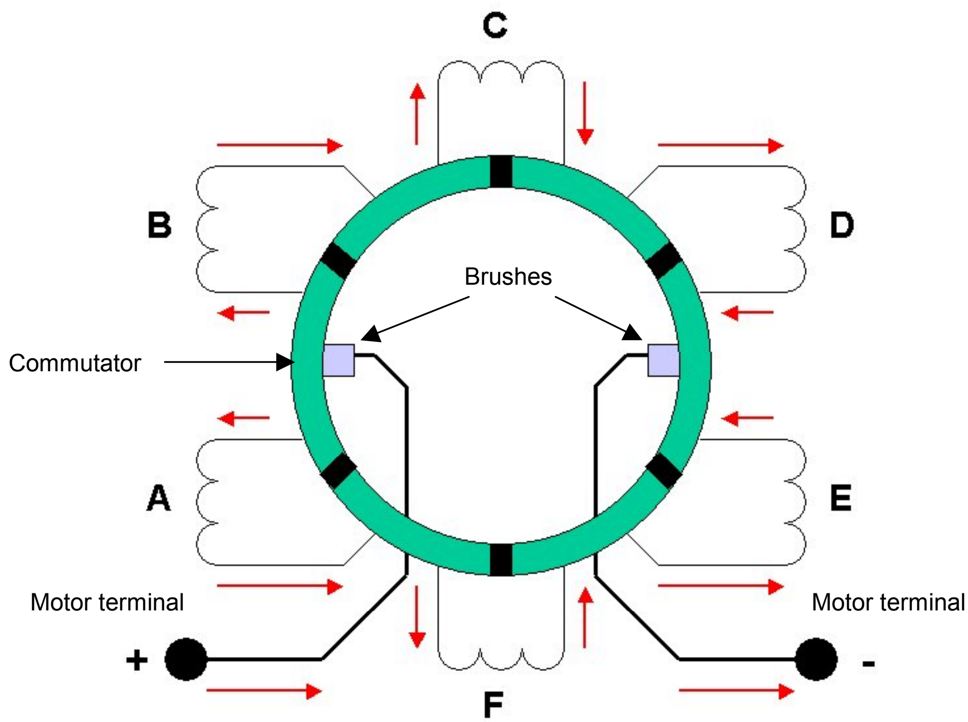

Fig. 2. Idealised brush and commutator arrangement in a brushed DC motor.

The commutator is usually in the form of a cylinder, consisting of segments of conducting material interspersed with, and insulated from one another by, an insulating material. Current flow is in through the left hand brush and out through the right hand brush, and is indicated by the red arrows. If the rotation is clockwise, it can be seen that one sixth of a revolution after the instant shown, the current in coils A and D will have changed directions. As successive commutator segments pass under the brushes, their current directions will also change. This “rotating” magnetic field interacts with the stationary field generated by the stator magnets, and the result is continuous rotational motion produced in the motor’s rotor.

Equivalent electrical circuit of a brushed DC motor.

The equivalent electrical circuit of a DC motor is show in Fig. 3.

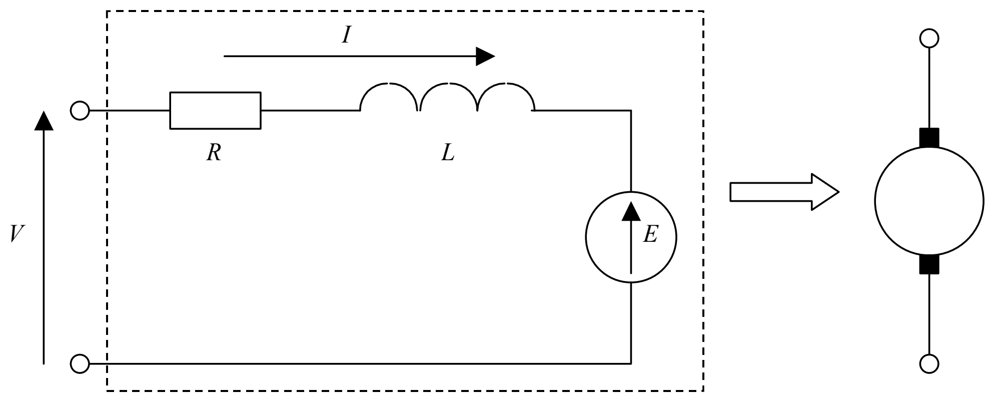

Fig. 3. The equivalent electrical circuit of a DC motor (left) and the commonly used circuit symbol for a DC motor (right).

Although a total of six distinct windings are shown in Fig. 1.2., the windings can be considered as one continuous larger winding. This is true for any number of configuration of windings in this type of motor. Referring to Fig. 1.3., "V" is the voltage applied to the motor terminals. "L" and "R" are the inductance and resistance of the motor windings respectively and "E" is the "motional e.m.f." developed internally across the motor windings. Motional e.m.f. arises as a consequence of the motor windings passing through the magnetic flux generated by the stator windings. When the unit is acting as a motor (rather than a generator), E always opposes V and is known as "back e.m.f."

The equation linking these terms is;

$$V = E + I.R + L.\frac{dI}{dt} \qquad (1)$$

Under steady state conditions, I is constant and this simplifies to;

$$V = E + I.R \qquad (2)$$

or

$$I = \frac{V - E}{R}$$

The "$I.R$" term in equation (2) is the loss that results in the motor windings due to the resistance of the windings. This is usually referred to as "copper loss" and results in a noticeable rise in temperature when the motor is running.

Voltage and current waveforms for a brushed DC motor controlled by a mechanical switch.

This circuit demonstrates the operation of the brushed DC motors using a simple mechanical switch as a control device (Fig. 4.) Motor current is measured by observing the voltage appearing across the 10mΩ shunt resistor.

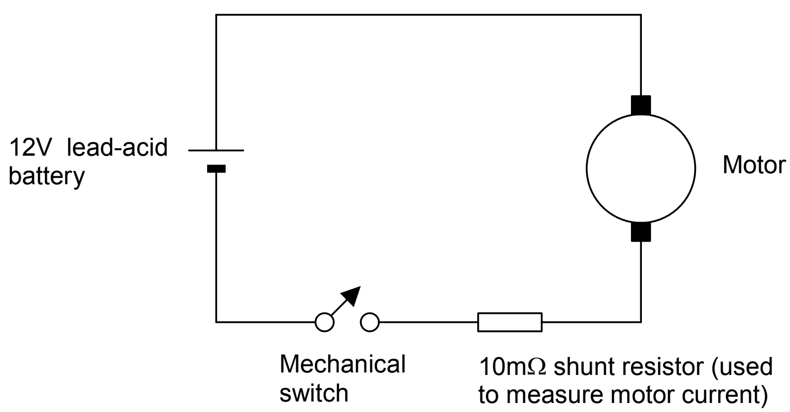

Fig. 4. Controlling a DC motor with a simple mechanical switch.

Current drawn by brushed DC motors at turn-on.

Fig. 5. demonstrates the current drawn at turn-on by the three ratings of motor. In all three cases the motors had no external mechanical load connected i.e. the only mechanical resistance to movement being internal friction of the motor bearings etc. Current was measured by measuring the voltage across the shunt resistor of Fig. 4.

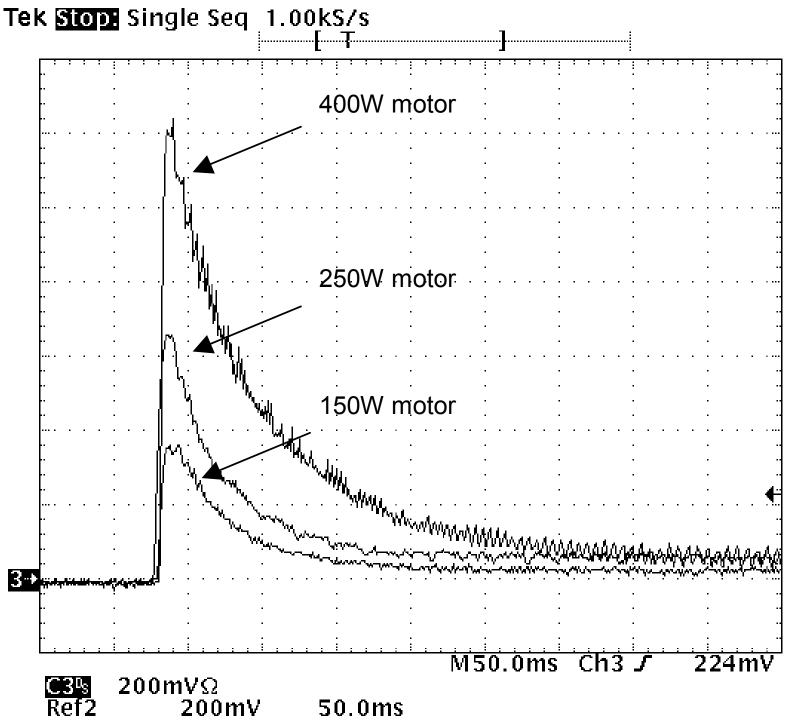

Fig. 5. Current drawn by 150W, 250W and 400W DC motors at turn-on under no-load conditions. Note that the vertical scale is 20A per division.

Considering equation (1) above, it should come as no surprise that a motor initially at rest should draw a large current when voltage is first applied to its terminals, since E is zero (no motional e.m.f.) and I is limited only by R and L. As the motor picks up speed, E starts to increase and opposes the flow of I, hence I rapidly falls from its initial high value towards a steady state, which itself depends on the mechanical load applied to the motor.

It should be clear from Fig. 5. that initial current surge drawn by a brushed DC motor, even in a no-load condition, can be enormous - as much as 120A for the 400W motor, and this surge current could cause excessive temperature rise in a semiconductor device driving the motor (see Fig. 6.) Management of such motor “in-rush” currents is an important part of motor control design, and will be considered in detail later in this article.

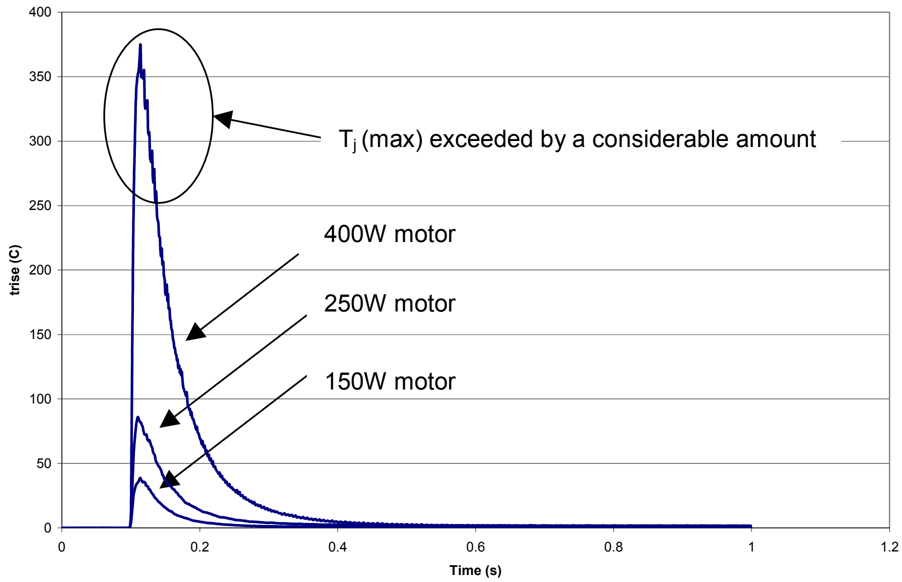

Fig. 6. Plot of junction temperature rise for 150W, 250W and 400W motors (no load) controlled by a BUK7528-55A

Current drawn by DC motors in the steady state.

Fig. 7. shows plots of motor current for the 150W motor for various levels of applied external load.

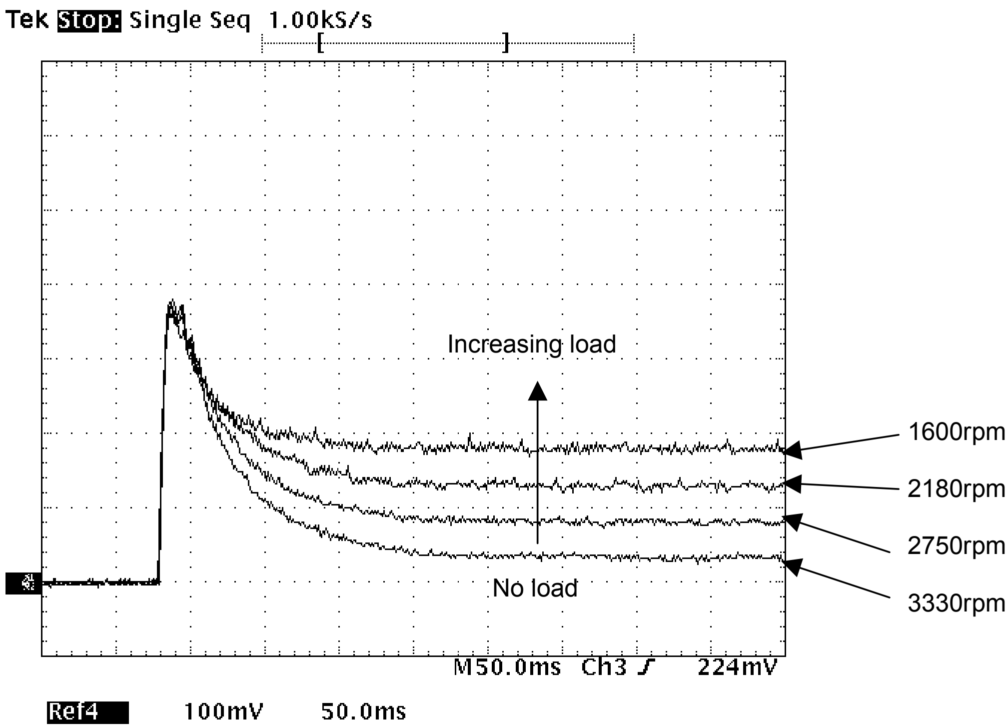

Fig. 7. Current drawn by 150W DC motor for various load conditions. Note that the vertical scale is 10A per division.

Although each current plot starts with a similar initial surge, the plots "flatten out" to different levels depending on the amount of external load applied. Increased load results in a higher steady state current and reduced motor speed. This is entirely consistent with equation (1.2) above. Assuming that applied voltage V stays constant under all load conditions, as load increases, so I increases and hence the $I.R$ term increases. At the same time the motor speed decreases and hence E (which is proportional to motor speed) decreases. In this way the equation remains balanced.

An important implied consequence of Fig. 7. is that, for a motor which is completely stalled (i.e. rotor locked in a stationary position), the motor current will stay at its initial peak value until the power source is disconnected. This situation could have serious power dissipation consequence for any power semiconductor which is being used to control the motor.

Current drawn by DC motors at turn-off.

The current drawn by the 150W motor at turn-off is shown in Fig. 8. At the time indicated, the mechanical switch is operated and the motor current ramps down to zero within approximately 200µs. Mechanical switches typically take several milliseconds to fully open, and within the timeframe considered the switch is in an "in-between" state where it is still able to provide a path for current; the energy stored in the inductance of the motor windings and external wiring "forces" the flow of current through the switch when it is in this indeterminate state, and approximately twelve volts appears across the switch contacts (note that, due to the switch orientation in the circuit, the displayed waveform is increasing in magnitude in the *downward* direction). The switch contacts are subject to arcing during this time – a destructive process which can drastically shorten the life of a switch or relay contacts. A power MOSFET would provide a much more reliable and long-lived method of switching the motor current.

When the supply current is removed from the motor it does not immediately cease rotation. Rather, being a mechanical system with momentum, the motor comes to a standstill relatively slowly. During this time it is acting as a generator and a voltage appears at the motor terminals.

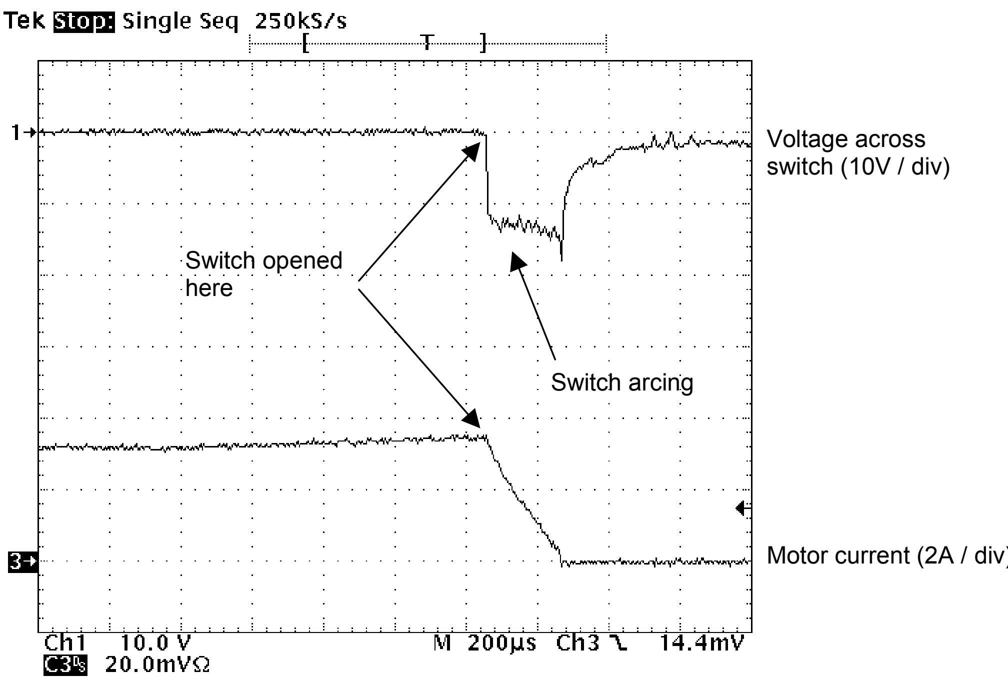

Fig. 8. Current drawn by 150W DC motor and voltage across the switch at turn-off.

2. SPEED CONTROL OF BRUSHED DC MOTORS.

<u>Introduction.</u>

The simplest form of speed control of a brushed DC motor can be achieved by applying a variable DC voltage to the motor terminals. It can be demonstrated that, for a constant mechanical load, the motor speed varies linearly with supply voltage. See Fig. 9.

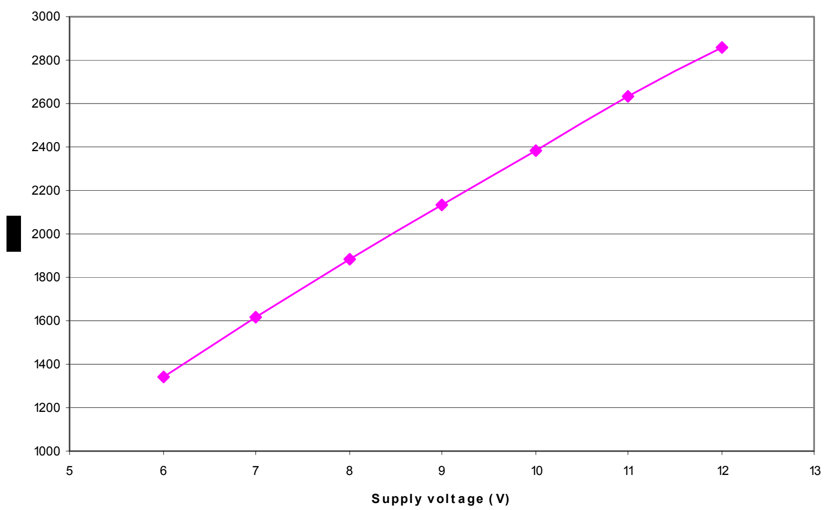

Fig. 9. RPM v Supply voltage for 150W brushed DC motor (no load).

Unfortunately, such an arrangement suffers from two serious disadvantages;

Efficiency. If the motor supply voltage is derived from a DC voltage source and variable resistor (in effect, a crude linear regulator) then, as motor speed is decreased, increasingly more power will be wasted as heat dissipation in the variable resistor. For all but the smallest of motors, this scheme would be impractical.

Speed regulation. The graph of Fig. 9. demonstrates a good linear relationship between supply voltage and motor speed – *provided the load on the motor remains constant*. If, however, we apply a particular supply voltage in order to set a specific motor speed, and then the load on the motor changes, it will be found that the motor speed also changes. Generally, if the supply voltage remains fixed, then motor speed decreases with increasing load. This is demonstrated in Fig. 10.

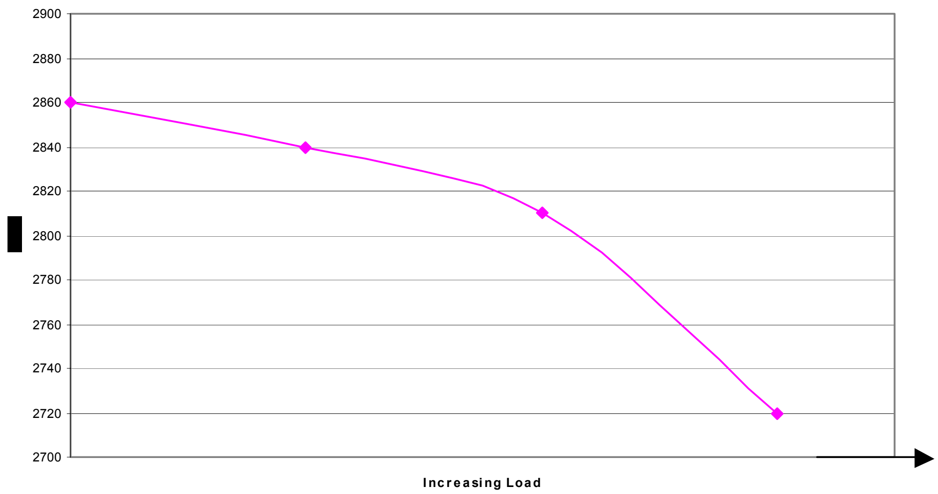

Fig. 10. Variation of RPM with load for a fixed supply voltage (150W brushed DC motor).

Why does motor speed vary with load?

In order to answer this question, we need to consider the simple motor model introduced earlier. See Fig. 11.

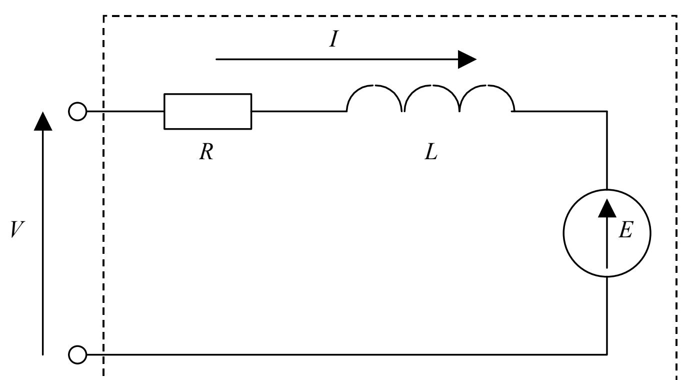

Fig. 11. The equivalent electrical circuit of a DC motor.

Assuming the motor is stationary before voltage is applied to the terminals, when voltage is applied, E will be zero and hence I will be limited only by R and L. As the motor builds up speed, E begins to increase, reducing the voltage across R and L and hence also reducing I. If the motor windings were made of perfect conductors, with no resistance, and the motor had zero friction in its bearings, then we would eventually reach a point where $E = V$ and $I = 0$. All that would be required to keep the motor running in this condition would be to maintain the voltage V at the motor terminals. Of course, real motors have both winding resistance and friction in their bearings, so even in a "no load" condition, I is greater than zero. Another way to look at this is to consider the motor as an "energy conversion" mechanism where;

electrical energy in = mechanical energy out + energy dissipated in bearings and windings.

If we now apply some load to a motor running in a "no load" condition, the speed of the motor will decrease. As a result, E will also decrease and hence I will increase. As I increases, the motor speed increases again and so E increases. In this way, the motor will quickly reach an equilibrium speed once more. Although this "self regulation" mechanism will go some way to restoring the original motor speed under the new load condition, it cannot compensate for the voltage dropped across (and power dissipated in) the motor winding resistance. Hence we see the motor speed drop with increasing load. Motors with higher winding resistance will tend to have worse speed regulation under load, as well as being less efficient – dissipating more power and running hotter for a given output power.

<u>Brushed DC motor speed control using a "chopper" circuit.</u>

The "chopper" speed control topology is the simplest brushed DC motor control topology, and is shown in its simplest form in Fig. 12.

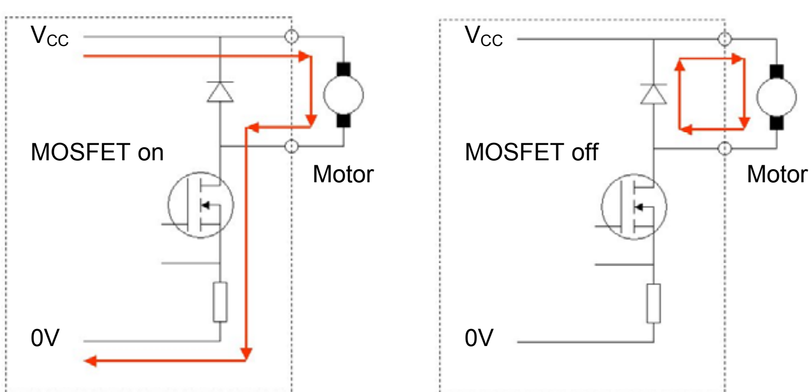

Fig. 12. Chopper control of a brushed DC motor. MOSFET conduction cycle is shown on the left, diode "flyback" conduction cycle on the right.

This topology bears more than a passing resemblance to a flyback SMPS, and the mode of operation is in fact quite similar. When the MOSFET turns on, the full voltage appears across the motor and the inductance of the motor windings causes the current through motor and MOSFET to ramp up. When the MOSFET turns off, the inductance of the motor windings forces the diode into conduction, with the diode anode (and MOSFET drain) biased at a diode drop above V_{CC}. During this time the current is ramping down. After a period of time, the MOSFET turns on again and the cycle repeats. Typical current and voltage waveforms for the motor and MOSFET are shown in Fig. 13. By varying the duty cycle of the MOSFET it is possible to vary the motor speed. It can be easily demonstrated that the motor speed due to the *average* voltage across the motor in a chopper circuit correlates well with the motor speed when the voltage is applied from a DC source.

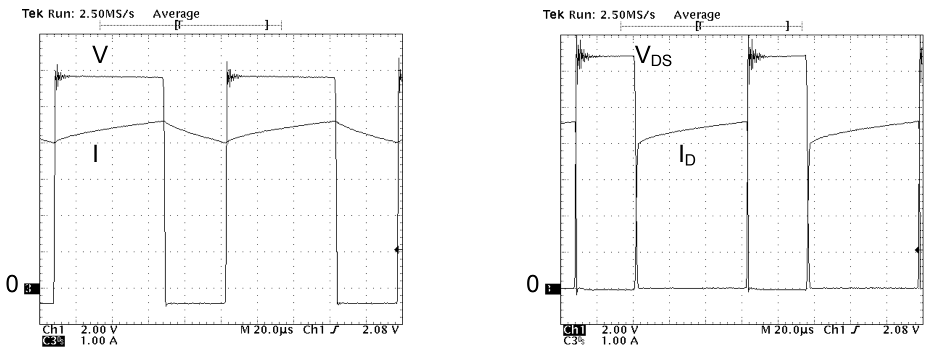

Fig. 13. Voltage and current waveforms for the motor (left) and MOSFET (right).
Current is 1A/div, voltage is 2V/div.

This topology shares the same efficiency benefit as its SMPS counterpart – the MOSFET is either hard on or hard off, and hence power dissipation in this component is kept very low. In addition, the switching frequency of such a circuit is much lower than that of an SMPS, typically 5 – 10kHz, so switching dissipation in the MOSFET is negligible. See Fig. 14. The only MOSFET loss of concern is on-state loss. The MOSFET drain is clamped at diode drop above V_{CC} by the diode, so the MOSFET V_{DS} rating need only be sufficient to deal with any ringing in the circuit at turn-off due to stray inductances – see right-hand oscillogram in Fig. 14. A 55V or 60V rated device is typical for a chopper circuit with a 12V input.

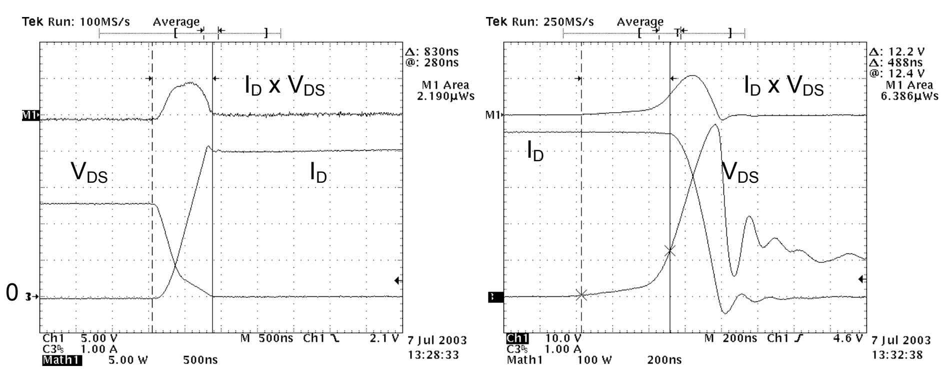

Fig. 14. Turn-on and turn-off loss for the MOSFET. Turn-on loss = 2.19µJ, turn-off loss = 6.36µJ. At a frequency of 10kHz this equates to a total switching loss of 85mW.

It is reasonable to wonder whether the surge current drawn by a motor at switch-on when controlled by a simple switch (see earlier) is also present when the motor is controlled by the more complex chopper circuit. It should be clear that the presence of such a surge could be damaging to the chopper MOSFET due to transient over temperature conditions which would be caused in this device, and hence the circuit design should seek to eliminate surge conditions. In the chopper circuit examined, a “soft start” technique has been employed, in that the motor is set in motion relatively slowly by periodic pulses of current, and large current spikes are avoided by initially only allowing the MOSFET to conduct for short periods of time. This is illustrated in Fig. 15.

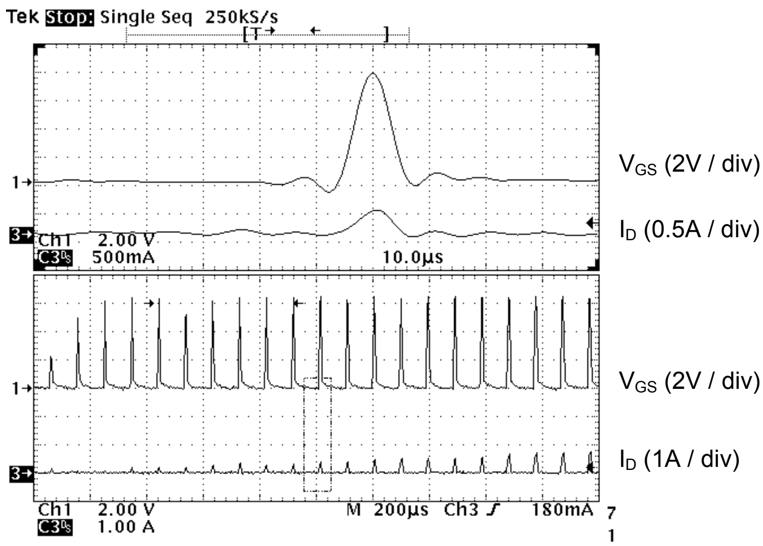

Fig. 15. MOSFET V_{GS} and I_D waveforms at start-up. Lower picture shows V_{GS} and I_D for the first few cycles at start-up, upper picture shows the detail of one cycle.

Whilst the chopper topology has greatly improved efficiency compared to linear regulation schemes, it has the disadvantage of being able to provide only limited speed regulation with load variation. The chopper circuit of Fig. 12. has only two connections to the motor – to the supply terminals – and has no way of determining the speed of the motor. As such, the speed regulation of the circuit is not good as it cannot compensate for the resistance of the motor windings. Some chopper circuits do allow a degree of speed compensation, whereby the MOSFET “on” time is increased slightly as motor current increases, to compensate for the increased voltage drop across the motor windings. A compensation scheme such as this needs to be manually “tuned” to the characteristics of a particular motor, and can provide an (albeit limited) method of speed compensation. The effects of speed compensation in a typical chopper circuit are shown in Fig. 16.

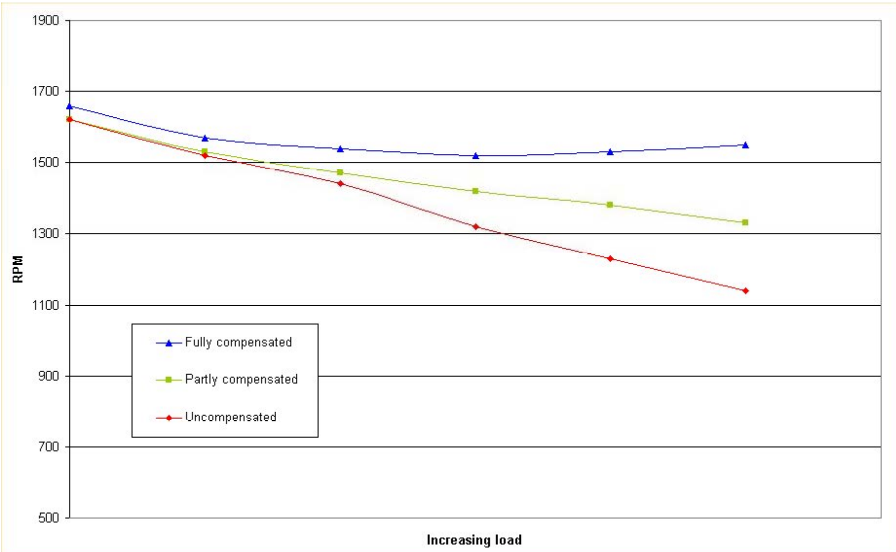

Fig. 16. The effect of speed compensation in simple chopper speed controller.

Brushed DC motor speed control using a full-bridge circuit.

The next level of sophistication in brushed motor control is represented by the full-bridge circuit (Fig. 17.). As with the chopper circuit, this topology bears a passing resemblance to a common SMPS topology, and there are some similarities in mode of operation.

The full-bridge topology has the advantage that it allows forward, reverse and braking control of the motor. Its main disadvantages are increased cost and complexity compared to simpler control schemes – particularly with regard to the need for floating gate drives for Q1 and Q3 (or choice of P-channel devices for these positions, which incurs extra cost per $R_{DS(on)}$) and the added complexity of driving four separate MOSFETs. Typically, a microcontroller is used as the controlling element in such a circuit.

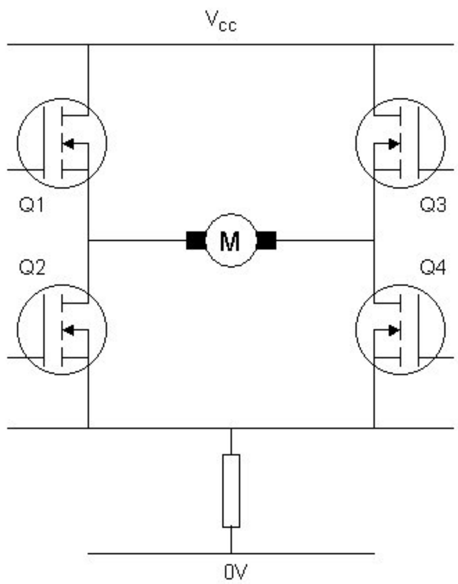

Fig. 17. Full-bridge control of a brushed DC motor.

Note that the resistor connected to Q2 and Q4 sources is used for current sensing. Operation in the forward direction is shown in Fig. 18.

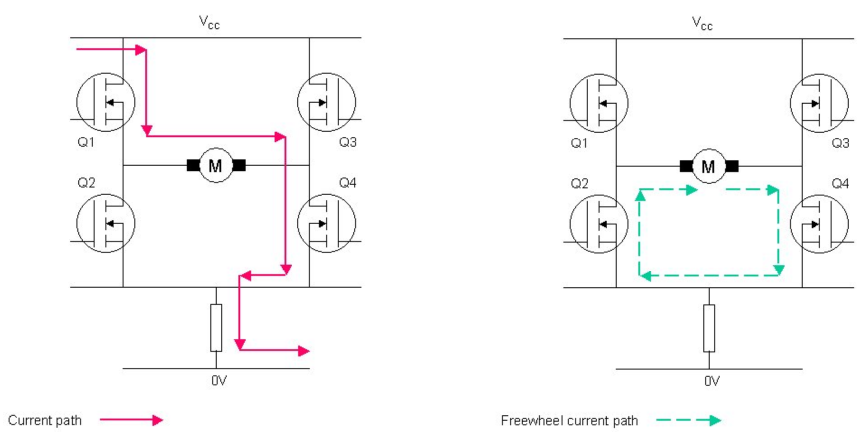

Fig. 18. Full-bridge operation in the forward direction.

For operation in the forward direction, Q4 is permanently turned on, whilst conduction alternates between Q1 and Q2. Braking is achieved by leaving Q4 on, turning Q1 off and turning Q2 on. The energy stored in the motor is dissipated during braking in Q2 and Q4. Similarly, for operation in the reverse direction, Q2 is permanently turned on, whilst conduction alternates between Q3 and Q4. Braking in the reverse direction is achieved by leaving Q2 on, turning Q3 off and turning Q4 on.

An equally valid mode of operation is to have one of the high-side MOSFETs (Q1 or Q3) turned permanently on whilst conduction alternates between the two opposite MOSFETs. So, in the forward direction, Q1 would be permanently on with conduction alternating between Q4 and Q3 body diode, and in the reverse direction Q3 would be permanently on with conduction alternating between Q2 and Q1 body diode. Braking is achieved by turning on Q1 and Q3. In this mode, Q1 and Q3 are often referred to as “steering” MOSFETs. The advantage of this mode is that the steering MOSFETs are *not* required to switch at frequencies of several kHz, and hence high-side TOPFETs may be used – thereby eliminating the need for additional floating gate drive circuitry.

As with the SMPS full-bridge topology, there is a need to avoid “cross conduction” between upper and lower MOSFETs. In the forward direction, Q1 and Q2 must never be on at the same time and in the reverse direction Q3 and Q4 must also never be allowed to cross conduct. Cross conduction is avoided by allowing a “dead time” between (for instance) Q1 turning off and Q2 turning on, or Q2 turning off and Q1 turning on. During the dead time, the appropriate MOSFET body diode is instead allowed to conduct. See Fig. 19.

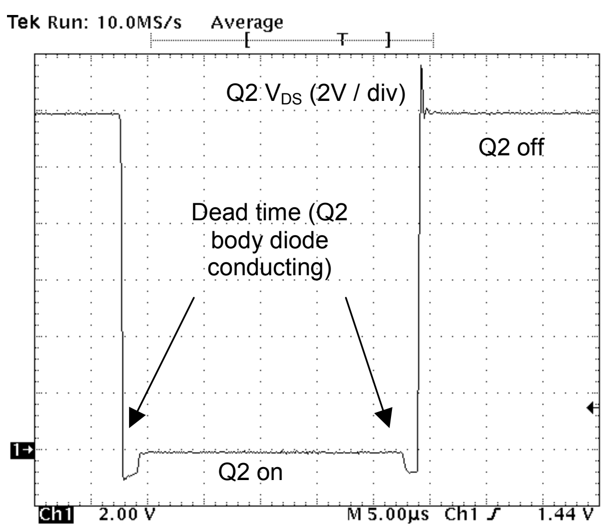

Fig. 19. Q2 V_{DS} waveform showing dead time.

In common with the chopper circuit, the MOSFETs in the full bridge are either hard on or hard off, and hence power dissipation in these components is very low. The switching frequency in full bridge circuits is also in the low kHz range, so switching dissipation in the MOSFETs is negligible. MOSFET loss in the full bridge is therefore dominated by on-state loss dissipation. The MOSFET drains are clamped at voltages close to V_{CC}, so the MOSFET V_{DS} ratings need only be sufficient to deal with any ringing in the circuit at turn-off due to stray inductances.

3. **AN INTRODUCTION TO BRUSHLESS DC MOTORS.**

Introduction.

Brushless DC motors utilise the same electromagnetic phenomenon as brushed motors in order to produce mechanical rotation - the force which is exerted on a current carrying conductor placed in a magnetic field. A brushless motor is mechanically simpler than a brushed motor and typically consists of two main parts;

Stator. The stationary part of the motor around which the rotor revolves. The stator typically takes the form of a central, axially mounted assembly which houses the motor windings, electronics and bearings which hold the rotor in place.

Rotor. The rotating part of the motor, mounted outside of the central stator. The rotor usually holds the motor's permanent magnets.

One way to consider the construction of a brushless motor is as an "inside out" brushed motor. In a brushed motor, the rotor (rotating part) holds the motor windings and the stator (stationary part) holds the motor magnets. In a brushless motor, this arrangement is reversed. The motor coils are stationary, the magnets rotate and the rotor is on the *outside*. An idealised cross-section of a typical brushless motor is shown in Fig. 20. Note that the control circuitry and power switches may also be housed in a separate module. An easy way to visualise the operation of a brushless motor is to say that the rotating rotor magnets "chase" the magnetic fields set up by stationary stator coils.

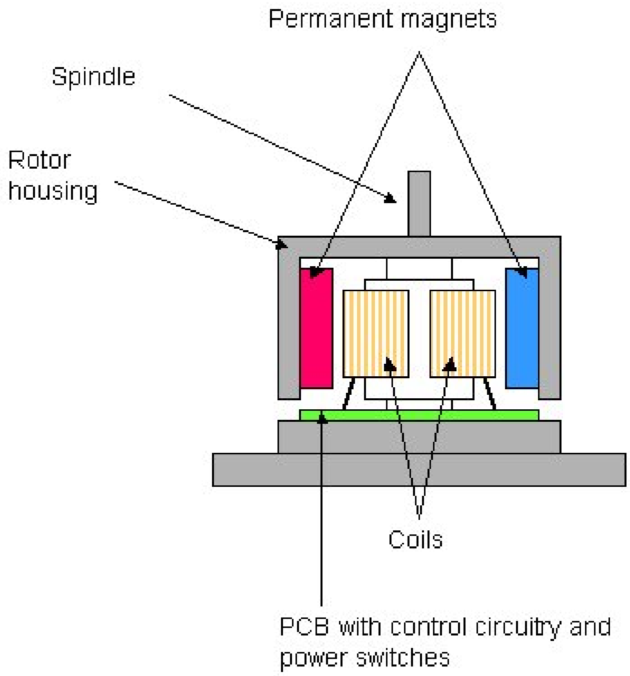

Fig. 20. Idealised cross-section of brushless DC motor.

An important difference between brushed and brushless motors is that the mechanical brushes of the former are eliminated (hence the name “brushless”) and all switching of motor currents is done electronically. The absence of brushes gives the brushless motor several notable advantages over its brushed counterpart;

1. Absence of arcing between contacts results in electrically quieter operation.
2. Motor life is considerably longer as there are no brushes to wear out.
3. Absence of mechanical friction allows for more efficient and smoother operation.
4. Much higher rotation speeds are possible – up to three times faster than for an equivalent brushed motor.
5. Good speed control is easily achieved.

These advantages dictate that brushed motors tend to be found in applications which require smoother, “cleaner” operation and long service life. Such applications include drive motors within PC disk drives, CD and DVD players, medical applications, precision instruments, lathes, etc., etc. Also, virtually every low voltage cooling fan in production today utilises some form of brushless construction.

Of course, the advantages offered by brushless motors come at a price, and this is mainly in the cost and complexity of the electronics which control the flow of current through the motor windings. A *brushed* motor is essentially “self driven” in that the flow of current in the rotor coils is determined by the mechanical construction of the motor and its rotational movement. In contrast, a brushless motor requires a series of semiconductor power switches to direct the current flow through the appropriate windings, as well as one or more position sensors (usually Hall effect devices) to determine the position of the rotor at any given time. Typically, a sub-10W motor will have two windings and a single position sensor – this is the arrangement commonly found in the cooling fans mentioned above. Motors with power ratings of more than ~10W will have three motor windings ("three-phase") and three position sensors. The control of these six elements is a complex task best achieved with an ASIC. The ASIC may also incorporate the power switches, in the case of a motor running at currents of a few amps. For higher currents (and hence higher switch dissipation), the power switches will usually take the form of external power MOSFETs.

4. DRIVING A THREE-PHASE BRUSHLESS DC MOTOR.

A three-phase motor requires a total of six power MOSFETs in its drive circuitry. See Fig. 21.

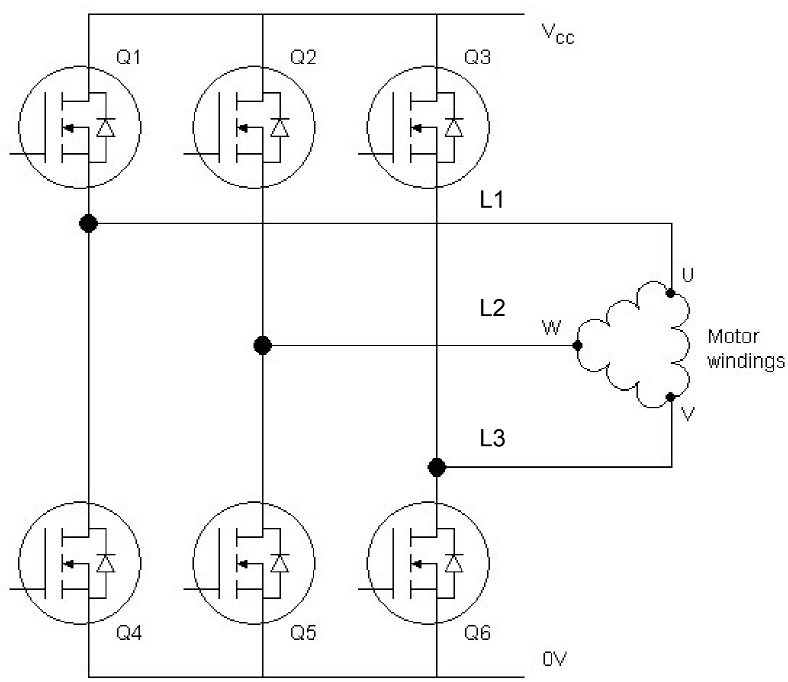

Fig. 21. MOSFET control of three-phase brushless DC motor windings.

This topology is often referred to as a "three phase bridge". There are numerous possible sequences of MOSFET operation, depending on the construction of a given motor. One typical sequence is shown in Fig. 22. This timing sequence was taken from a Papst "Variodrive" brushless DC motor, part number VD-3-54.14. This is a modern, "off the shelf" unit from a contemporary motor manufacturer, and is considered to be representative of the present "state of the art". The VD-3-54.14 will be used in all the following analysis. Note that current flowing into the motor windings is shown as positive current flow, current flowing out of the motor windings is shown as negative current flow.

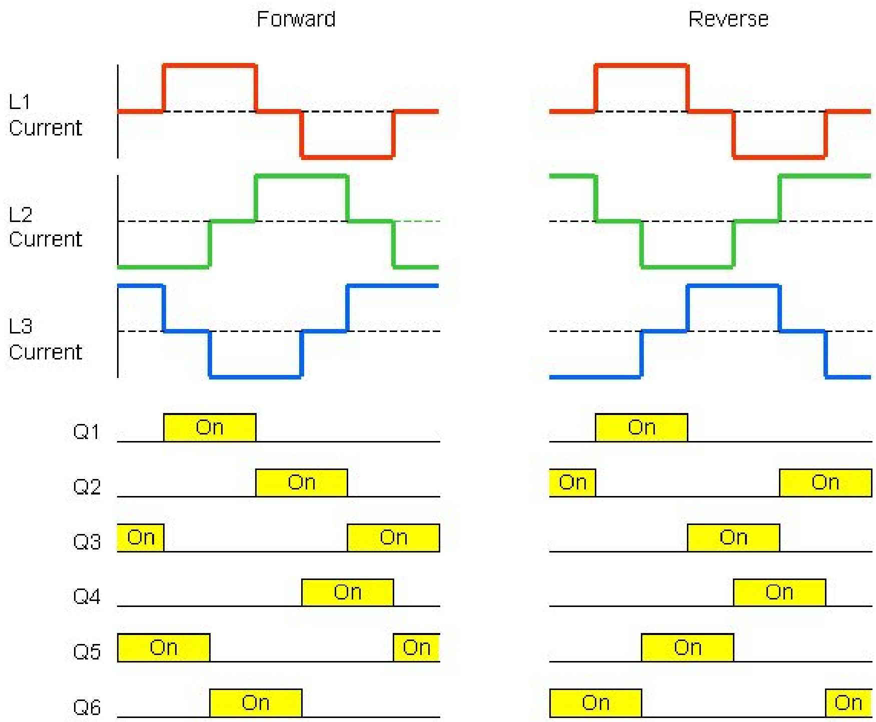

Fig. 22. Current waveforms for a Papst “Variodrive” brushless DC motor, part number VD-3-54.14.

In order that current is applied to the appropriate winding at any given time it is essential that the rotor position is always known. To determine the position of the rotor, a series of Hall effect sensors is used. The rotor is effectively a rotating series of magnets, and the Hall effect sensors are triggered by the proximity of these magnets. See Fig. 23. Braking is achieved by simply turning on MOSFETs Q4, Q5 and Q6, thereby shorting the motor windings to 0V.

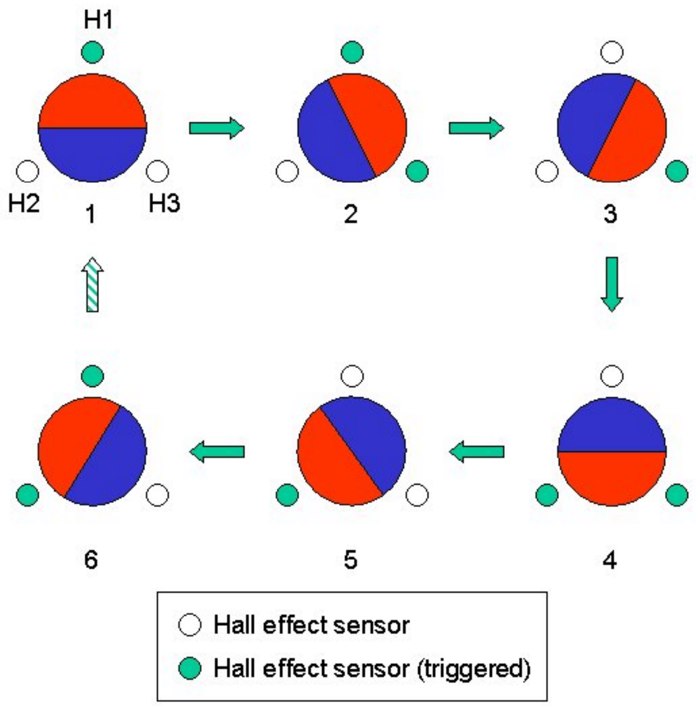

Fig. 23. Hall effect sensors used to determine rotor position.

It can be seen from Fig. 23. that each rotor position is associated with a specific pattern of triggered Hall effect sensors. Controller ASICs usually have on-board logic which is able to decode the patterns of sensor information and hence turn on the appropriate MOSFETs in the bridge circuit. Other arrangements of rotor and sensors are in common usage – the rotor may consist of more than the two magnetic poles shown in Fig. 3.4., and the sensors may be spaced at 30, 60 or 90° rather than the 120° shown. An added benefit of using rotor position sensors is that the controller circuitry can easily produce a pulse train output whose frequency is directly proportional to true motor speed – a feature offered by many of the presently available control ASICs. This feature is usually referred to as a “Tachometer” or “Tacho” output.

Current drawn by brushless DC motors at turn-on.

Each of the brushless DC motor windings may be represented by the circuit shown in Fig. 3., with the same electrical characteristics. “R” and “L” are the winding resistance and inductance, whilst the motional EMF due to the movement of the rotor magnetic fields through the winding is shown as voltage source “E”.

As is the case with the brushed motor, each of the brushless motor windings will conduct large currents when the motor is stationary as there is no motional EMF to oppose the flow of current. It is therefore necessary to employ similar “soft start” techniques to ensure that damagingly high currents are not conducted through the controller circuit power switches at start-up. Typically, the motor is set in motion relatively slowly by periodic pulses of current, and large current spikes are avoided by initially only allowing the bridge MOSFET to conduct for short periods of time. This is illustrated in Fig. 24. where the waveforms for motor current L3 and MOSFET Q6 gate waveform (from Fig. 21) are shown.

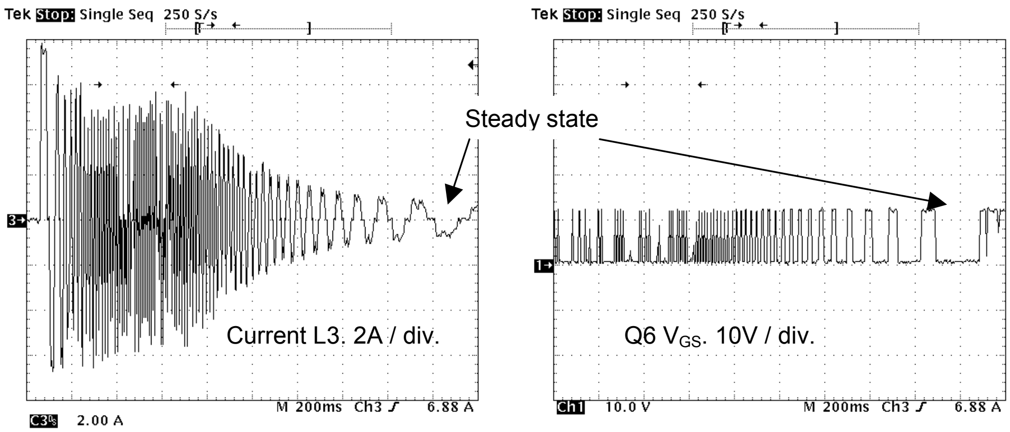

Fig. 24. Motor current and MOSFET gate drive at start-up.

MOSFET operating conditions in the three phase bridge circuit.

In common with the chopper and full bridge circuits, the MOSFETs in the three-phase bridge are either hard on or hard off, and hence power dissipation in these components is very low. The switching frequency in three-phase bridge circuits is also in the low kHz range, so switching dissipation in the MOSFETs is negligible. MOSFET loss in the three-phase bridge is therefore dominated by on-state loss dissipation. The MOSFET drains are clamped at voltages close to V_{CC}, so the MOSFET V_{DS} ratings need only be sufficient to deal with any ringing in the circuit at turn-off due to stray inductances.

5. BRUSHED AND BRUSHLESS MOTORS COMPARED

Characteristic	Motor Type	
	Brushed DC	Brushless DC
Mechanical complexity	High – incorporates a complex “brush arrangement” in addition to rotor and stator.	Low – essentially only two main mechanical components.
Electrical complexity	Low – current switching is carried out by “self driven” mechanical switches (the “brushes”).	High – controller circuit must generate three “quasi sinusoidal” current waveforms which are at 180° phase angles. Controller must also continuously sense the position of the rotor in order to ensure correct current switching.
Reliability	Low – continuous friction inevitably causes brush wear with subsequent degradation of performance and eventual failure of the motor.	High – a typical brushless motor can operate for 20,000+ hours without maintenance.
Controller complexity	Can be very low – the simplest PWM speed controller does not even require a dedicated control IC. More complex H-bridge controllers offer increased flexibility – forward, reverse and braking functions – at the cost of increased component count and complexity.	High – unlike brushed motors, brushless motors cannot be simply connected to a DC supply – they always require a complex control circuit to operate. This is usually in the form of a three-phase bridge with Hall effect sensors to monitor rotor position. The co-ordination of the control elements usually requires a dedicated ASIC.
Quality of motion	Quite low – the necessary friction between brushes and rotor results in an uneven feel to the motion.	High – with good quality bearings the motion of the motor can be extremely smooth.
Power losses and efficiency	“Copper losses” in windings, electrical losses in brushes, mechanical losses due to brushes and bearings.	“Copper losses” in windings, electrical losses in control circuit power semiconductors, mechanical losses due to bearings.
	At lower power levels, friction from the brushes tends to make the brushed motor less efficient. At higher powers, copper losses dominate and, for motors with similar windings, there is less difference in efficiency between the two motor types.	
Electrical noise	Can be high – especially when brushes have worn and arcing occurs.	Low – no mechanical switches to produce arcing.
Maximum RPM	Typically 16,000 – 20,000	Typically 30,000 – 60,000

6. A COMPARISON OF BIPOLAR AND MOSFET POWER SWITCHES.

Introduction.

This section will highlight the main differences between bipolar and MOSFET devices when used in typical DC motor drive applications. The comparison will focus on the following parameters;

Ease of drive.

On-state losses.

Paralleling of devices for increased power handling.

Utilisation of the MOSFET integral “body diode” in bridge circuits.

The comparison will demonstrate that in typical low voltage, medium power circuits, MOSFETs are a far superior choice of switching device.

Ease of drive - switching behaviour compared.

There is a fundamental difference in the way that MOSFET and bipolar devices are switched on and off;

To turn a MOSFET on it is necessary to charge up the MOSFET gate-source capacitance to a particular voltage level. Once this voltage level is reached, the MOSFET is in the “on” state and the circuitry driving the MOSFET gate needs do no more work, other than ensuring that the charge in the MOSFET gate-source is not allowed to leak out. Similarly at turn-off, the drive circuitry only needs to discharge the small amount of charge held on the gate-source capacitance and subsequently ensure that the MOSFET gate is held at 0V potential. See Fig. 25. (right).

A bipolar transistor is turned on by causing a current to flow from the device base to emitter – the base-emitter junction is “forward biased” and behaves in a similar manner to a diode. The forward biasing of the base-emitter must continue for the entire time that the transistor is turned on otherwise the device will turn off again. See Fig. 25. (left). At turn-off, the drive circuitry must allow a large current to flow out of the base-emitter if a fast turn-off is to be achieved.

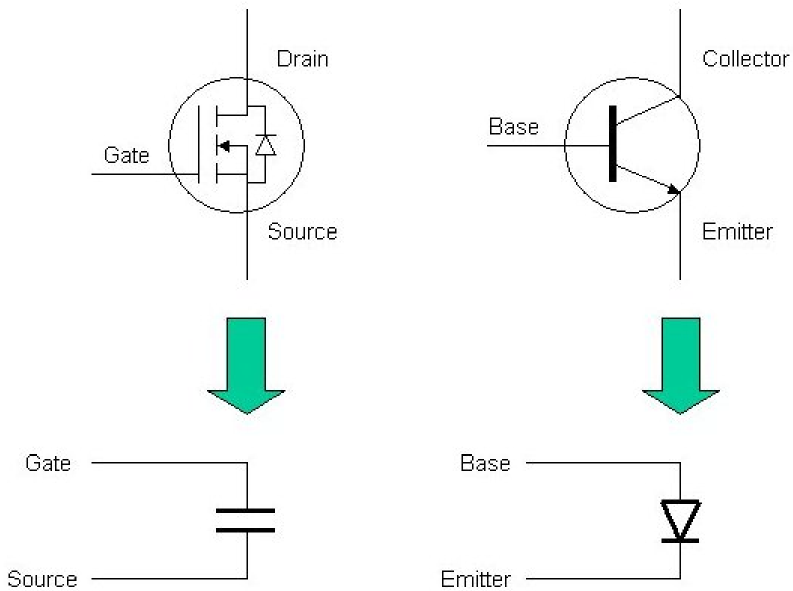

Fig. 25. Equivalent circuit of a MOSFET gate-source (left) and bipolar transistor base-emitter (right).

There are two main consequences of the switching behaviour of a bipolar device;

The need to supply continual base-emitter current ("I_B") results in increased dissipation in the external driver circuitry whilst the transistor is in the on state. On-state power dissipation due to the device collector-emitter voltage drop ("V_{CE}") is partly determined by I_B (V_{CE} decreases as I_B increases), so reduced V_{CE} loss comes at the expense of increased I_B and increased driver dissipation.

The transistor drive current I_B is a source of power dissipation within the transistor itself. Current I_B flows through base-emitter junction whilst voltage V_{BE} appears across the base emitter. The power dissipation within the transistor base-emitter is therefore I_B x V_{BE}, and this figure can be significantly high in a bipolar power transistor.

The difference in drive currents for a BUK7506-55A MOSFET and a MJE3055T bipolar transistor is illustrated in Fig. 26. In this circuit, the devices were driven by an EL7104CN driver IC and a 10Ω series gate (base) resistor.

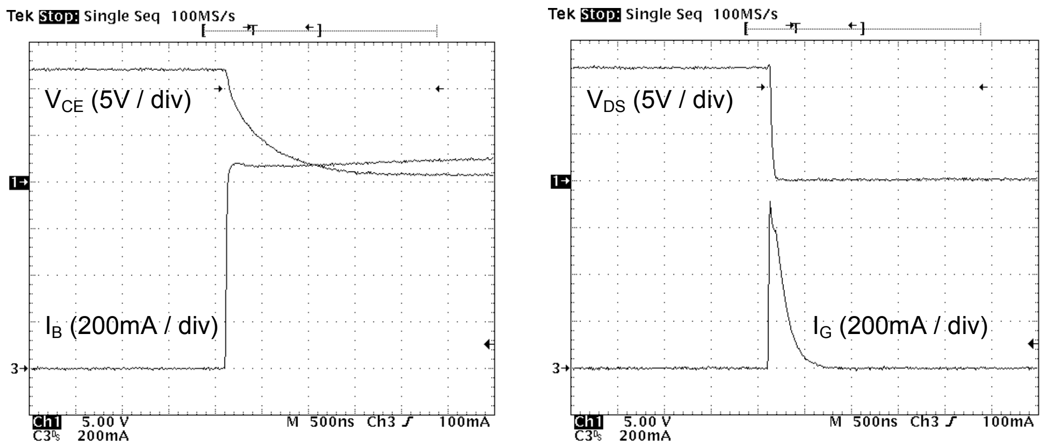

Fig. 26. Drive currents for an MJE3055T bipolar transistor (left) and BUK7506-55A MOSFET (right). Waveforms are shown at device turn-on.

Note the two important differences between the waveforms of Fig. 26;

The MOSFET gate current consists of a brief spike of current which rapidly decays to zero as the device turns on, whereas the bipolar transistor base current remains at a level of approximately 850mA, even after the device has turned on.

In comparison to the bipolar transistor, the MOSFET turns on considerably faster. This can be seen by comparing the rate at which the V_{DS} and V_{CE} waveforms fall to their on-state levels.

On-state losses.

On state loss in a MOSFET occurs due to current I_D flowing through the MOSFET drain-source whilst the device is conducting with on-resistance $R_{DS(on)}$. The loss ("P_D") is then simply;

$$P_{D(MOS)} = I_D^2 \,.\, R_{DS(on)}$$

Similarly, for a bipolar transistor, on state loss is found from multiplying the collector current ("I_C") by the collector emitter on-state voltage, V_{CE}. So;

$$P_{D(Bip)} = I_C \, x \, V_{CE}$$

Taking a typical operating current of 10A, and device V_{DS} or V_{CE} rating of 55V or 60V, even the best bipolar device available in TO220 has an on-state V_{CE} of around 1V, resulting in an on-state dissipation of 10W. In comparison, 55V TO220 MOSFETs are available in $R_{DS(on)}$ values as low as 6mΩ, allowing MOSFET on-state losses as low as 0.6W. A comparison of on-state losses is shown in Fig. 27.

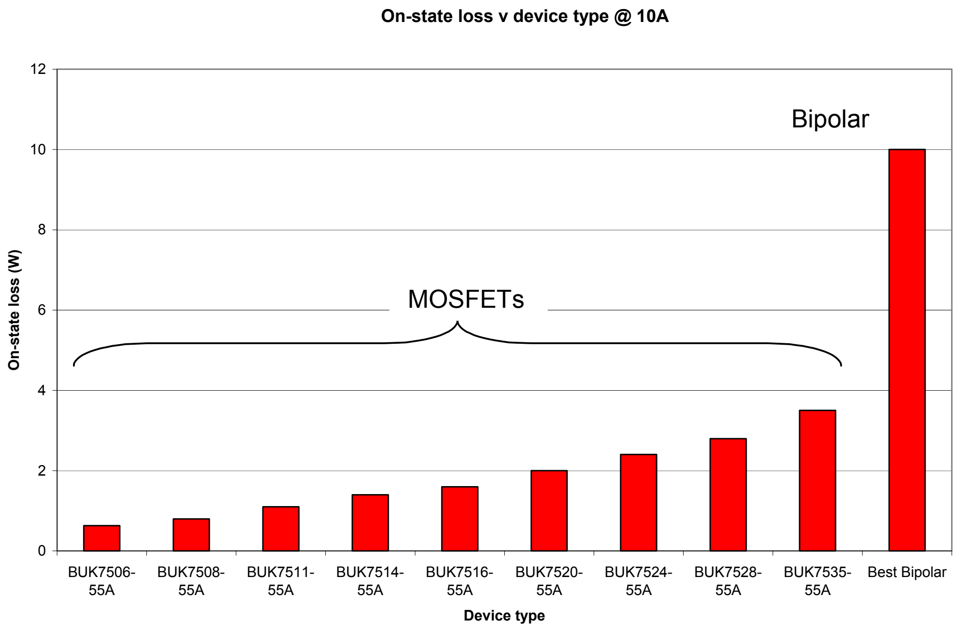

Fig. 27. Comparison of MOSFET and bipolar on-state losses.

<u>Paralleling of devices for increased power handling.</u>

If the application power handling requirements exceed that which can be achieved with a single device, then increased power handling capability may be realised by connecting two or more devices in parallel. Equal current sharing between parallel connected MOSFETs is easily achieved due to the positive temperature coefficient of $R_{DS(on)}$ inherent in the MOSFET structure. The positive temperature coefficient ensures that device $R_{DS(on)}$ increases with increasing temperature.

To illustrate this point, imagine that we have two of the same type of MOSFET connected in parallel, both sharing the same total load current. Due to manufacturing spreads, there will inevitably be small differences in $R_{DS(on)}$ between the two devices, so that in the initial state the device with lower $R_{DS(on)}$ will conduct more than 50% of the total load current. As this device is conducting more current, its power dissipation will be higher and hence it will experience a greater rise in junction temperature ("T_j"). As T_j rises, the device $R_{DS(on)}$ also rises and so the amount of current conducted by the device will tend to decrease – with the overall result that current sharing between the two devices will tend to automatically equalise without the need for any external current monitoring circuitry. Using this principle, the circuit designer may connect two or more MOSFETs in parallel and be assured that equal current sharing will occur, and the only additional circuitry that is required is an individual gate resistor for each MOSFET (see Fig. 28).

In comparison, equal current sharing in paralleled bipolar transistors is much harder to achieve. Bipolar transistors are certainly not “self equalising” – the bipolar collector-emitter drop (V_{CE}) exhibits a *negative* temperature coefficient and tends to *decrease* with increasing temperature. As a result, if we simply connect two bipolar devices in parallel, the device with the lower V_{CE} will initially conduct more current and hence dissipate more power. As more power is dissipated in the device, so its temperature increases and V_{CE} decreases further, allowing even more current to flow, and so on. The final result would be either very unequal current sharing (with one of the two devices running much hotter than the other) or, in extreme cases, a device which is forced into thermal runaway and eventual failure.

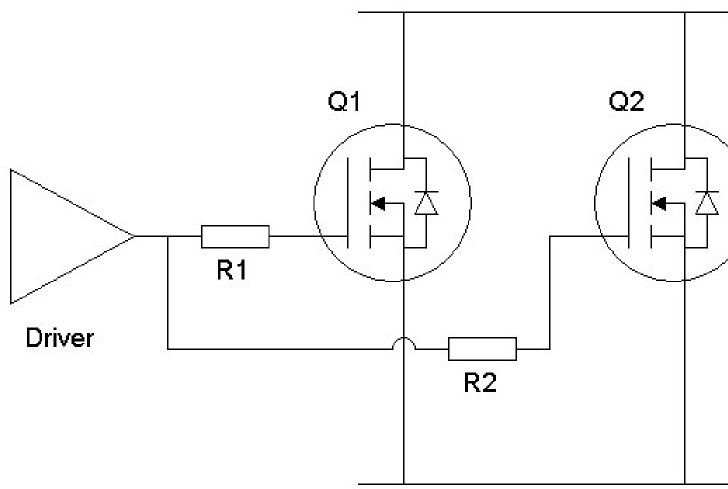

Fig. 28. Individual gate resistors R1, R2 used when driving parallel MOSFETs Q1 and Q2 from a common driver circuit.

<u>Utilisation of the MOSFET integral “body diode” in bridge circuits.</u>

Several of the common motor control topologies utilise MOSFET body diodes during the “freewheel” part of their conduction cycle. A typical example of this is shown in Fig. 29. In the example shown, Q2 (the “steering” MOSFET) is permanently on whilst Q1 and Q3 are turned on and off alternately. When Q1 is on, Q3 is off and vice versa. In order to ensure that Q1 and Q3 are never conducting at the same time, a gap (“dead time”) is left between Q1 turning off and Q3 turning on and between Q3 turning off and Q1 turning on. The gap is typically 1 or 2us in duration. During the dead time, Q1’s body diode is allowed to conduct the motor current. The provision of dead time in such a circuit is essential, and there is no easy way to design a bridge circuit which does not need dead time. Consequently, there is always a need to provide a current path from Q1 source to drain – even when Q1 is turned off (the same also applies to Q2 as the circuit is symmetrical). If bipolar transistors were used in the Q1 and Q2 positions then external diodes would need to be fitted across the collector/emitters of these devices as no equivalent of the MOSFET body diode exists within a bipolar transistor. Such external diodes are often referred to “anti-parallel” diodes.

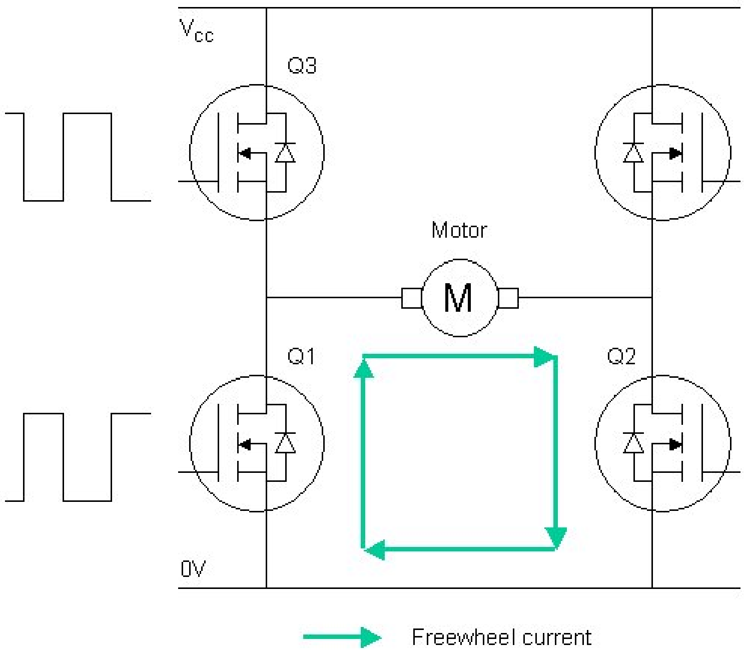

Fig. 29. MOSFET body diode conduction in a full bridge configuration.

<u>Summary.</u>

Characteristic	Bipolar transistor	MOSFET
Ease of drive	Base-emitter behaves like a diode	Drain-source behaves like a capacitor
	Base current must be supplied with current for the whole time the device is on	Gate source capacitance need only be fully charged to turn the device on
	Driver and base-emitter dissipations are high	Driver and gate-source dissipations are negligible
On-state losses	On-state losses are high as on-state V_{CE} cannot be reduced below ~1V at high currents.	On-state losses are much lower as extremely low $R_{DS(on)}$ values are possible.
Paralleling of devices	Difficult as parallel devices will not automatically adopt equal current sharing	Easy as devices will share current equally with the addition of almost no additional circuitry.
Anti-parallel diodes	External components must be fitted to fulfil this function.	“Built in” body diode may be used.

7. CHOICE OF POWER SEMICONDUCTORS IN MOTOR DRIVE CIRCUITS.

MOSFETs in the simple chopper, full bridge and three-phase bridge topologies.

MOSFET V_{DS} rating.

The MOSFET V_{DS} rating indicates the maximum drain-source voltage that the device is guaranteed to withstand when in the “off” state. In all the topologies considered, the MOSFET drain is clamped at a voltage close to V_{CC}, so the MOSFET V_{DS} rating need only be sufficient to deal with any voltage spikes appearing across the device at turn-off. The magnitude of the voltage spikes appearing at turn-off will depend on several factors, including device switching speed and the presence of stray inductances in the PCB layout. For a given device, faster switching will usually result in higher peak voltages. Speed of switching will itself depend on the peak current capability of the MOSFET drive circuit and the magnitude of the impedance in the gate path. This principle is demonstrated in Fig. 30. In this example, the left-hand trace was created with a high gate impedance (slow switching) and the right-hand trace was created with a low gate impedance (fast switching). Note the considerable difference in peak voltage – approximately 16V.

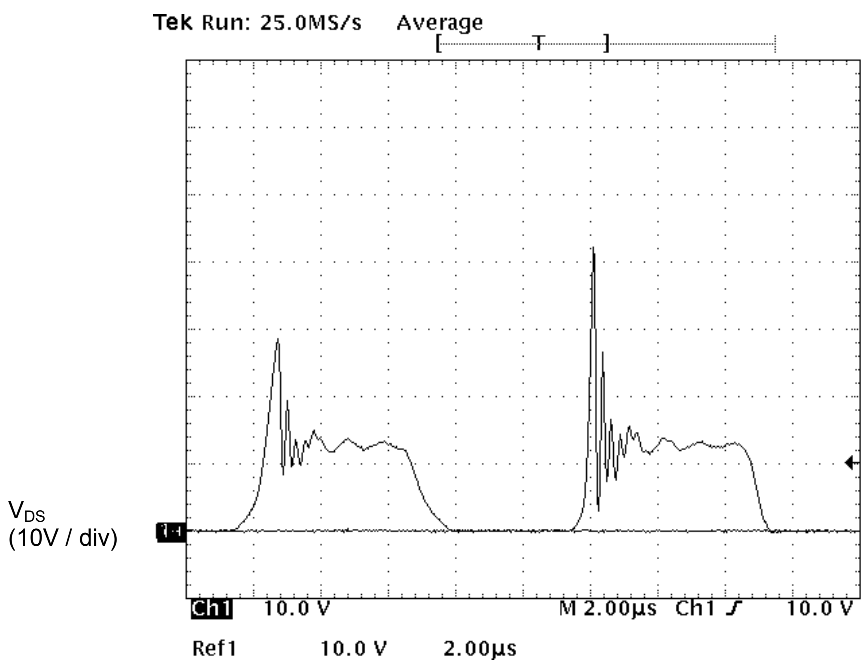

Fig. 30. MOSFET drain-source voltage at turn-off with high gate impedance (left) and low gate impedance (right)

It is also advisable to monitor the MOSFET drain-source voltage at the upper limit of the circuit’s acceptable supply voltage range, and also during soft-start and short-circuit conditions, in order to determine the highest possible voltage to which the device may be subjected under any combination of circumstances.

As a general rule, 40V, 55V or 60V rated devices are appropriate for circuits running from a 12V supply, whilst 60V, 75V or possibly 100V rated devices would be chosen for circuits running from a 24V supply.

MOSFET $R_{DS(on)}$ rating.

In motor drive topologies operating at frequencies of up to a few tens of kHz, the dominant MOSFET loss is on-state loss and hence $R_{DS(on)}$ is the dominant factor in determining loss in this component. Lower $R_{DS(on)}$ (higher I_D rated) devices will have lower power dissipation, but will be more expensive. The choice of device $R_{DS(on)}$ will largely depend on what measures are available to cool the MOSFETs. Typically, the chosen devices will be in TO220 packages and will be bolted to some form of heatsink, the function of the heatsink being to help channel the MOSFET power dissipation into the surrounding air. A larger heatsink will allow the use of higher $R_{DS(on)}$ MOSFETs. If however, there is no room for a large heatsink then lower $R_{DS(on)}$ devices may be needed in order to ensure that these devices run at acceptable temperatures. This may also be considered as an exercise on balancing MOSFET and heatsink costs. TO220, in order that the device may be bolted to a heatsink, though surface-mount devices are also available. See Tables 1 and 2.

Logic Level or Standard Level?

The terms “Logic Level” and “Standard Level” refer to the gate drive voltage required to drive a MOSFET into its fully turned-on state. A Logic Level device will have a data sheet $R_{DS(on)}$ specified at a gate drive voltage (V_{GS}) of 5V, indicating that a voltage of at least 5V must be applied in order for the MOSFET to be fully turned on. At a lower voltage than this, the device may not be fully turned on and will have an $R_{DS(on)}$ much higher than that specified in the data sheet. Similarly, a Standard Level device will have an $R_{DS(on)}$ specified at V_{GS} = 10V and, again, a lower applied V_{GS} will result in a higher $R_{DS(on)}$.

The maximum gate –source voltage which may be applied to a device ($V_{GS(max)}$) is also determined by whether it is Standard or Logic Level part. Generally, a Standard Level device has a $V_{GS(max)}$ rating of ±20V and a Logic Level device has a $V_{GS(max)}$ rating of ±10V. Operation beyond these limits will usually result in premature device failure.

In summary, therefore, the choice of Standard Level or Logic Level is determined by the level of gate drive available whilst ensuring that device $V_{GS(max)}$ ratings are not exceeded.

V_{DS}	$R_{DS(on)}$	Package		Standard / logic level
(V)	(milliohms)	TO220	D^2-Pak	(S / L)
40	2.8	PHP222NQ04LT	PHB222NQ04LT	L
	3.1	PHP225NQ04T	PHB225NQ04T	S
	4	PHP174NQ04LT	PHB174NQ04LT	L
	4.3	PHP176NQ04T	PHB176NQ04T	S
	5	PHP129NQ04LT	PHB129NQ04LT	L
	5.2		PHB143NQ04T	S
	7		PHB95NQ04LT	L
	8	PHP101NQ04T	PHB101NQ04T	S
55	3.7	PHP191NQ06LT	PHB191NQ06LT	L
	4	PHP193NQ06T	PHB193NQ06T	S
	5.4		PHB146NQ06LT	L
	6		PHB145NQ06T	S
	7	PHP110NQ06T	PHB110NQ06T	S
	7.1	PHP119NQ06T	PHB119NQ06T	S
	75	PHP20N06T	PHB20N06T	S
	75	PHP21N06LT	PHB21N06LT	L
60	12	PHP83N06T		S
	14	PHP73N06T	PHB73N06T	S
	22	PHP52N06T		S
	40	PHP32N06LT	PHB32N06LT	L
	150	PHP3055E		S

Table 1. 40V, 55V and 60V MOSFETs.

V_{DS}	$R_{DS(on)}$	Package		Standard / logic level
(V)	(milliohms)	TO220	D^2-Pak	(S / L)
75	5.6	PHP160NQ08T	PHB160NQ08T	S
	6.1	PHP153NQ08LT	PHB153NQ08LT	L
	9	PHP110NQ08LT	PHB110NQ08LT	L
	9	PHP110NQ08T	PHB110NQ08T	S
	50	PHP29N08T	PHB29N08T	S
100	8.8	PSMN009-100P	PSMN009-100B	S
	15	PSMN015-100P	PSMN015-100B	S
	25	PHP45NQ10T	PHB45NQ10T	S
	28	PHP47NQ10T	PHB47NQ10T	S
	40	PHP34NQ10T	PHB34NQ10T	S
	51	PHP27NQ10T	PHB27NQ10T	S
	70	PHP23NQ10T	PHB23NQ10T	S
	90	PHP18NQ10T	PHB18NQ10T	S

Table 2. 75V and 100V MOSFETs.

<u>Simple chopper circuit - diode.</u>

The only diode loss of concern is on-state loss (I_F x V_F) and hence forward voltage ("V_F") is the dominant factor in determining loss in this component. Lower V_F (higher I_F rated) devices will have lower power dissipation, but will be more expensive. As with the MOSFET, the diode V_{RRM} rating need only be sufficient to deal with any voltage spikes appearing across the device at turn-off. Note that V_F (and hence on-state loss) is generally higher for devices with higher V_{RRM} ratings, so in order to minimise dissipation it is a good idea to choose the lowest V_{RRM} rated device possible. In general, schottky diodes are appropriate for this application, with a 45V rated device being suitable for a chopper circuit running from a 12V supply, and a 60V or possibly 100V rated device would be chosen for a circuit running from a 24V supply. Choice of package is usually TO220, in order that the device may be bolted to a heatsink. Surface-mount devices are also available. See Table 3.

V_{RRM}	V_F	Package		Comments
(V)	(mV)	TO220	D^2-Pak	
45	570	PBYR745		
	570	PBYR1045	PBYR1045B	
	570	PBYR1545CT		Dual device - connect the two halves in parallel
	570	PBYR1645	PBYR1645	
	570	PBYR2045CT		Dual device - connect the two halves in parallel
	620	PBYR2545CT	PBYR2545CTB	Dual device - connect the two halves in parallel
60	700	PBYR1060		
100	700	PBYR20100CT	PBYR20100CTB	Dual device - connect the two halves in parallel

Table 3. 45V, 60V and 100V diodes.