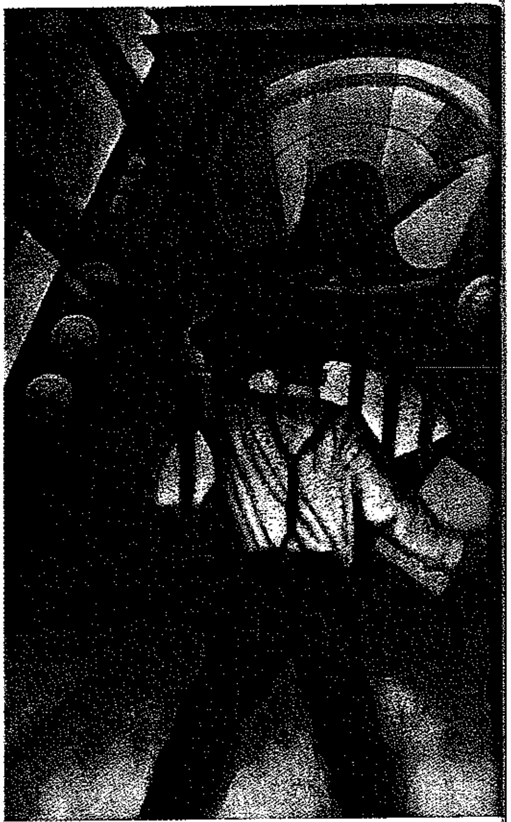
Designer's power supply

Ian Hickman – who frequently designs circuits – shares his experience of power supplies and presents a solution combining versatility with features necessary for powering up a prototype without risking damage.



Of the various self-designed power supply units gracing my workbench, all but one provide only a single-rail. The exception is a dual 15V, 1A unit, with the facility for use as tracking ±15V supplies, as a 30V 1A supply or a 15V 2A supply. Perhaps because of this versatility, it is the one that I use most often, despite the fact that the current limit on each section is fixed at 1A.

Having experienced the inadequacies of my existing supplies, it seemed a good idea to take a fresh approach when designing a new one.

Useful features

An important feature is adjustable current limit. Additionally, in the interests of flexibility, I wanted my design to be easily modified to accommodate different output voltages and/or currents.

Since much of my work involves low-level analogue signals, in the interests of low noise, the design would be a linear regulator, with the inefficiency that this admittedly involves.

As the design would spend most of its working life powering circuitry under development, I placed emphasis on good performance in the constant voltage, or cv, mode, making very low hum ripple – even at full load – a priority. Performance in constant-current, or cc, mode is less important. Constant-current mode was intended primarily as a safety feature, to protect both the supply and the circuit being tested under fault conditions. Dual 15V supplies were envisaged, with provision for independent operation, operation in series, and operation with the voltage of one unit – the slave – automatically set to the same value as the other, acting as master. In this mode, the two units may be paralleled to provide double the current available from each separately, or connected in series to provide tracking positive and negative rails.

I adopted a fairly standard approach, Fig. 1, with a constant-voltage loop controlled by IC_1 and a constant-current loop by IC_2. With the wiper of R_v set to ground, i.e. fully clockwise, output voltage is determined by the ratio of R_f and R_i, and the voltage at the non-inverting input of IC_1. On the other hand, with the wiper of R_v set fully anticlockwise, if R_i/R_f equals R_a/R_b the output voltage will be zero.

In constant voltage mode, the constant-current loop is inactive, since the volt drop across the current sense resistor R_c is small compared with the voltage at the non-inverting input of IC_2.

Of course, Fig. 1 is purely diagramatic; in order for it to work, either the op-amps must have n-p-n open-collector outputs, or the output of each must be connected to the base of the pass transistor via a diode. Furthermore, there must be a dummy load across the stabilised output, to provide a pull-down for the emitter of the pass transistor at low output voltages. But apart from that, the scheme is plausible.

Specifications of the power supply

These specifications are for the basic 15V, 1A supply, but the design is versatile and is easily modified for other voltages and currents.

Output voltage	15V max. nominal
Continuously adjustable	0V to max output
Noise, hum and ripple	<100μVrms
Output current	1A max. nominal
Current limit continuously adjustable	From max. to 50μA
Noise, hum and ripple in constant current	<8mV peak-to-peak
Regulation	
Output resistance – not in current limit	50mΩ
Peak deviation	700mV*
Recovery time	10μs*
* for step load change 50% to 100% of rated current.	
Stabilisation	
Output voltage variation	1mV for ±10% mains voltage change

For the 15V 1A version, each mains transformer secondary, combined with the bridge rectifier and 2200μF reservoir capacitor, should be capable of delivering 21V dc at 1.3A continuously.

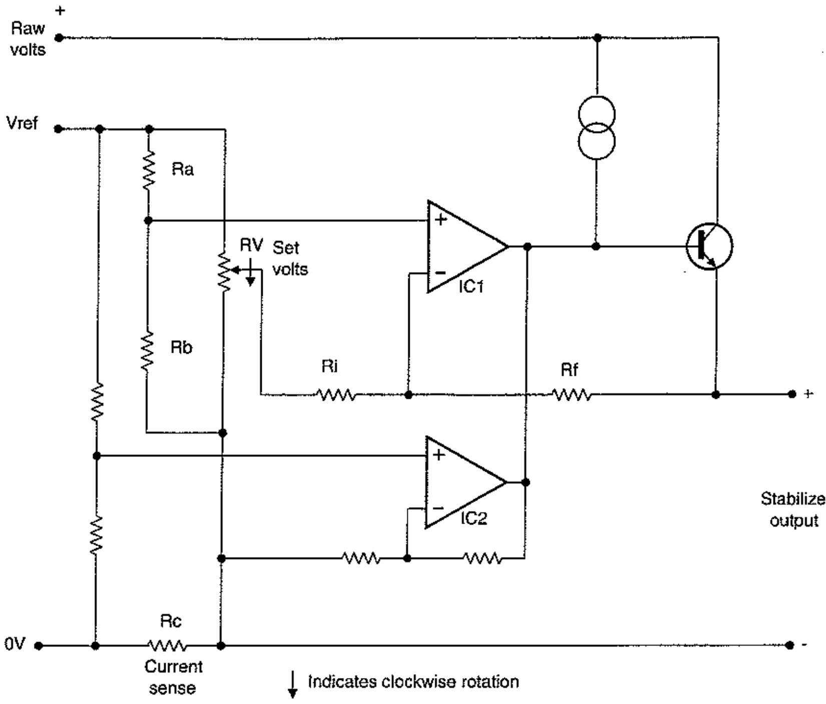


Fig. 1. Simplified circuit of a laboratory bench power supply.

When it comes to the detailed design, practical difficulties emerge. Op-amps with open-collector outputs are not generally available. Although comparators fill the bill in this respect, they are notoriously unstable when operating in a linear regime.

Another problem with the Fig. 1 scheme is that the op-amps must be able to pull the base of the pass transistor right down to the negative stabilised output terminal while sinking the current from the constant current generator. But op-amps capable of this are limited as to the maximum supply voltage they can stand.

In the event, the ICs in Fig. 1 were realised with discrete devices. Using discretes provides you with much greater design flexibility.

My choices

My chosen design was based on Fig. 1, but with a number of variations. For instance, n-p-n current mirrors, such as the Texas Instruments *TL0xx* range, are readily available, but p-n-p mirrors are not.

In principle, you could use devices in a pack of matched p-n-p transistors from the *RCA CA3xxx* range, but the solution adopted here was to use a resistor supplying current from an auxiliary supply of voltage higher than the positive raw supply. The final circuit is shown in **Fig. 2**. A mains transformer from stock was used, providing a 21V raw supply. Allowing for about 2.5V peak-to-peak ripple across the reservoir capacitor C_3 at 1A full load, this transformer allowed a generous margin of V_{ce} for the pass transistor – even at –10% mains voltage.

The raw positive supply uses a bridge rectifier circuit as this makes the best use of the transformer's secondary copper. The modest size reservoir capacitor allows appreciable ripple voltage, resulting in lower copper losses due to a longer conduction angle than would apply with a larger reservoir. An additional half-wave doubler circuit provides the auxiliary supply.

Reference voltage is provided by an op-amp and zener circuit. This is a convenient arrangement using readily available devices, but you

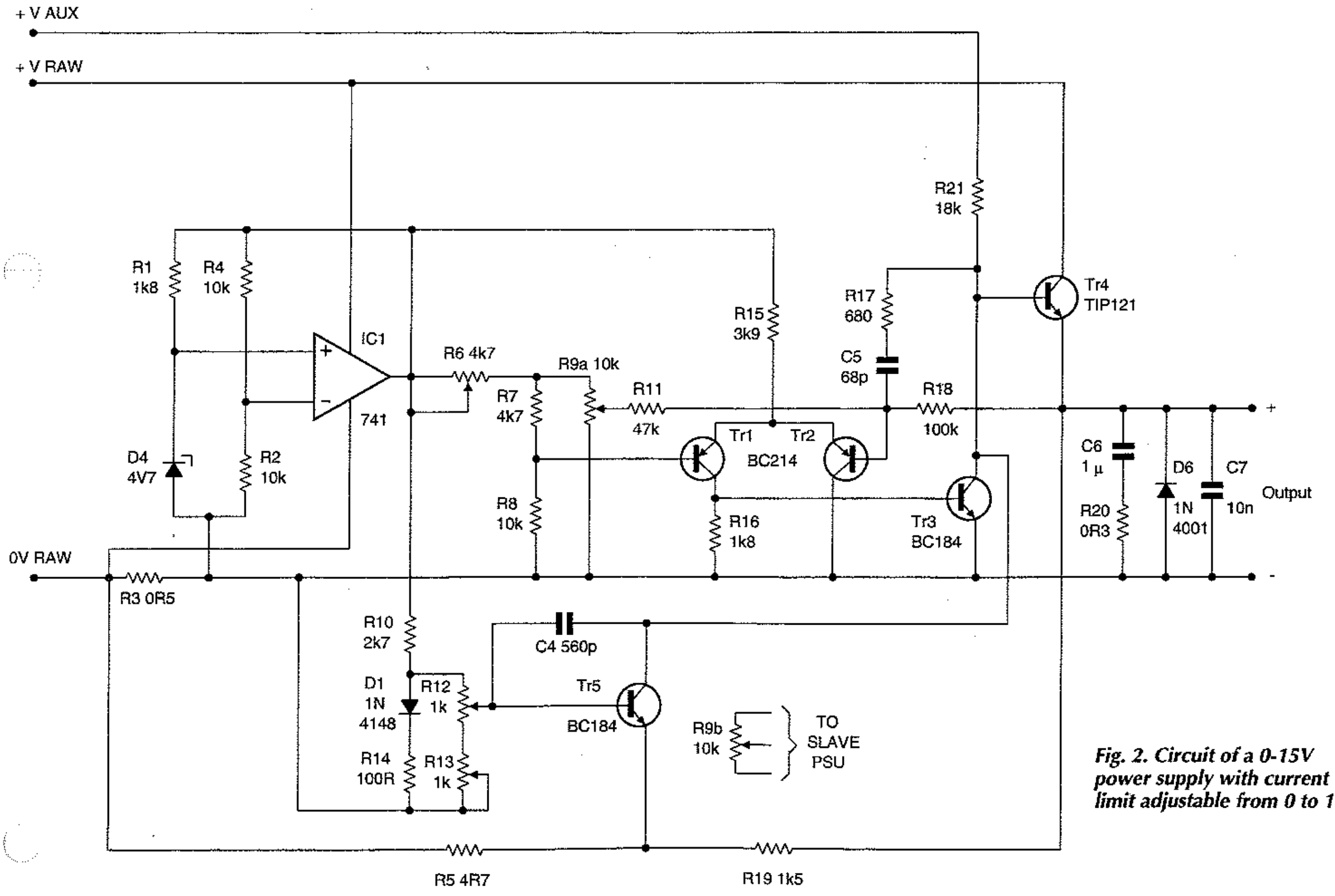


Fig. 2. Circuit of a 0-15V power supply with current limit adjustable from 0 to 1A.

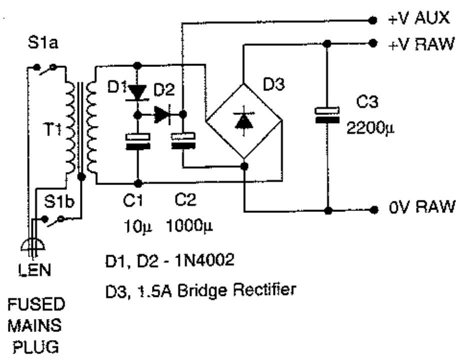


Fig. 3. Raw and auxiliary supplies. Two of these are needed, derived from separate secondaries and isolated from each other.

may prefer to use your own favourite IC voltage reference circuit, of which there are many on the market.

The op-amp provides the reference for both constant-voltage and constant-current loops. It also supplies tail current for the long tailed pair $Tr_{1,2}$. Together with Tr_3, these two replace of the IC_1 of Fig. 1.

Transistor Tr_3 drives the base of the pass transistor, a *TIP121* Darlington device which is adequate for a 15V 1A supply, given a generous heat sink. Actually, the 18kΩ resistor drives the pass transistor, Tr_3, simply sinking the excess current as necessary, to maintain the set output voltage. Capacitors $C_{6,7}$ maintain a low output impedance at frequencies where the loop gain starts to fall off. In conjunction with these, C_5 and R_{17} provide the necessary roll-off of loop gain for the constant-voltage loop. Resistors $R_{7,8,11,18}$ should preferably be 1% metal film, and R_6 permits the constant-voltage loop reference voltage to be set to 7.5V exactly.

In constant-voltage operation, Tr_5 remains cut off. At fully clockwise rotation of R_{12} its wiper is at the end of the track connected to R_{13}. This latter is set so that at an output voltage of 15V, the maximum available output current is, say, 1.1A. As R_{12} is rotated anticlockwise, the base voltage of Tr_5 rises. As a result, a smaller voltage drop across R_3 suffices to turn on Tr_5, limiting the available output current to a lower level. Transistors $Tr_{3,5}$ operate as a linear 'or' gate; whichever pulls the base of Tr_4 lower, that device controls the output voltage.

Constant current criteria

Unlike the constant-voltage loop, the loop gain of the constant-current loop is quite low. Such a low loop gain would result in the short-circuit output current being considerably greater than the maximum current available at output voltage of 15V.

This undesirable state of affairs is avoided by the judicious application of a little positive feedback from the output. The feedback is applied, via R_{19}, to the emitter of Tr_5, which is returned to the negative end of the raw supply via R_5. Thus as the output voltage falls, the additional drive, necessary to turn on Tr_5 harder, is supplied via its emitter. So an increase in output current, to provide an extra drop across R_3, does not occur. The result is that, with the component values shown, there is actually a small degree of 'fold back', that is to say that the short circuit current is actually slightly less than the maximum that can be supplied at an output voltage of 15V.

In addition, R_{19} plus R_5 form a dummy load, providing the necessary pull-down to enable the output voltage to be adjusted fully down to zero. In fact, on no-load, there is a residual output voltage of about 75mV – even when the demanded voltage is zero. This is due to some 50µA flowing via R_{11}, whose left-hand end is then at +7.5V, and R_{18}, producing the said drop across R_{19}. But this residual output voltage is of little consequence since the available current, into a short circuit, is of course no more than 50µA – even if the current limit setting of the constant-current loop be 1A.

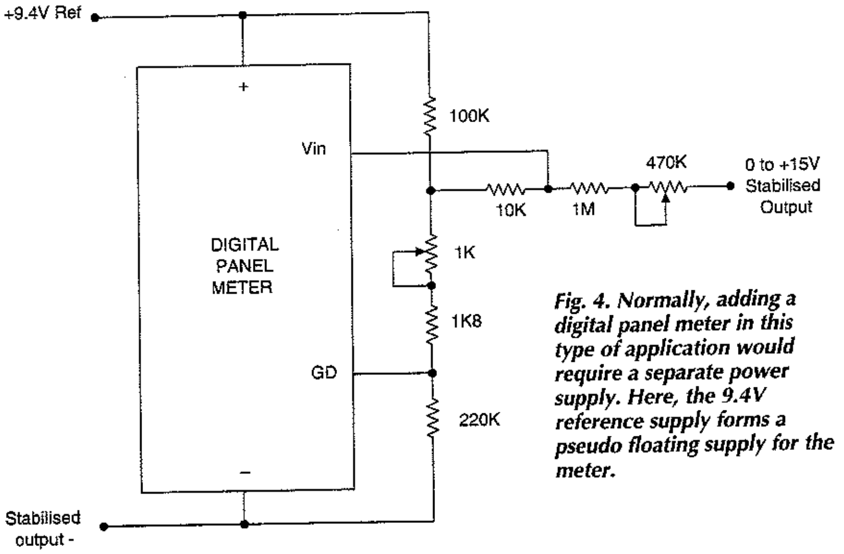


Fig. 4. Normally, adding a digital panel meter in this type of application would require a separate power supply. Here, the 9.4V reference supply forms a pseudo floating supply for the meter.

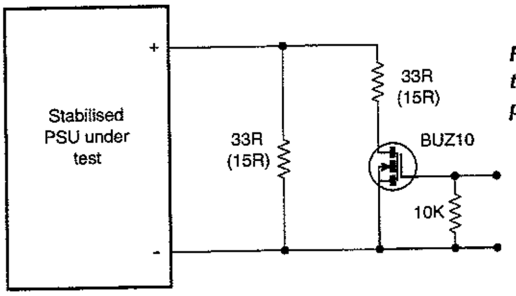


Fig. 5. Circuit used for testing the transient response of the power supply.

Duals and slaves

The mains transformer used had two similar secondaries, **Fig. 3**. These powered two identical sets of raw and auxiliary supplies – completely isolated from each other – and two almost identical Fig. 2 type stabiliser circuits.

Figure 2 actually shows the master supply, R_9 being a two-gang linear 10kΩ potentiometer. Resistor R_{9A} controls the output voltage of the master unit. The corresponding 10kΩ potentiometer in the slave is a single gang unit, its track being in parallel with that of the second gang, R_{9B}, of the master unit.

In the slave unit, R_{11} is connected to a single-pole changeover switch. This enables the slave output voltage to be controlled either by its own single-gang R_9, or by the R_{9B} of the master unit. In the latter case, the output voltage of the slave tracks that of the master, enabling their outputs to be paralleled to provide up to 2A, or connected in series to provide tracking positive and negative supplies.

Metering outputs

Having a power supply with built-in metering is useful in that it frees up the dvm for other tasks. It is particularly convenient when checking a circuit under test for correct operation over the design supply voltage range, such as 4.75 to 5.25V. Digital panel meters are available at very attractive prices, so built-in metering is no longer a luxury*. One popular type is built around the *ICL7106CPL* chip, which is produced by a number of semiconductor manufacturers.

Such panel meters consist of no more than the IC, a liquid-crystal display and a dozen or so discrete components. Designed primarily for use in small free-standing digital voltmeters, the IC is usually powered by a standard 9V PP3 battery, drawing no more than a miserly 1mA.

The basic range of a digital voltmeter based on this chip is 200mV. Series limiters and shunts are needed for other voltage ranges, and for current reading. The 200mV input terminals are designated V_{in} and *GD*, the input

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* One of the cheapest digital panel meters available in the UK, based on the *ICL7106* and available exclusively to *Electronics World* readers at a special price, is described on page 25.

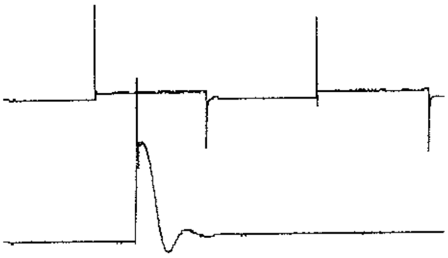


Fig. 6. Transient response of the power supply when the load switches between 0.5A and 1A; upper trace 200mV/div, 200µs/div: lower trace 200mV/div, 5µs/div.

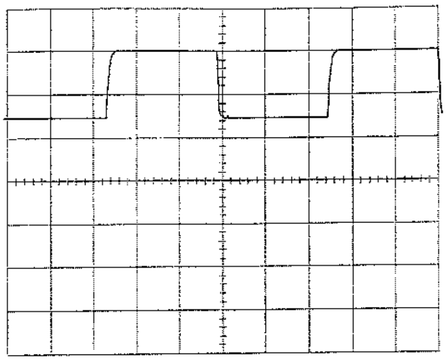


Fig. 7. Load switching between 33Ω and 17.5Ω, with the demanded output voltage set to 15V but the current limit reduced to roughly 0.5A, i.e. such that at 17.5Ω the voltage collapses to 7.5V; 5V/division, centre line = 0V, 200µs/division.

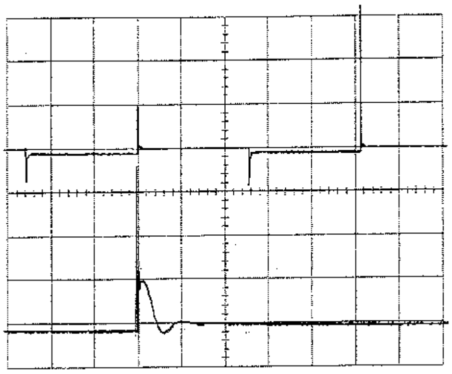


Fig. 8. Transient response of the psu when the load switches between 1A and 2A; upper trace 500mV/div, 200µs/div: lower trace 500mV/div, 5µs/div

resistance between them being more than 100MΩ.

However, the common-mode input resistance between these terminals and the negative end of the +9V supply is undefined. The IC is normally operated with the 9V battery floating, the *GD* terminal sitting at about two thirds of the supply voltage, or +6V. Though high, the common-mode input resistance is by no means to be ignored, being non-linear to boot. If the *GD* terminal is tied to a fixed voltage other than that at which it normally floats, the display shows the overload indication as a lone '1' in the left hand digit.

On the other hand, the need to supply a floating +9V is clearly an inconvenience for the designer. However, it turns out that with a little ingenuity, the 9.4V reference supply to the constant-voltage and constant-current loops can be pressed into service.

Figure 4 shows the scheme: the reference supply is used as a pseudo-floating supply by translating and scaling the 0 to 15V output to be measured to a 200mV range at the *7106*'s natural common-mode input voltage. This is carried out at a high impedance level – possible in view of the panel-meter's very high input resistance – thus avoiding pulling the common mode input voltage away from its preferred level.

The resistance values required are not what you would calculate on the basis of an infinite common-mode input resistance. The proper values are in fact not easily derived, given the non-linear common mode input resistance. As a result, I made them adjustable via trimmer potentiometers. These were set to give the right readings at output voltages of zero and +15V.

As the adjustments interact, they must be iterated to achieve the correct final settings. Adjusted thus, the panel meter agreed with the readings on a Philips *PM2521* dvm to well within ±1% over the whole 0 to 15V range. The dvm was reading the actual 0 to +15V output of the power supply, while the dpm saw a 0 to 150mV input. But linking the appropriate points on the rear pcb of the panel meter, namely jumper P2, activates a decimal point to indicate a 00.00 to 19.99 range.

Three samples of panel meter were tested in

Tips on power supply use

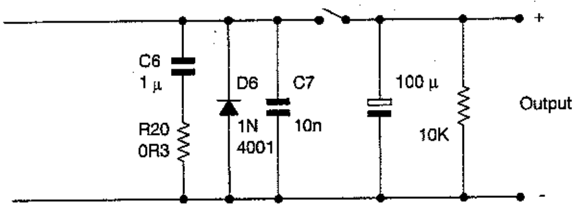


Fig. 10. When the separate output switch is closed, the 100µF capacitor causes the output voltage momentarily to collapse almost to zero. Output voltage then ramps up with the power supply in current limit, until the preset voltage is reached, or until the limited available current is drawn through a fault in the circuit under test.

With one or two amps of current available at whatever output voltage has been set, up to 15 or 30V, there is always the possibility of damage to a newly constructed prototype circuit connected to the power supply, when first powered up.

Some engineers are supremely confident of their design and workmanship, and thus have no qualms. For my part, there is always the worry that some misconnection – or even more likely, an undetected solder bridge – will result in the damage or destruction of one or more devices.

A safe way of powering up in such circumstances is to make use of the continuously variable current limit. The supply is set to the desired output voltage, and the current limit control then set fully anticlockwise, causing the output voltage to collapse to zero. The current meter is then set to a range appropriate to the current which the circuit under test is expected to draw, and circuit under test connected to the power supply.

The current limit control can now be advanced slowly clockwise, keeping a weather eye on the current meter and another on the voltmeter. If the current starts to rise alarmingly before the output voltage is anywhere near the preset value, it is prudent to switch off and recheck the circuit under test for faults. If the power supply is to be used in this way, it is advisable to use a reliable long-life potentiometer for the current limit control R_{12}, such as a cermet type.

There is an alternative mode of use, which though not offering such certain safety, will usually prevent any damage. This mode is useful where the supply is to be used by all and sundry. This is to fit an on/off switch for the power supply output, independent of the mains on/off switch. Downstream of this switch is a 100µF capacitor and discharge resistor, as in **Fig. 10**.

At switch-on, charge sharing between C_6 and the 100µF capacitor causes the output voltage to collapse to 1% of the preset value, e.g. 15V down to 150mV. The output voltage then ramps up at the set current limit until either the preset output voltage is reached, or the fault current drawn by the circuit equals the current limit.

If the fault current is only tens of milliamps – more than adequate to power a good deal of c-mos circuitry – usually no permanent damage will result, and the fault can then be cleared at leisure.

the circuit of Fig. 4. Only minor readjustments of the trimmer potentiometers were needed for each.

Current indication

A second panel meter can be used as a dedicated current meter, but an op-amp stage would be needed to suitably scale and translate the 0 to 500mV developed across R_3 to the desired level. But my personal preference for a dedicated current meter is a moving coil analogue type, since this provides an instantaneous visible indication of the current drawn. A versatile, fully protected circuit is described later on.

Using a digital panel meter, with its reading rate of about three readings a second, and allowing for settling time, a clear indication of the current drawn would not be instantly available. Indeed, if the current being drawn by the load has an appreciable ripple, the last few digits may be constantly flashing.

An analogue meter, by contrast, has a degree of built-in smoothing, due to the inertia of the movement. Nevertheless, a digital readout of current can be useful for testing purposes, so perhaps the best of both worlds would be an analogue meter permanently indicating the current being supplied, and a digital meter normally indicating output voltage, but switchable by means of a biassed toggle, to read current when required.

A useful performance

I tested the 15V 1A power supply of Figs 2 and 3 for the usual performance parameters, with the following results.

Direct-current output resistance measured 50mΩ, while the change in output voltage for a 10% change in mains voltage was barely 1mV. Output ripple in constant-voltage mode, supplying 1A at 15V, was estimated at around 200µV peak-to-peak, as measured on the 2mV/division range of a Thurlby-Thandar digital sampling adaptor type *DSA524* with averaging mode selected.

In view of the low signal level, to avoid possible errors due to earth loops, the reading was repeated, using the audio-frequency millivoltmeter section of the laboratory amplifier described in Ref. 1, with its balanced floating input stage. There was no indication on the 3mV rms full scale range, confirming that the full load ripple is below 100µV rms.

With the same load resistance and set voltage, the current limit was reduced to enter constant-current mode. Ripple voltage across the load was then 8mV pk-pk at 900mA, reducing *pro rata* with current, reflecting the lower gain of the constant-current loop.

An important parameter of a power supply is the transient response when the demanded load current changes abruptly. **Figure 5** shows a simple test circuit which was used to switch the load between 0.5A and 1A approximately, at a rate of 1kHz. The transient was captured using the *DSA524*. The result is illustrated in **Fig. 6**, at 200µs/division, upper trace, with an expanded view of the transient at 5µs/division, lower trace.

When the load drops from an amp to half an amp, there is a momentary positive-going spike of some 700mV. But since the width of this measured out at just 100ns, the energy associated with it is low. Thereafter, there is a well-controlled transient, settling within 10µs to the steady level.

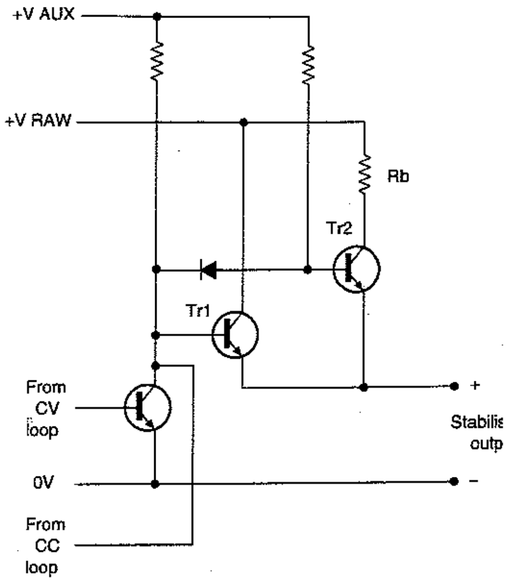


Fig. 9. Updated version of the McPherson regulator. Of the worst case total dissipation, only around a third ever appears in either output transistor.

The story when the load switches from 0.5A to 1A is similar; the spike just looks smaller in the upper trace as a sampling pulse does not happen to have caught the peak. **Figure 7** shows the same load and set voltage, but with the current limit set to roughly 0.5A, so that at the lower value of resistance, the output voltage drops to 7.5V. The response is overshoot-free, as the constant-current loop is, if anything, overdamped.

The prototype is stable both on and off load in both constant voltage and constant current modes with 1000µF in parallel with the output. Of course, a 1000µF capacitor reduces the 7.5/15V switching waveform of Figure 7 to pretty well an 11V straight line, and even just 10µF turns it into something approaching a triangular wave.

Variations on a theme

As mentioned in the introduction, the circuit is designed to be 'stretchable', both in voltage and current. Typical ratings for commercial laboratory bench power supplies are 15V or 30V, at 1A, 2A or occasionally 5A.

Figure 8 shows the output of the psu when the load switches between 1A and 2A, the 33Ω resistors in Fig. 5 having been replaced by similar wirewound 15Ω resistors. As the raw supplies with pass transistor Tr_4 and its heat sink were not rated for continuous use at 2A, the test was not continued for longer than necessary to obtain the results shown.

To enable the unit to provide 2A, even in the short term, current sensing resistor R_3 was temporarily shorted to defeat the current limit – not a practice to be recommended. A proper 2A version requires only the beefing up of the raw supplies, a pass transistor with a higher maximum dissipation than the *TIP121* used in Fig. 2 – with suitable extra heatsinking – and halving the values of R_3 and R_{21}.

Similarly, few changes are required for a 30V version, other than attention to voltage

Supplying the panel meter

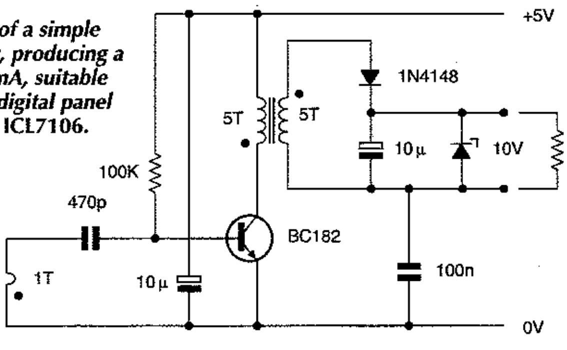
The stratagem described in the main article, to permit the panel meter to be powered from a non-floating supply, is not always convenient. In this case, an inverter can be used to produce a suitable floating 9V supply from whatever rail voltage is available.

Figure **11** shows a very simple flyback inverter for operating a digital panel meter from a +5V rail. At under 60%, the efficiency when supplying close on 10V at 1mA, is not wonderful. However, the odd 3.8mA is hardly a heavy load on the 5V supply.

The prototype circuit ran at about 170kHz, producing 9.52V off load, 9.46V into a 10kΩ dummy load simulating a digital panel meter. The dual five-turn windings were of bifilar wire, on a Mullard *FX2754* two-hole balun core having an A_L of 3500nH/turn^2.

Direct-current-wise, the output voltage is floating, but the 100nF capacitor is added to prevent switching frequency ripple appearing on the output relative to ground. The circuit is readily adapted for other supply voltages, and as the required output power is less than 10mW, efficiency will not usually be an important consideration.

Fig. 11. Circuit of a simple flyback inverter, producing a nominal 10V 1mA, suitable for powering a digital panel meter using the ICL7106.



ratings of capacitors and semiconductors – and one other point. If you are using a $3^1/_2$-digit panel meter in a 30V version, provision must be made to switch the latter from 19.99V full scale to 199.9V full scale. A useful halfway house, providing more than 15V output but without the complication of dpm range switching, is a 20V design. This will enable circuitry designed for either 15V or 18V nominal supplies to be tested at both top and bottom supply limits.

Whatever the rating chosen, a useful feature to incorporate is a non-locking push-button wired across the output terminals. Pressing this will put the psu into current limit, and R_{12} can then be adjusted for a lower limit than the maximum, if required.

More variations

The *TIP121* Darlington is so cheap and convenient, it is worthwhile considering whether it can be used in higher power designs. For example, in a 15V 2A design, two can be used in parallel. Each needs to be fitted with a 0.5Ω emitter ballast resistor to prevent current hogging by one of them. Heatsinking must be adequate to handle the total worst case dissipation, with a short circuited output and the highest mains voltage. However, the two devices are equivalent to a single Darlington with half the junction-to-heatsink thermal resistance of a single device.

For even higher powers, the McPherson circuit, Ref. 2, is attractive. Its patent has probably by now expired. An updated version of

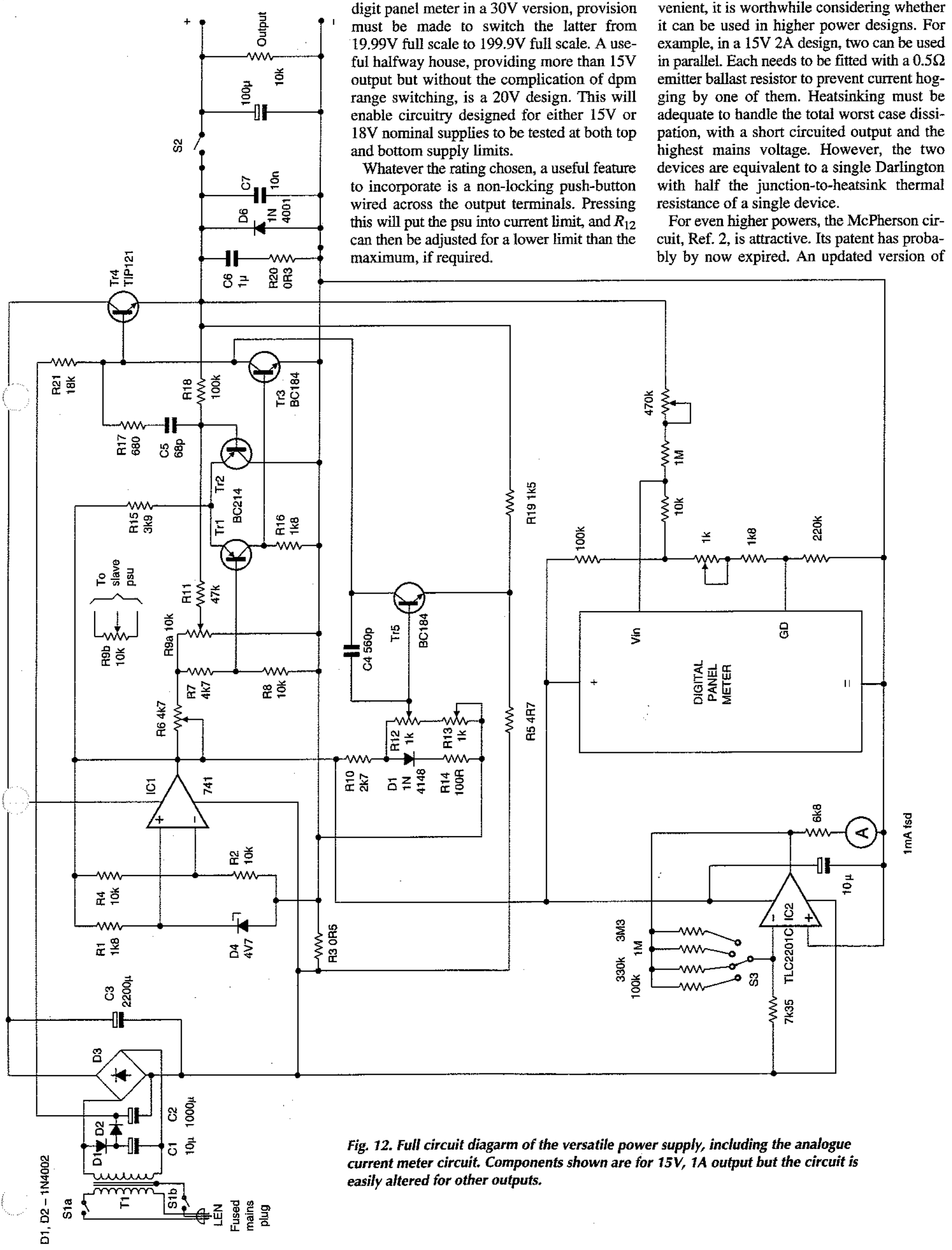


Fig. 12. Full circuit diagarm of the versatile power supply, including the analogue current meter circuit. Components shown are for 15V, 1A output but the circuit is easily altered for other outputs.

this scheme is shown in **Fig. 9**. If you imagine the raw voltage to be only marginally greater than the maximum rated output voltage, for example at minimum mains voltage, then only a quarter of the worst case power dissipation ever appears in either transistor; often it is much less. This is because at rated maximum current on short circuit, Tr_1 is cut off, Tr_2 bottomed, and all the dissipation takes place in ballast resistor R_b, which is $V_{rated(max)}/I_{rated(max)}$. At maximum rated current at maximum output voltage, Tr_2 can make no significant contribution, so all the current is supplied via Tr_1, whose V_{ce} is then however minimal.

There are two worst cases; the first is full output current at half output voltage. Here, Tr_2 is bottomed and supplies half the current, while Tr_1 supplies the other half, with a V_{ce} of half the raw volts. The other is negligible output voltage at half rated current. Here, Tr_1 is off and Tr_2 supplies half the rated current with half the raw volts collector to emitter. Either way, only a quarter of the maximum power dissipation appears in either transistor, and never in both at the same time, so they can usefully share the same heatsink.

In practice, the worst case transistor dissipation is somewhat more than this, especially at top mains voltage. But it is still much lower than schemes where all the dissipation occurs in pass transistors. Clearly, a considerable saving in the heat sinking requirements is achieved. Most of the dissipation occurs in the wirewound resistor, or resistors, which can handle heat at a 300°C surface temperature, against 125°C for a semiconductor junction.

Reference 2 describes how the scheme can be extended to four transistors, three with appropriate value resistors in their collector circuits. Turning on one or more, as required, in sequence, keeps most of the dissipation in the various ballast resistors, a very effective arrangement.

Variations on the current limit circuit are also possible. **Figure 10** shows a versatile analogue current meter circuit. An op-amp is used to amplify the 0.5V maximum drop across the current sense resistor R_3 to 6.8V, to drive a 1mA full-scale deflection meter, scaled 0 to 1A and 0 to 300mA.

Other values of feedback resistor may be selected, giving a choice of 30, 100, 300 and 1000mA ranges. On the most sensitive of these, the full-scale voltage drop across R_3 is only 15mV, so an op-amp with low offset voltage is indicated.

A *TLC2201/C* being to hand, this device – with its typical offset of 100μV – was used. In fact, with its low maximum input offset of 500μV (or 200μV on the /AC and /BC versions), the *TLC2201* comes without offset adjust inputs, and at 1pA its bias current is not large either. But a more mundane op-amp, complete with offset adjustment, would suffice.

The circuit shown protects the meter against overload. If the psu supplies 1A when the meter is switched to the 30mA range – representing a 33× overload – the op-amp output can only reach something less than +9.4V, limiting the actual meter overload to less than 50%.

Another variation can be useful, where the maximum power available from the raw supply at +7% mains voltage, is greater than the pass transistor can dissipate indefinitely with the output short-circuited. For example, on a 15V 1A unit, the current limit could be set at 1.5A at 15V, folding back to 1A when the output is shorted. This merely involves raising the value of R_5. A further ploy is to thermally couple Tr_5 to Tr_4; the short circuit current can then be set to, say, around 1.2A with the unit cold. On an extended short circuit, the V_{be} of Tr_5 will then fall by about 2.2mV/°C as the heatsink and pass transistor warm up, gradually reducing the short circuit current back to 1A.

References

1. Hickman, I, Listening for clues *Electronics World* July/August 1996 pp 596-598.
2. McPherson, J W, Regulator Elements Using Transistors *Electronic Engineering*, March 1964, p. 162.

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