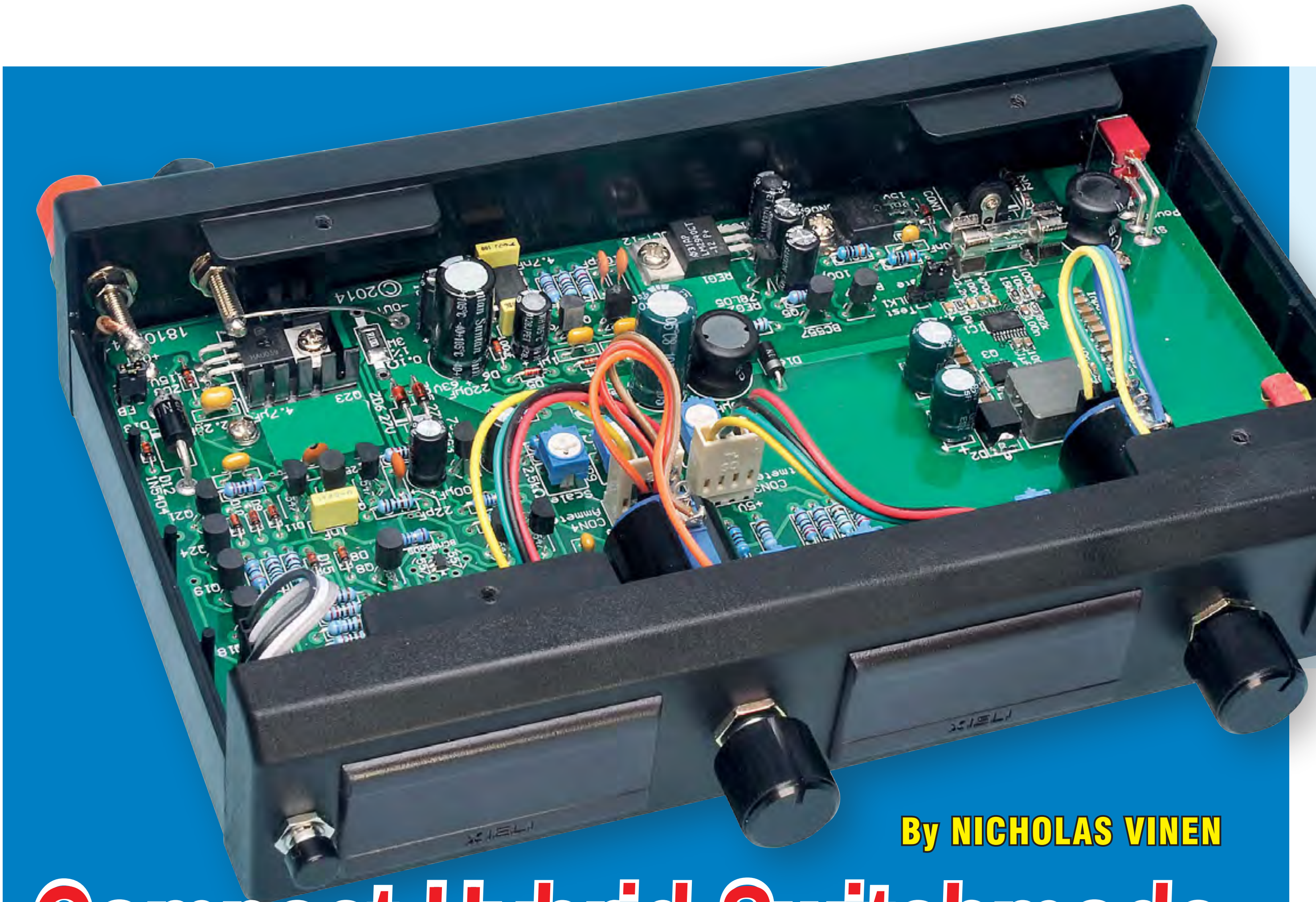


By NICHOLAS VINEN

Compact Hybrid Switchmode 100W Bench Supply – Part 1

Includes dual voltage and current metering

This very compact bench supply can deliver 0-40V at up to 5A with accurate and fast current limiting and has 3.5-digit 7-segment LED readouts for simultaneous voltage and current display. You can power it from any 12-24V DC supply such as a PC or laptop power supply or lead-acid/lithium battery. It uses a combination of switchmode and linear circuitry to obtain good regulation and low residual noise.

NORMALLY, YOU would expect any adjustable power supply capable of producing up to 40V and 5A to be a great deal larger than this little unit. In fact, it fits into a tiny half 1U rack plastic case. It measures just 209 × 43 × 122mm (W × H × D), not including the knobs and rear terminals. So how have we managed this feat of miniaturisation?

The first point is that it is really just an elaborate regulator and is meant to be powered by a laptop supply or similar. Second, the circuitry is housed on one double-sided plated-through PCB which employs some surface mount devices and MOSFETs selected for low-on resistance to produce very little heat dissipation inside the case.

We have combined the benefits of switchmode and linear regulator

Left: the unit is built into a compact half 1U rack plastic case measuring just 209 × 43 × 122mm (W × H × D), not including the knobs and rear terminals. It comes with the panel meters and load switch already fitted.

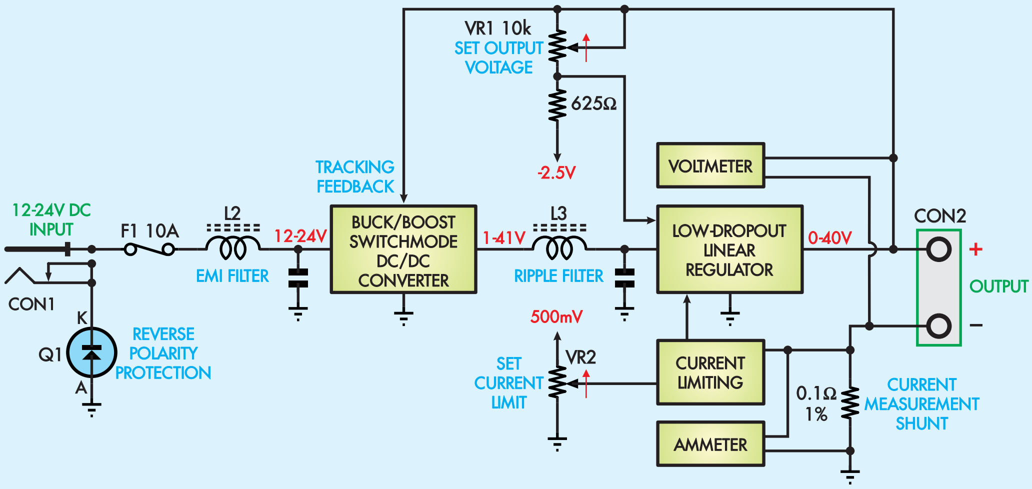


Fig.1: block diagram of the switchmode/linear bench supply. The output voltage is adjusted using VR1 which provides feedback to the low drop-out linear regulator section, which acts to maintain the feedback potential at 0V. VR2 sets the current limit to 0-5A, while the buck/boost switchmode section monitors the output voltage of the linear regulator and adjusts its output to provide about 0.7V 'headroom' at the regulator's input.

circuitry. Its output is adjustable over the range of 0-40V and unlike some designs, goes all the way down to 0V. Its current limit is adjustable from 0-5A and has fine resolution so that low currents can be accurately set. The dual LED panel meters constantly display the output voltage and current and the current limit can be displayed and set without having to short the outputs. It also has a front-panel load switch; this lets you set up the required voltage before switching on power to the load.

Being a hybrid design (switchmode *and* linear), it has much lower output noise (hash) than a pure switchmode bench supply and also doesn't need a large output capacitor bank that would then be dumped into the load in case of a short circuit. In fact, when the current limit kicks in, the output voltage drops very rapidly and the unit goes into current regulation mode. In other words, it can also be used as a near-ideal current source.

Since it runs off a low-voltage DC input, it can even be used away from 230VAC mains and powered from a car/truck/caravan battery or even a portable battery pack.

The dual voltage/current displays are really handy for a bench supply if you need to check that you have set the right output voltage and monitor the current draw while you are performing your tests. It's also quite handy to be able to see what the output voltage has dropped to, should current limiting be activated.

One feature missing from some cheap current-limited bench supplies is the ability to view the current limit setting without shorting the output leads. This is especially useful if you want to adjust the current limit while the load is powered since otherwise you really have no way to know what you've set it to while the load is drawing less current than the limit.

Buck/boost converter

The switchmode-based bench supplies we have published in the past have typically used a relatively large mains transformer to charge a capacitor bank to around 50V. They then used a step-down ('buck') switchmode converter to produce the required output voltage efficiently. This means the supply produces much less heat than a linear design of an equivalent power level.

In this case though, we wanted to fit the supply into this neat case from Altronics which comes with the panel meters and load switch already fitted. That ruled out using a large internal transformer. So we had the idea of powering it from a high-current DC supply which constructors may already possess, such as an old PC power supply or laptop charger. If you're like us, you have a few of these lying around, just waiting to be used for something grand.

PC supplies usually deliver the most current from their 12V output while laptop supplies normally give 15-24V with the most common being 17V. This means that our bench supply needs to be able to step the incoming supply voltage either up or down, depending on the required output voltage. And to be truly useful, it needs to do this efficiently at a reasonably high power level, matching that available from a typical laptop supply (60-100W).

To achieve this, we are using a 'buck/boost' switchmode converter. This is similar to the more common 'buck' type, but it can produce an output voltage that's higher, lower or the same as the input voltage. The particular chip we are using (the LM5118 from National Semiconductor, now Texas Instruments) operates in buck mode, boost mode or an intermediate buck/boost mode, depending on the ratio of the output to input voltages. We'll explain how this works in more detail below.

As a result, this supply can deliver plenty of current at lower voltages, up to about 15V, and then a lesser but still significant current up to the maximum 40V output (2.5A+, depending on the input DC supply voltage and power). Most bench supplies only go up to 30V and while this is sufficient for many tasks, we sometimes find it a bit limiting, hence the decision to go to 40V, even with a reduced current capability.

Performance

As mentioned briefly above, switchmode-based bench supplies always have some of the high-frequency

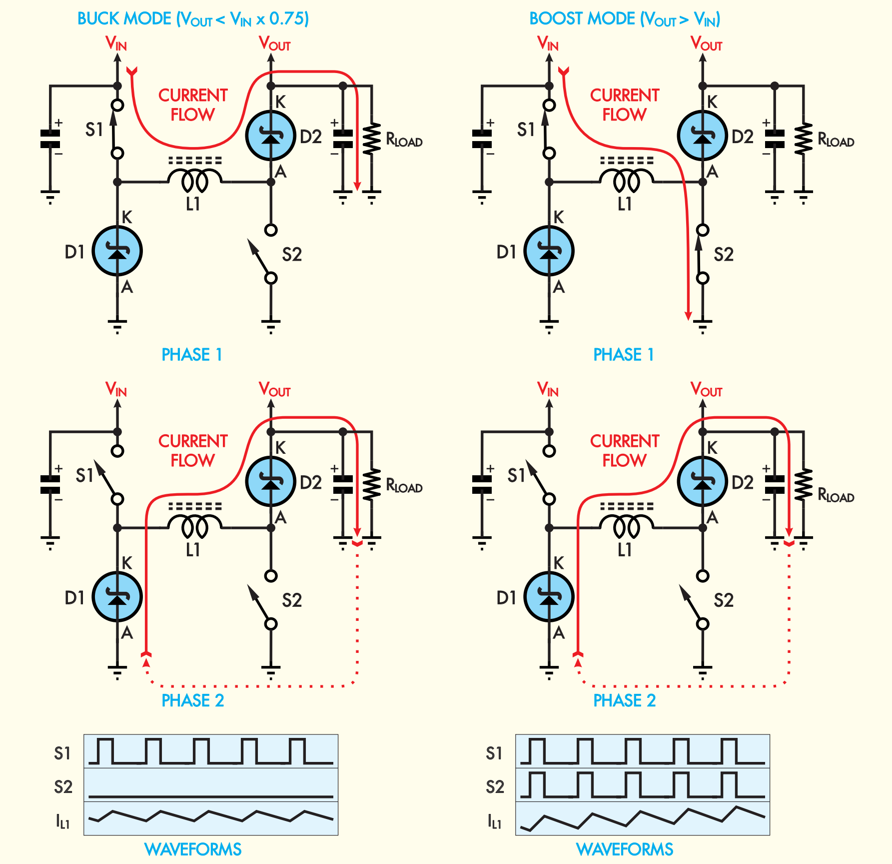


Fig.2: an illustration of how the switchmode converter works, in buck mode (diagrams at left) and boost mode (diagrams at right). The mode of operation is determined by whether S2 (actually a MOSFET) is switched with S1 or just left open (ie, off). In buck mode, as the duty cycle approaches 100%, the output voltage approaches the input voltage. In buck/boost mode, a 50% duty cycle gives an output voltage that's equal to the input, with higher duty cycles boosting the output voltage above the input, approximately doubling it at 75% duty cycle, quadrupling it at 87.5% and so on.

switching components present as 'hash' in the output, while an ideal bench supply should have pure DC with no noise or hash. In many cases, a switchmode-based bench supply will have an LC filter (or possibly a more complex filter involving a differential-mode choke) at the output to attenuate the noise, but this is only partially effective and also adversely affects output regulation.

Adding a linear regulator stage after the switchmode stage is a better proposition. This can offer greater noise and ripple rejection and depending on the dropout voltage it operates with, can also result in much better transient load response. In other words, it can cope better with sudden changes in load impedance/current draw, resulting in smaller variations in output voltage under these conditions.

Without getting into a lot of detail, the reason for this is that switchmode regulators tend to be quite heavily 'compensated', ie, their 'closed loop' bandwidth is purposefully reduced to a few kilohertz. This is necessary because the inductor and capacitors which are used to convert the switching output to a smooth voltage form a low-pass filter which leads to a delay between changes in the switching waveform and changes in the output voltage.

This delay is a form of phase shift; quite a large one in fact. And feedback systems with large phase shifts are unstable unless the gain is limited at higher frequencies. Linear regulators (depending on design) can have much smaller phase shifts, allowing more feedback bandwidth and thus much more rapid response to any changes in the output voltage due to the behaviour of the load.

Having a linear regulator at the output also means that we can implement the current limiting feature there. With the linear regulator's high bandwidth, that means it can provide a smooth current flow even in the face of rapidly varying load impedance and it also means there can be a very small output capacitance, so that there isn't much stored energy that will flow through the load before the current limit is effective. In this case, a 2.2µF output capacitor delivers a maximum of 1.75mJ of energy (with the output set to 40V) into a dead short.

So you can see that combining a switchmode and linear regulator gives us the best of both worlds. Well, it isn't quite perfect – some switching noise will still make it through the linear regulator, for example. However, it's a highly effective combination and a better compromise than either type of regulator by itself. Implementing it effectively is a bit tricky though, as we shall see.

Design concept

Fig.1 is the block diagram, which shows the overall design of the supply. The output voltage and current are controlled by the low-dropout linear regulator section, with VR1 adjusting the voltage and VR2 the current.

VR1 forms part of a voltage divider between the output and a –2.5V reference voltage. If the feedback voltage is above 0V, the regulator reduces its output, while if the feedback is below 0V, the output voltage is increased. The values selected give the unit a range of 0-40V.

The LDO regulator needs an input voltage that's slightly higher than its output voltage for proper regulation (at least 0.1V but ideally a bit more). As a result, the switchmode converter monitors this output voltage and attempts to maintain its own output at a slightly higher voltage. The 'headroom' is set at around 0.7V, so the output of the switchmode regulator will go slightly above 40V and normally never drops to zero.

An LC filter between the two regulators reduces high-frequency ripple fed to the linear regulator, as its input supply/ripple rejection is best at lower frequencies. There is a similar filter at the input of the switchmode regulator to stop too much noise coupling back to the input and possibly radiating EMI from the input wiring.

A 10A fuse protects the circuit against serious faults, however if the switchmode section is working normally, its cycle-by-cycle current limiting will mean that the fuse should never blow. Q1 provides input reverse polarity protection; although it operates as a diode, it is actually a MOSFET to avoid reducing the supply voltage too much and wasting a lot of power, as a standard diode would.

The voltmeter is wired across the output terminals while the ammeter displays the voltage across the shunt. Note that the voltage across the shunt is effectively subtracted from the output voltage, but the way the feedback network is connected automatically compensates for this (as explained later).

We have used 10-turn potentiometers for voltage and current adjustment as this makes it easier to set these values accurately; we recommend constructors do the same – however, there is nothing stopping you from using the cheaper 270° rotation pots should you wish.

Buck/boost operation

Most of the switchmode regulators we have published in the past have been one of three types, either 'buck', 'boost' or based around a transformer. The buck and boost types are the simplest; however, the former can only reduce the input voltage while the latter can only produce an output greater than the input. Hence the use of buck/boost, which gives a much wider range of output voltages.

The LM5118 IC operates in buck mode when the output voltage is less than three quaters the input voltage, and boost mode when the output voltage exceeds the input voltage. Between these, it operates in an intermediate mode which is partly buck and partly boost, ie, buck/boost.

Fig.2 shows the difference between the two main modes. At left are the two states used for buck mode. When S1 is on, current can flow from the input straight to the output, via inductor L1 and Schottky diode D2. During this time, L1's magnetic field charges up and the current flow smoothly ramps upwards, at a rate determined by the voltage across L1 and its inductance.

When S1 is switched off (below), L1's magnetic field continues to drive current through the load via D2, but this current can no longer come from V_{IN}, so it must flow through Schottky diode D1 from ground. The dotted line shows how current recirculates – the only source of energy during S1's off time is L1's magnetic field. As such, the current flow smoothly drops, again at a rate limited by the voltage across L1 (now roughly equal to the output voltage) and its inductance.

This cycle repeats and the ratio of S1's on-time to off-time, in combination with the load impedance, determines the ratio of the output voltage to the input voltage, but this is always less than one. Some example waveforms are shown below these diagrams, for a steady state (ie, constant load and output voltage).

Compare these diagrams to those at right, which show operation in boost mode. The difference is that now S2 switches on simultaneously with S1. This increases the voltage across L1 to be the full input supply voltage and this does not drop over time, so L1's magnetic field charges up much faster. Thus, more current is delivered during the off-time (below) and hence the output voltage is higher for the same duty cycle as buck mode.

It stands to reason then that the ratio of the output voltage to the input voltage can be greater than one and in fact, it is inversely proportional to the duty cycle. Thus, the maximum output voltage is limited mainly by the maximum duty cycle, which for the LM5118 is related to the operation frequency (as there is a fixed minimum off-time). In our circuit, maximum duty cycle is about 85%, giving a maximum boost ratio of about 4:1, certainly sufficient to get an output of over 40V from an input of 12V.

When in the intermediate mode mentioned above, the only difference is that S2 switches off before S1, thus giving three phases for each cycle, equivalent to phase 1 for boost, followed by phase 1 for buck and then phase 2 (same in either mode). Thus, the boost ratio is not as high as in pure boost mode. This intermediate mode means there is no discontinuity in the converter's operation or output voltage.

Circuit description

Now let's turn to the full circuit. Fig.3 shows the main section. At its heart is the buck/boost switchmode converter, controlled by IC1 (LM5118). First, let's look at the 'output' side of IC1, ie, pins 12-20. These drive the MOSFETs which do the actual switching.

Pin 19 is the high-side driver output which connects to the gate of Q2 (S1 in Fig.2). Pin 20 is connected to the source of this MOSFET, which is the 'floating' node that switches between ground and the incoming supply rail. This pin is used as the ground return for the discharge current from the MOSFET gate and as a negative reference when driving it high, charging the gate to this voltage plus about 7V.

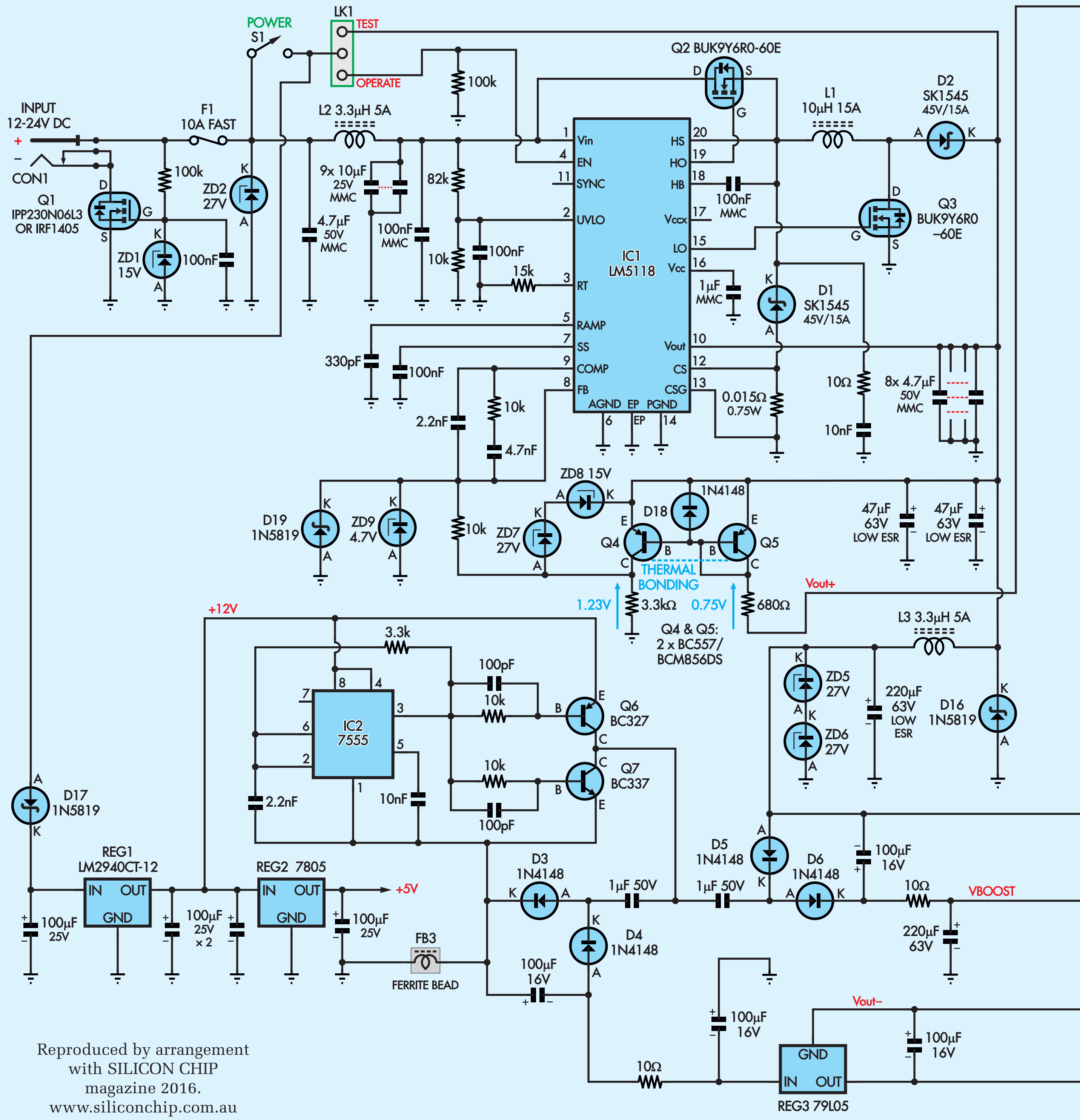
Thus if the input voltage is say 14V, the gate of Q2 must be driven to 21V. To generate this higher voltage, IC1 has an internal charge pump and it uses the 100nF capacitor between pins 18 and 20 to accomplish this. This capacitor is charged to 7V from the input supply when pin 20 is low and current then flows back from it into the MOSFET gate when pin 20 goes high, boosting the pin 18 voltage to the required level. This arrangement is known as a 'floating high-side driver'.

The low-side (boost) MOSFET, Q3, is driven from pin 15. This does not require a charge pump as its source terminal is connected directly to ground and thus the gate only needs to reach about 5V for full conduction.

Q2 and Q3 are logic-level MOSFETs which are switched fully on with a gate-source voltage of 5V. IC1 has an internal 7V regulator with a 1µF output filter capacitor connected from pin 16 to ground and this determines the maximum gate-source voltage fed to the two MOSFET gates. An external supply can be connected to pin 17 (V_{CCX}) but the internal regulator can supply enough current to operate the MOSFETs at 350kHz without excessive dissipation (a maximum of about 650mW).

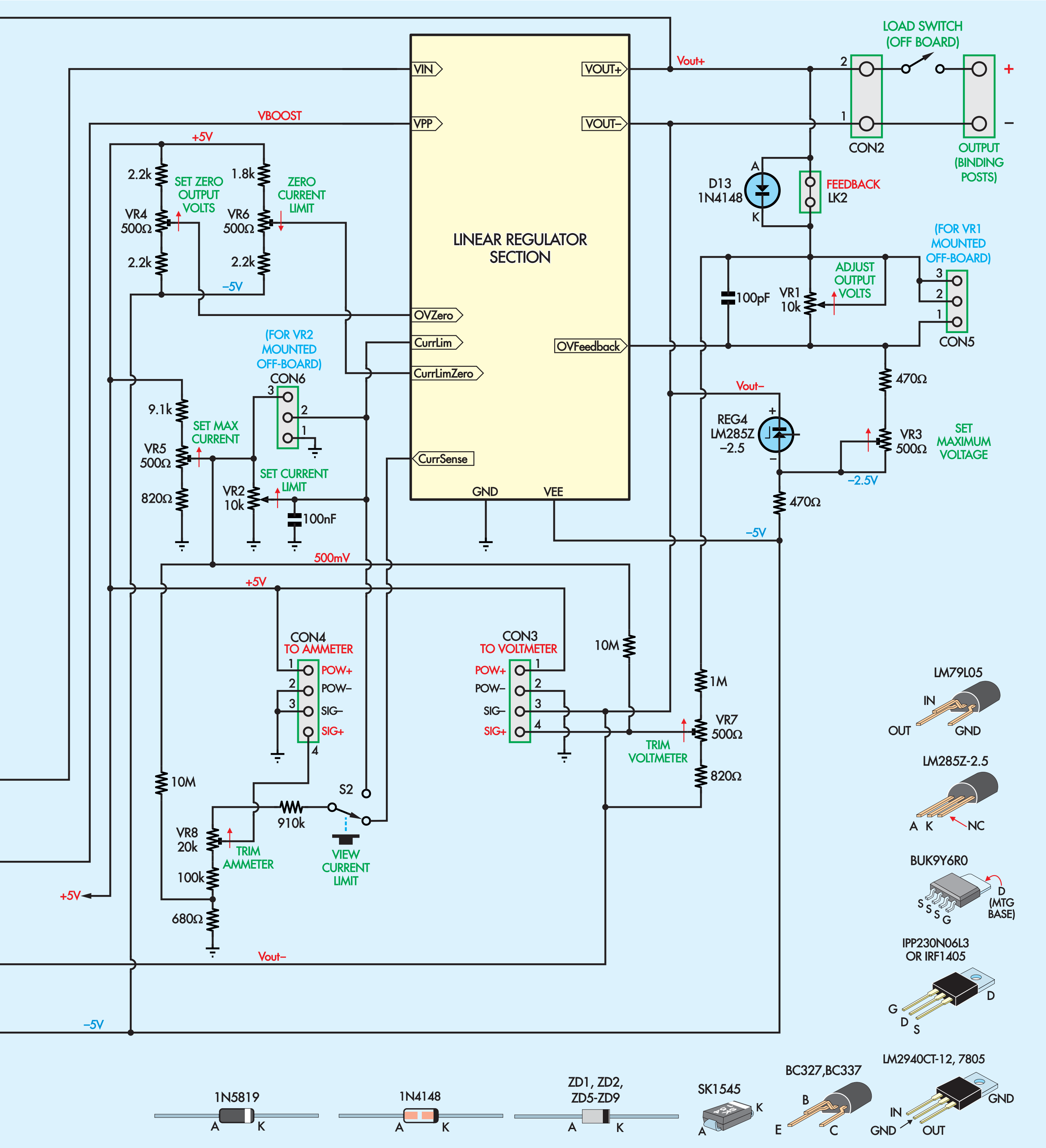
Ground return for the low-side MOSFET driver is pin 14 (PGND) while pins 12 and 13 are used to sense the voltage across a 15mΩ shunt connected in series with the buck recirculating diode, D1. This sets the peak current limit to 125mV / 15mΩ = 8.3A in buck mode and 250mV / 15mΩ = 16.6A in boost mode (close to the inductor's saturation current). The inductor current is sampled just after the MOSFET(s) switch off, when it is at its peak, just after D1 becomes forward biased.

Note that the switchmode arrangement is based largely on the sample circuit in the LM5118's data sheet, which provides a design with similar requirements to ours. We require a maximum boost of 40V / 12V = 3.3



SWITCHMODE-LINEAR HYBRID BENCH POWER SUPPLY MAIN CIRCUIT DIAGRAM

Fig.3: the main section of the bench supply circuit. IC1 is the buck/boost controller and this drives MOSFETs Q2 and Q3, which form the switchmode converter in conjunction with inductor L1 and Schottky diodes D1 and D2. The output then goes to the linear regulator (shown in detail next month) and then to the output terrminals. The output current is checked using a 0.1Ω shunt resistor which is monitored both by the ammeter panel and the current limit circuitry. REG1-REG3 and IC2 generate two extra supply rails for the linear regulator circuitry, one about 10V above the main supply rail (V_{BOOST}) and a –5V rail, using a charge pump driven by Q6 and Q7. VR1 and VR2 set the output voltage and current limit while VR3-VR8 zero and adjust the meter displays, output voltage range and current limit range.



times with an input current of around 8A (100W / 12V), while their design is for a maximum boost of 12V / 5V = 2.4 times with an input current of around 8A (36W / 5V).

We are using 45V 15A Schottky diodes for D1 and D2 since these are the components in the switchmode section which dissipate the most power. They are relatively compact devices, which is important because the PCB layout of this section is critical.

The BUK9Y6R0-60E MOSFETs were chosen due to their low gate charge (minimising switching losses), low on-resistance (minimising I^2R power dissipation) and ease of soldering. While the LM5118 data sheet says a snubber is not required, we have fitted one – consisting of a 10Ω resistor and 10nF capacitor in series, across D1 –

The PCB assembly is designed to mate with integral pillars inside the half 1U rack plastic case. Part 2 next month has the full assembly details. Note that the final board will differ slightly from this prototype.

to reduce voltage spikes and thus EMI during MOSFET switching.

A parallel array of eight 10µF 25V multi-layer ceramic capacitors is used for input supply filtering. This arrangement has a very low ESR and is also relatively cheap. A similar arrangement of nine 4.7µF 50V ceramic capacitors is used for the output, where low ESR is important to minimise ripple. These form a low-pass filter in combination with inductor L1. A couple of 47µF low-ESR electrolytics are paralleled for 'bulk capacitance', which helps switchmode feedback loop stability.

The output of the switchmode regulator passes through another LC low-pass filter, consisting of a 3.3µH bobbin inductor (chosen for its low losses and low price) followed by a 220µF low-ESR electrolytic capacitor. This attenuates the output ripple of the switchmode regulator and this voltage then feeds into the linear regulator (to be described next month).

Feedback and control circuitry

Pin 1 of IC1 is the supply input for its high-side and this is decoupled with a 100nF ceramic capacitor close to that pin of the IC. The resistive divider connected to pin 2 sets the under-voltage lock-out threshold at 11.3V [1.23V × (82kΩ + 10kΩ) / 10kΩ]; while the IC can run from at little as 5V, we want to avoid excessive input current draw at low supply voltages.

The 100nF capacitor at pin 2 sets up the 'hiccup' over-current protection; if a prolonged over-current condition is detected, pin 2 is pulled to ground and this capacitor takes some time to charge to the 11.3V threshold, preventing excessive current draw in case of a prolonged short or other overload.

A resistor from pin 3 to ground sets the operating frequency of the switchmode regulator, with 15kΩ giving operation at around 350kHz. Higher frequencies mean less RMS ripple voltage at the output but in exchange for that, switching losses are higher (due to more frequent transitions at the output). Also, inductors tend to be lossier at higher frequencies.

Pin 4 is the enable input and must be pulled high for the regulator to operate. This is connected to power switch S1 via link LK1, with a 100kΩ pull-down resistor. Thus, when power switch S1 is off, voltage is still applied to the input of IC1 but it is disabled so the output is at 0V. This avoids S1 having to switch a high current. LK1 is used to disable and bypass the switchmode regulator for testing the rest of the circuit independently.

The capacitor connected from pin 5 to ground sets the time constant for the 'emulated ramp compensation'. This value is chosen to match the time constant for the rate of change of current flow through the output inductor (L1) and allows the IC to perform ramp compensation without needing to measure the current through L2. Ramp compensation is required for feedback loop stability at higher duty cycles.

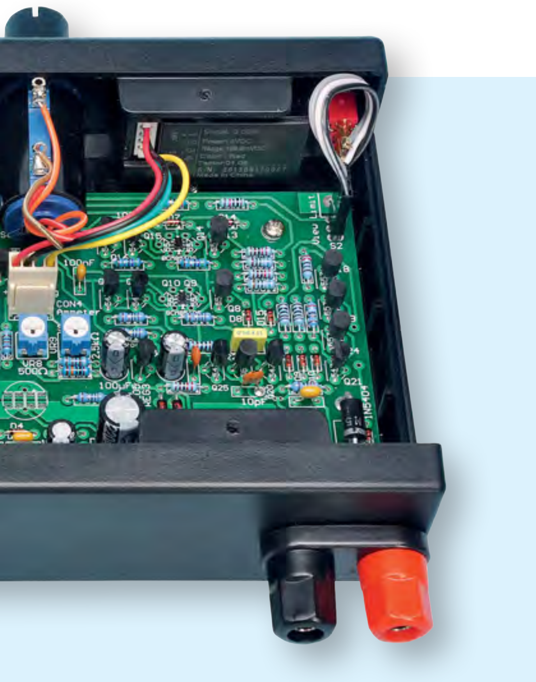
In brief, ramp compensation involves feeding back a voltage related to the output duty cycle to the input of the error amplifier, in order to avoid the duty cycle oscillating either side of the required stable value without settling down.

Pin 6 is the analogue ground pin, ie, the ground return for the components connected to pins 1-9. Pin 7 is the soft start pin and the connected capacitor is charged at power-on, with the output duty cycle being limited until it is fully charged, to prevent the IC from drawing very high input currents while the output capacitor bank is charged.

Pin 8 is for voltage feedback and is the input to the error amplifier. When pin 8 is below 1.23V, the output duty cycle increases and if it is above 1.23V, the output duty cycle is reduced. Normally this is connected to the output of the regulator via a resistive divider, to set a fixed output voltage, or with one resistor replaced by a rheostat or potentiometer to give an adjustable output voltage.

But in this case, we don't want to set the output voltage of the switchmode regulator directly. Instead, we want its output to be slightly higher than that of the linear regulator, so that it has 'headroom' to operate and deal with load transients but without dissipating much power in the linear pass element.

This is achieved using PNP transistors Q4 and Q5. These form a current mirror and their emitters are tied to the output of the switchmode regulator. The difference between this and the output of the linear regulator causes a current to flow through the 680Ω resistor at Q5's collector. When



the difference is 0.75V, that current is (0.75V – 0.5V) / 680Ω = 0.37mA.

Being a current mirror, this also flows through the 3.3kΩ resistor at the collector of Q4 which gives a voltage of 3.3kΩ × 0.37mA = 1.23V, which is IC1's internal reference voltage level. Thus, its negative feedback will maintain the switchmode output voltage about 0.75V above the linear regulator's output voltage. Being effectively a common-base amplifier, this arrangement has very little phase shift and thus does not affect IC1's feedback loop stability.

Note, though, that the output of the switchmode regulator can't drop below 1.23V as this is the minimum, so dissipation in the linear regulator will be a little higher when its output voltage is below 0.5V or so.

ZD7 and ZD8 prevent the output of the switchmode regulator from exceeding 45V in case of a feedback fault. The 10kΩ current-limiting resistor in combination with zener diode ZD9 and Schottky diode D19 protect the feedback input (pin 8 of IC1) from going outside the range of –0.3 to 5V, which would risk destroying the IC.

Diode D18 protects Q4 and Q5 from damage due to base-emitter junction reverse breakdown.

Linear regulator supply rails

The linear regulator requires a positive supply rail (V_{PP}) that is at least 7V above the output in order to switch its internal MOSFET on fully to supply a high load current. It also requires a well-filtered negative rail (V_{EE}) several volts below ground so that it can turn that MOSFET fully off when required (allowing for the voltage drop of several internal driving transistors).

These are supplied from a charge pump shown in Fig.3. This is based on REG1, a 12V low-dropout regulator and IC2, a 7555 CMOS timer IC. IC2 is set up to provide a 12V square wave at 100kHz at its output pin 3. This drives a complementary pair of bipolar transistors, Q6 and Q7, which form an inverting buffer. The 100pF capacitors across their base current-limiting resistors speed up switch-off to prevent cross-conduction.

The buffered output drives two 1µF capacitors, one of which is charged to the switchmode output voltage by D5 and the other which is clamped to ground by D3. When the buffered output of Q6/Q7 goes positive, D6 becomes forward biased and the connected 100µF capacitor is charged to roughly 10V above the switchmode supply rail. Similarly, when the collectors of Q6/Q7 go negative (to ground), D4 becomes forward biased, charging the connected 100µF capacitor to about –10V.

These two new voltage rails are then filtered using RC filters (10Ω/220µF and 10Ω/100µF respectively) to remove most of the 100kHz component, forming relatively smooth DC supply rails. The –10V rail is then regulated by REG3 to a stable and clean –5V for the linear regulator's V_{EE} rail. This regulation is necessary for two reasons: (1) any noise or ripple on this line will affect the regulator's output; and (2) this is also used as the reference for setting the output voltage.

Adjustments and trimming

As stated earlier, the linear regulator acts to keep the feedback voltage at around 0V. This is determined by the output voltage in combination with the position of 10kΩ potentiometer VR1 (Fig.3). This can be a 10-turn potentiometer to give finer output adjustment. It acts as a voltage divider in combination with trimpot VR3 and the 470Ω resistor connecting them.

VR3 is connected to a –2.5V rail, derived from the –5V rail by voltage reference REG4. The circuit will operate without REG4 – however, it will be subject to output voltage variations due to thermal drift in –5V regulator REG3; REG4 has much better thermal stability and is not dissipating anywhere near as much power either (typically <1mW compared to ~100mW for REG2).

When VR1 is at minimum resistance, the output rail is effectively connected directly to the feedback point and so the output voltage is at 0V. As VR1 is turned clockwise and its resistance increases, the output voltage must increase in order to keep the feedback voltage at 0V. For example, when the resistance of VR1 is around 625Ω, matching that of VR3 plus the 470Ω series resistor, the output is at around 2.5V.

VR3 is used to trim out variations in the other components, giving 40V at the output with VR1 fully clockwise.

The output voltage is fed to VR1 via link LK2 in parallel with diode D13. This is intended to allow for some wiring voltage drop compensation to be used due to the need to run a wire from the output terminal to the off-board load switch, in which case LK2 is removed and a wire is run from the supply side of the load switch to the lower terminal of LK2. In case this connection fails, D13 limits the output voltage from rising more than 0.6V.

The current limit is set using VR2, another 10kΩ potentiometer which can also be a 10-turn type. Its wiper voltage is filtered with a 100nF capacitor and fed into the linear regulator where it is compared with the voltage across a 100mΩ shunt. There is 500mV across VR2, derived from the +5V rail by trimpot VR5. This +5V comes from linear regulator REG2, which supplies several reference voltages but also the power for the two panel meters.

The voltmeter reads the voltage across the output terminals, but this is a 200mV full-scale meter so the output voltage is divided down by 1MΩ and 1kΩ resistors plus 500Ω trimpot VR7 for fine tuning. The panel meter's input impedance is 100MΩ so we are using relatively high value resistors here.

Similarly, to get a reading of up to 5A on the ammeter panel, we need a 0-50mV signal and so the 0-500mV from the shunt (at CurrSense) is divided down by a factor of 10 by a 910kΩ resistor, 100kΩ+1kΩ resistors and 20kΩ trimpot VR8. S2 allows the voltage feeding this divider to be switched from the current feedback to the current-limit setting, so that the limit can be viewed without having to short the output terminals.

The 10MΩ and 2.2MΩ resistors provide a small bias current to the two panel meters so that they do not give a negative reading when the output voltage is 0V or no current is being drawn.

The two remaining trimpots, VR4 and VR6, are used to trim out any offset error in the voltage feedback and current-limiting circuitry respectively. These inputs have a low impedance to ground so the adjustment ranges span just a few millivolts either side of 0V.

Remaining circuitry

The circuit is protected from a reversed input supply polarity by MOSFET Q1. When the supply is connected the right way, Q1's gate is pulled positive by the 100kΩ resistor and clamped at a safe level by the 15V zener diode. This switches it on and allows ground current to flow from the circuit back to the supply.

If connected backwards, the gate is pulled negative and so Q1 remains off. Its body diode is also reverse-biased and thus very little current will flow. The 100nF capacitor from its gate to ground slows its turn-on to avoid large current spikes charging the input capacitor bank when power is first supplied; IC1 has a soft-start feature, so it's just this input bank that can draw a high current initially.

A 10A fuse protects the circuit against serious faults, and 27V zener diode ZD2 conducts if the input supply voltage becomes too high. If that excessive voltage is maintained for very long, it will blow the fuse. The clamping voltage is above the 25V rating of the input capacitor bank, but they are unlikely to fail due to a brief over-voltage of just a few volts and we don't want ZD2 to conduct any significant current with the supply below 25V.

A 4.7µF capacitor and 3.3µH inductor (L2) prevent much switching noise from passing back through the input leads, which could lead to electromagnetic interference being radiated from them. Power switch S1 enables the switchmode regulator and at the same time, applies power to the rest of the circuit.

When LK1 is moved to the 'Test' position, the linear regulator remains off and power can bypass it from S1 straight to the output. This is so that the constructor can check the linear regulator and other circuitry is working before activating the switchmode portion; otherwise troubleshooting could be very difficult.

Finally, there is a Schottky clamp diode (D16) at the output of the switchmode regulator so that its output can not be pulled very far below ground by the linear regulator at start-up. There is also a clamp consisting of two 27V zeners (ZD5 and ZD6) in series after filter inductor L3, so that if the switchmode regulator feedback fails (including the ZD7/ZD8 voltage clamp), its output will not go high enough to damage the 63V filter capacitors or any part of the linear regulator circuitry.

LDO operation and construction

That's all we have room for this month. Next month, we'll describe the linear regulator section and begin the construction.