Ian Hegglun thinks it can, and explores how the technique can be applied.

Switched capacitor power supplies

Switching power supplies normally use inductors and/or transformers for voltage conversion. But this is not the only way to transform voltage efficiently. Voltage multipliers and voltage dividers are another option.

Voltage dividers are less well known. They are simply multipliers reconfigured for voltage step down. There are lots of similarities. Like multipliers, they need no inductors and like multipliers they are limited, mainly because they are difficult to regulate efficiently over a wide range of input or output voltages.

Pulse-width modulation can certainly vary the average voltage of inductorless circuits but the losses in the smoothing process are hardly any different to linear regulation. This is because the equivalent circuit of a switching capacitor arrangement is a low-pass *RC* filter which is inefficient when pulse modulated at low duty cycles. However, my earlier work showed efficient regulation is possible if a small value air-cored inductor is added to ac input of a full wave multiplier[1]. Regulation is then achieved by varying frequency.

This article looks at step down switched capacitor converters. In particular, converters that are capable of delivering tens of watts at 3V and converters delivering hundreds of watts at around 50V output. When cascades, these converters could form the basis for computer power supplies, but how practical is this?

240V Mains to 3V without inductors?

Breaking the task into two stages with an intermediate voltage at around 30 or 50V helps. This would work in well with the distributed supply technique where one main power supply delivers only one voltage which then supplies various boards, each board has their own converter for their particular needs.[2,3] This solves a number of problems and makes it relatively easy to add batteries to the intermediate voltage for an uninterruptible supply.

Each converter requires a step-down ratio of at least 8:1. Since there are so many stages, each stage must be very efficient. If you aim for, say, 70% overall efficiency then each of two converters need to be around 85% efficient. With two or three cascades within each converter, efficiency for each stage must be no less than 95%.

Although difficult, this is not impossible. The increasing use of synchronous rectifiers such as the *MA767* can give this efficiency even at the 3V level.[4] Recent developments

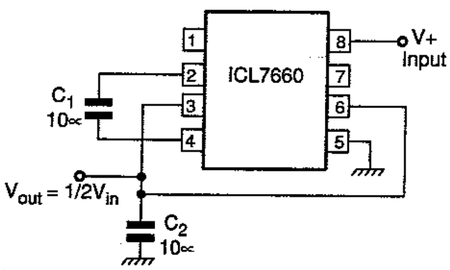


Fig. 1. Voltage divider using a popular IC.

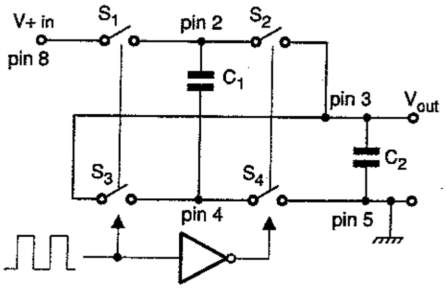


Fig. 2. Voltage divider internal circuit.

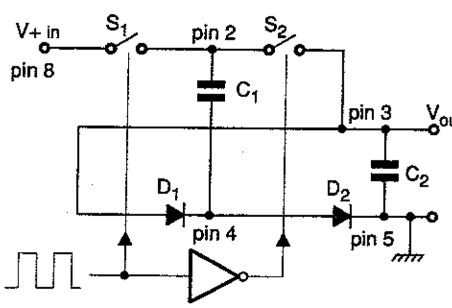


Fig. 3. Diodes being used in place of S_3 and S_4.

with mosfets, namely Philips' *Trenchmos* and SGS-Thompson's mesh overlay, should reduce the cost of the silicon needed for low losses.

Other necessities for mains conversion are power factor correction and an earthable or non-floating output. Both of these can be provided without using an inductor or transformer.

Literature on voltage dividers

One reference to a voltage divider can be found in the *ICL7660* data sheet.[4,5] **Figure 1** shows the connections required and **Fig. 2** shows the internal switching with mosfets.

In the first half cycle when S_1 and S_3 are closed, the supply voltage divides across the capacitors in a conventional way proportional to their value. In the next half cycle when S_2 and S_4 are closed, the capacitors switch from a series connection to a parallel connection. This forces the capacitors to have the same voltage; the charge redistributes to maintain precisely $^1/_2$V+, across C_1 and C_2.

A Linear Technology application note describes using four discrete mosfets to convert 12V to 5V at 1A without using an inductor in a similar way[6]. Regulation over a small range is achieved by varying the charge duration while S_1 and S_4 are on.

Two of the switches, S_3 and S_4, can be replaced with diodes, **Fig. 3**. With diodes, efficiency is reduced for light loads. This is because mosfets acting as synchronous rectifiers in Fig. 2 can have a low forward voltage drop at low currents, unlike normal diodes.

In the diode version, the output capacitor can be removed and the circuit still operates. Ripple voltage increases, as expected, but the ripple frequency halves – something not expected.

Investigating further, it appears that a minimum output capacitance is needed to pull the ripple voltage average of the first phase to the same level as the second phase. Simulation on *Tina* showed this effect. The minimum capacitance appears to be around $^1/_3C_1$, although in tests the ripple current rating requires a minimum of $^1/_2C_1$.

I also tried simulation on *Electronics Workbench* V4. Although the simulation ran, it did not give the correct waveforms using the same models as *Tina*. In these simulations, the mosfet switches were simply voltage-controlled switches with a series resistance to simulate mosfet on resistance. Doing this speeds up simulation considerably and is obviously acceptable with *Tina*.

Simplifying the switches in this way may explain why *Workbench* gave incorrect results, but I am unable to repeat the analyses with proper mosfets since I no longer have access to the software. Anyone interested in seeing the Workbench file can do so by sending an stamped, self-addressed envelope to *Electronics World*'s editorial offices.

Using an equivalent circuit

The equivalent circuit is useful for predicting output voltage, ripple voltage and efficiency of a converter. **Figure 4** shows the equivalent circuit. The derivation is given in a separate panel. The voltage source is $^1/_2V_{in}$ and there is a V_D diode drop with a triangular ac ripple voltage (ΔV_{pp}) superimposed on the output.

Ripple voltage is determined mainly by C_1. Bear in mind that C_1 supplies output current for both phases in this type of converter. Therefore is best to place most of the available capacitance at C_1.

Finally, there is a series resistance of $R_{DS(on)}$ for the mosfet

ESR of aluminium electrolytic capacitors

Losses in a capacitor are represented by an equivalent series resistance, or esr. Low esr is needed in switching power supplies. A low esr means low internal heating due to ripple current and low ripple voltage. It is common to parallel two or more small capacitors to give an esr lower than a single capacitor of the same total value.

Electrolytic losses vary with frequency and temperature. The graph shows the impedance of several aluminium electrolytic capacitors. At higher frequencies the capacitor becomes net inductive. Some capacitors have a flat region over one or two decades of frequency.

Minimum esr is also related to voltage and capacitance. In general, the same cans size fixes the esr and hence ripple current rating. This can be seen by scanning through the ripple current ratings for various capacitors of the same size within the same brand range.

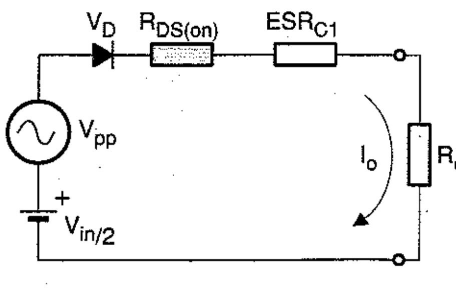
For non-solid electrolytic capacitors reliability is related to power loss (average I^2esr) and the ambient temperature.

Philips' electrolytic capacitor data book 8 states that a 50% drop in reliability can be expected for each 10°C increase in temperature. Solid electrolytics are not degraded by temperature to the same extent and offer much longer life. Philips quotes a failure rate of 10^{-9}hours for solid electrolytics compared to 10^{-6}hours at a 60% confidence level for their non-solid electrolytics.

The company also gives an equation for calculating ripple current.

$$I_R = \sqrt{\frac{P}{ESR}} = \sqrt{\frac{\alpha S(T_C - T_{amb})}{ESR}}$$

where I_R is the ripple current (A), P is heat dissipation (W), a is the heat-transfer coefficient (W/m²/°C), S is surface area (m²), T_C is the case temperature (°C) and T_{amb} is the ambient temperature (°C).



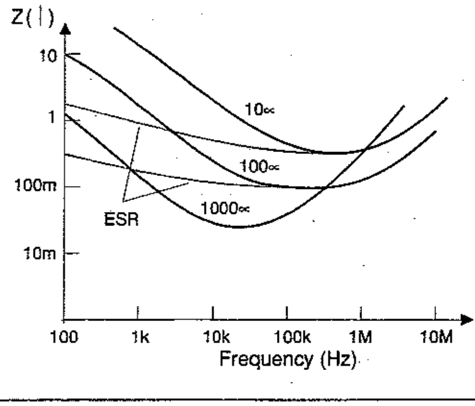


Fig. 4. Equivalent circuit for Fig. 3, and curves showing the variation of impedance with frequency for various aluminium electrolytic capacitors.

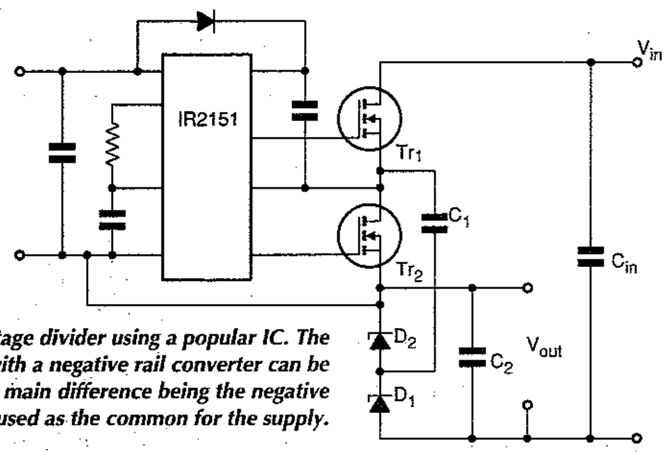


Fig. 5. Voltage divider using a popular IC. The similarity with a negative rail converter can be seen – the main difference being the negative rail is used as the common for the supply.

switches plus ESR_{C1}, as in the diagram in the panel. With significant dead time the losses must be increased according to the form factor.

A 30 to 15V converter

Figure 3 is a relatively easy circuit to implement. Test circuit **Fig. 5** uses an *IR2151* to drive two mosfets. Three 6.8μF, 20V tantalum capacitors measuring 7 by 5 by 4mm were paralleled for C_1. Each capacitor has an esr of 400mΩ at 100kHz. Two of these capacitors are used for the output C_2.

The input capacitor consists of three 22μF, 50V YXB aluminium electrolytics. These allowed the circuit to deliver 5A continuously. The input voltage was adjusted to give 15V output , requiring 34.9V.

Efficiency is easily calculated by 2×15/34.9, giving 87% efficiency. This calculation is possible because the ratio of output current to input current is precisely 2, due to charge conservation and current used by the driver is negligible at full load.

Analysing losses, 2W is lost in diodes, 2W in capacitors, mainly C_1, and 4W in the mosfets. This gives a total of 8W for 75W output.

Figure 6, top trace, shows output ripple voltage with a 5A load: the average of this ripple is offset 15V above ground. The lower trace shows the voltage at the output of the mosfets feeding C_1. Note that the ripple frequency is twice the oscillator frequency.

Dead time can also be seen. It totals about 2μs which means about 40% more current is required causing 40% more ripple and losses. Current through C_1 is 10A peak falling to 5A or 7.5A average. You may notice the similarity with a negative rail converter. The main difference is that the negative rail is used as the common for the supply.

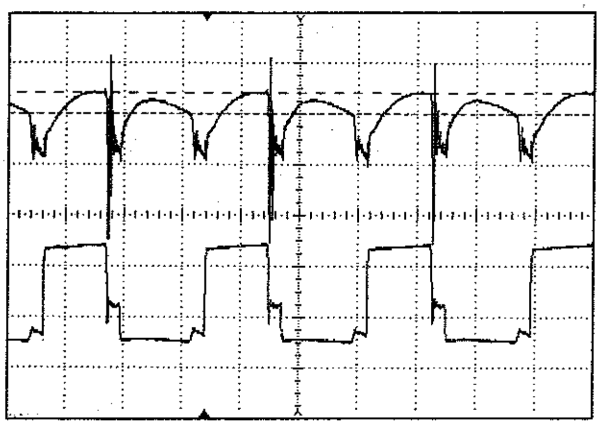


Fig. 6. Ripple voltage (upper trace 1V/div) and the ac drive voltage to C_1 (lower trace 10V/div) with a 5A load. Time base is 5μs/div. The ringing was traced to the driver. Adding gate damper resistors of 47Ω eliminates this ringing.

Due to the dead time, the ripple is increased by 50%. Increasing the frequency fails to reduce ripple because the dead time component remains constant. Adding a small 1μH inductor to form a low-pass filter between the load and output with a shunt capacitor across the load can reduce ripple to low levels if required.

The inductor used was removed from a 10A switching power supply output and measures 6mm dia. and 15mm long with five turns of 1.5mm wire. With a 220μF 200mΩ esr electrolytic the ripple was reduced by a factor of five with a 70kHz operating frequency. If the dead time could be reduced and the frequency increased to 150kHz the ripple would fall to around 10mV rms with this filter.

With 5A load, the tantalums ran at about 50°C, or about 35°C above ambient. Operating the converter at higher ambient temperatures, of say 50°C, the current would need to be halved. This means that ripple voltage would also be halved.

Another method for reducing ripple is by running two converters in parallel and clocking them 90° out of phase. This can reduce ripple by a factor of two – more with an *LC* filter – while also doubling the output current. This technique can also be applied when cascading two converters for higher step down ratios.

At about 1A per capacitor, this is a remarkably high power density. The chip inside measures only 3mm diameter and 3mm long, or $^1/_{25}$th the volume of an aluminium capacitor of similar rating. These capacitors were removed from a computer pcb some time ago and I don't know where they can be sourced. However, similarly rated special polypyrole *ECG* series electrolytics by Panasonic are readily available.

Later tests on an ECG 6.8μF/16V electrolytic revealed an esr of only 40mΩ – the data sheet quotes 0.4Ω maximum at 400kHz – and a minimum impedance at 2MHz. It appears some manufacturers are very conservative when it comes to esr values, possibly because there is a large spread in values. These ECG capacitors are rated at 1A at 105°C, but if they are sorted, most will run at higher currents. Although to make use of this capacity, the frequency needs to be closer to 1MHz.

I also tested a 4.7μF/16V multilayer ceramic in a 1206 package. Its capacitance measured on a dmm was only 3μF at 25°C and only 1μF at 1MHz. Minimum impedance occurred at 8MHz with an esr of 130mΩ. It uses Y5V dielectric and has a temperature coefficient of 1%/°C from 25°C which means the capacitance falls to $^1/_5$th of its room temperature value when heated to 80°C. These Y5V capacitors are not practicable for converters that operate above 40°C.

Comparison with buck

Comparison with a conventional buck converter is a useful exercise. The *L296* application note gives complete design details for a buck converter[7]. Equations for the value of the inductor and output ripple voltage are given below assuming the inductor ripple current is the usual 30% of the peak inductor current,

$$L = \frac{(V_i - V_o)V_o}{0.3V_i\, f\, I_o}$$

$$\Delta V_C = \frac{(V_i - V_o)V_o}{8V_i f^2 L C_O}$$

$$\Delta V_R = 0.3\, I_o\, ESR_{Co}$$

Note that the capacitor ripple voltage is made up of two parts. Since theses components are in quadrature they should not be added algebraically, although, if one component dominates by more than three times, the other can be neglected, due to the nature of quadrature addition. For a 2:1 buck configuraand $V_i = 2V_o$ the inductance required and ripple voltages are,

$$L = \frac{V_o}{0.6 f I_o}$$

$$V_C = \frac{I_o}{24 f C_o}$$

$$\Delta V_R = 0.3 I_o \, ESR_{C_o}$$

Using the *L296* as described in the application, the frequency is 100kHz and two 100μF and 150mΩ esr output capacitors are paralleled. When loaded at 4A the ripple voltage is calculated to be 90mV pk-pk and esr ripple dominates.

For the switching capacitor converter of Fig. 3 the ripple voltage is approximately,

$$\Delta V \cong \frac{I_o}{4C_1 f} \cong \frac{I_o}{3\, C_O f}$$

where C_2 is $^2/_3\, C_O$ and C_1 is $^1/_3\, C_O$. The input capacitor is assumed to be the same in both cases. This equation comes from the equation $I_C = C^{\Delta V}/_{\Delta t}$, where the capacitor current pulse rate is $2f$ and current and the ripple voltage across C_1 is divided by two at the output and verified by measurements. It was also verified that there is no esr ripple component present – unlike the buck converter.

Using the same value of output capacitance as used in the *L296* example, the output ripple at 4A is 60mV pk-pk compared to 90mV. However, since the capacitors used in this example are only carrying 1.2A of ripple current (0.3×4A) the capacitance required for the same temperature and same life must be three times higher for a third the esr. A plus for the switching capacitor converter is lower electromagnetic radiation, lighter and more compact. The buck scores high on being able to vary the ratio over a wide range efficiently and requires less components.

Silicon utilisation

The buck converter requires one transistor, minimally rated at V_{in} and I_o (average) plus a diode for the same V_{in} and I_o. A switching capacitor converter such as Fig. 3 requires two transistors, each minimally rated at $^1/_2 V_{in}$ and I_o, plus two diodes each rated at $^1/_2 V_{in}$ and I_o. This is twice the number of devices although the total switching capacity ($V \times I$) is the [illegible].

With more parts there is a cost penalty but it also helps to distribute heat and improves reliability. Distributing heat makes natural cooling easier which improves reliability. In many cases a fan can be avoided which is often the most unreliable component.

Single converter divide-by-*n*

A circuit for a 4:1 step down, **Fig. 7**, has appeared in this magazine[9] and more recently in another magazine[10]. In this arrangement four capacitors, labelled $C_{1\text{-}4}$ on the circuit, are charged in series and later discharged in parallel. They operate as a Mosmarx[11] step down converter.

With this technique a 60V peak-to-peak input voltage can be stepped down to around 12V with an efficiency of around 75% using conventional diodes or around 90% with schottky diodes. Adding each extra stage in this circuit involves three additional diodes and one capacitor.

Figure 8 shows a modification that allows the step down ratio to be varied from 4:1 to 1:1 in 4 steps which is useful for regulation or power factor correction for off-line converters. A ratio of three for example is obtained when Tr_1 is held on for both phases, bypassing C_1 during the charge phase and therefore charging three capacitors in series rather than four.

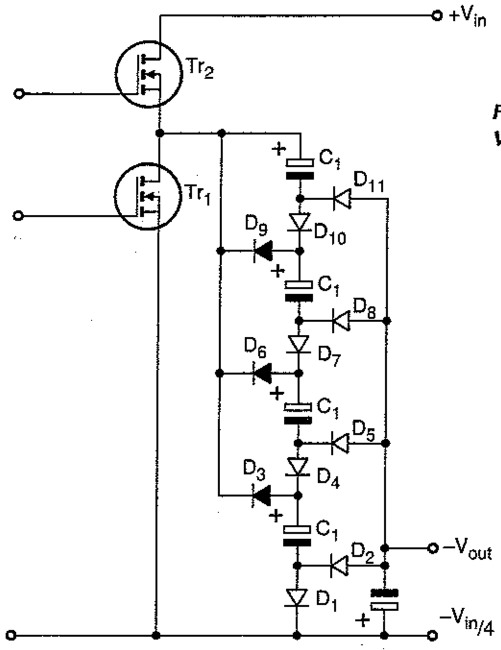


Fig. 7. This network converts V_{in} to negative $^1/_4\ V_{in}$.

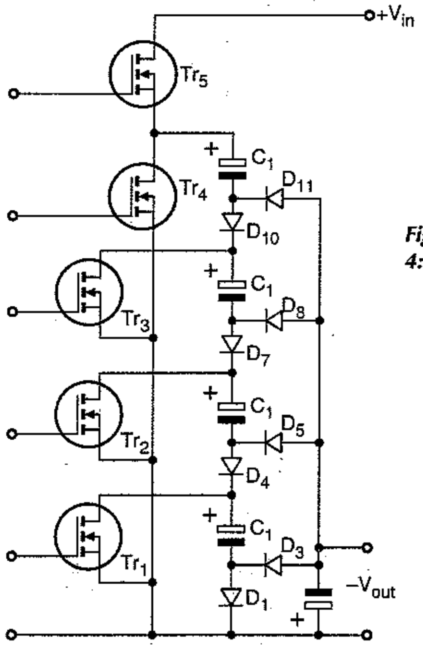


Fig. 8. Allows variable ratios of 4:1 to 1:1 in 4 steps.

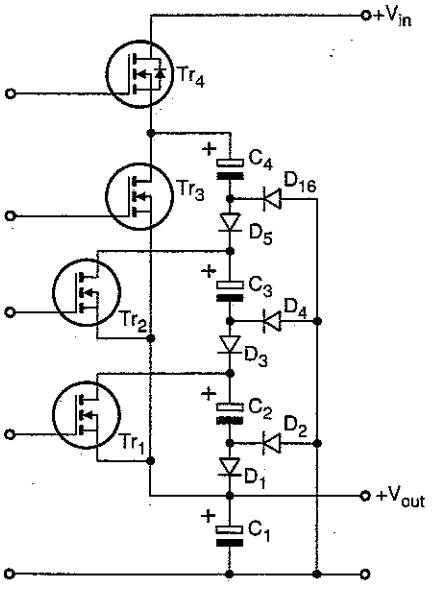


Fig. 9. Converts Vin to positive 1/4 Vin with 4 steps.

If a low drop series regulator is added after the converter, the ripple can be removed and regulation over a wide range of input voltages is possible with high efficiency.

Figure 9 shows a non-inverting version. It uses two fewer diodes, one fewer transistor and one fewer capacitor. This makes it slightly more efficient. However, the transistors in a selectable ratio converter requires individual high side drive making driving more difficult.

So can the Cockcroft Walton multiplier be reversed for step down to give ratios higher than 2:1? **Figure 10** shows how it can using mosfets on the improved version.[12] This allows the number of diodes to be reduced, improving efficiency.

Parallelling two dividers as in **Fig. 11** reduces the ripple in the input current and allows much smaller input and output capacitors. Minimal dead time is required for best results.

Equations for switching converters

Neglecting ac ripple component on C_1 and the effect of energy loss from capacitor charge redistribution, the equations for steady state output are as follows. In all cases the output voltage is slightly lower due to charge redistribution losses[4] but appears to contribute only a few percent of losses since ripple is usually less than 10% of the dc voltage.

Synchronous converter, Fig. 2. All switches are mosfets with equal $R_{DS(on)}$ since they carry the same peak current.
Charge phase,

$$V_C = V_{in} - I_O\left(ESR + 2R_{DS(on)}\right) - V_{out} \qquad (1)$$

Discharge phase,

$$V_{out} = V_C - \left(ESR + 2R_{DS(on)}\right)I_O \qquad (2)$$

Substituting (2) into (1),

$$V_C = V_{in} - I_O\left(ESR + 2R_{DS(on)}\right) - V_C - \left(ESR + 2R_{DS(on)}\right)I_O$$

So,

$$V_C = \frac{V_{in}}{2}$$

giving,

$$V_{out} = \frac{V_{in}}{2} - \left(ESR + 2R_{DS(on)}\right)I_O$$

Single ended converter using diodes, Fig. 3. Each diode has a forward drop V_D and ohmic resistance R_D.
Charge phase,

$$V_C = V_{in} - I_O\left(ESR + R_{DS(on)} + R_D\right) - V_D - V_{out} \qquad (3)$$

Discharge phase,

$$V_{out} = V_C - I_O\left(ESR + R_{DS(on)} + R_D\right) - V_D \qquad (4)$$

Substituting (4) into (3) also gives,

$$V_C = \frac{V_{in}}{2}$$

hence,

$$V_{out} = \frac{V_{in}}{2} - V_D - \left(ESR + R_{DS(on)}\right)I_O$$

From this, the equivalent circuit of Fig. 4 can be drawn.

Parallel converter, Fig. 11. This circuit shares the output current with $I_O/2$ through each stage giving,

$$V_{out} = \frac{V_{in}}{2} - \left(\tfrac{1}{2}ESR + R_{DS(on)}\right)I_O$$

Converting from 30V down to 3V

This 30V to 15V converter gives valuable insights into the operation of a divide-by-two converter. At the time of writing, I had not tested the 30V to 3V converter, although a 24V to 12V synchronous rectifier version, as in Fig. 2, has been run, giving the target of 95% efficiency. To get this efficiency the output current is only a fraction of the mosfets current rating. The latest mosfets will improve this.

Philips now offers a 55V device in a TO220 package with an 8mΩ on resistance. Even lower resistance devices in smaller packages are likely soon.

Conversion from 30V to 3V requires a ratio of 10:1. This can be achieved a number of ways with dividers, either two divide-by-threes or three divide by twos. It appears that the $V \times I$ capacity is about the same in both cases.

A practical converter requires regulation of the output voltage. If the input varies over a wide range it is not practical to use pulse modulation as mentioned.

Recent tests using a small inductor within a divider demonstrated efficient regulation over a 2:1 input voltage range. This allows the 30V input on the low voltage converter to

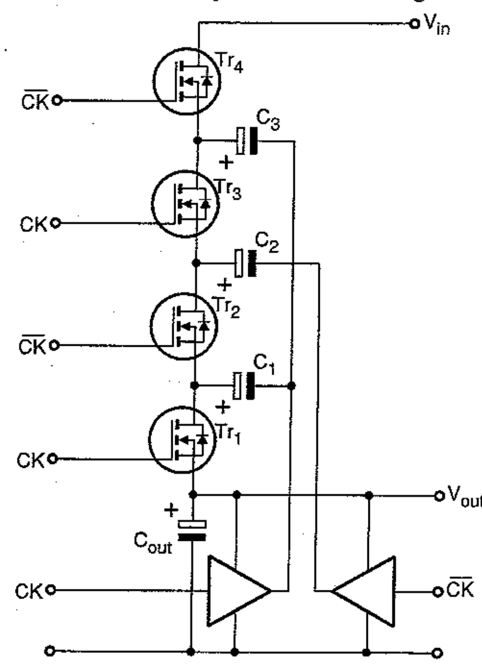


Fig. 10. Improved Cockcroft Walton multiplier can be reversed if diodes are swapped for mosfets.

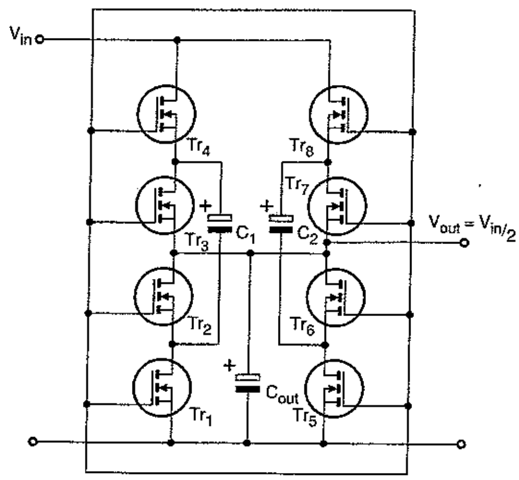


Fig. 11. Parallelling two dividers reduces the ripple in the input current and allows much smaller input and output capacitors. Minimal dead time is required for best results.

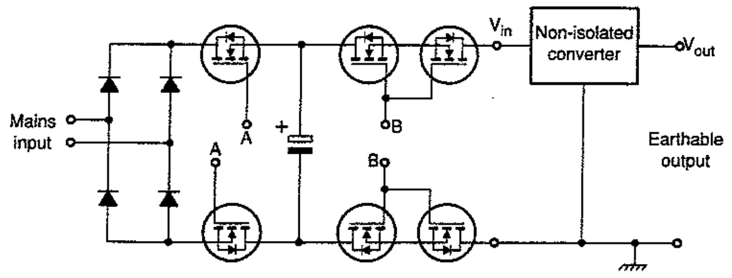


Fig. 12. The flying capacitor technique allows the normally floating output of a bridge rectifier to be connected to earth without damaging the circuit.

> ***Safety warning**
>
> When Ian speaks of mains isolation, this does **not** mean safety isolation. Device failure could result in a direct path between the mains live side and the converter's output. As a result, all parts of any such converter should be considered to be lethal, and never touched. Wiring regulations in your country should also be consulted before such a circuit is connected – Ed.

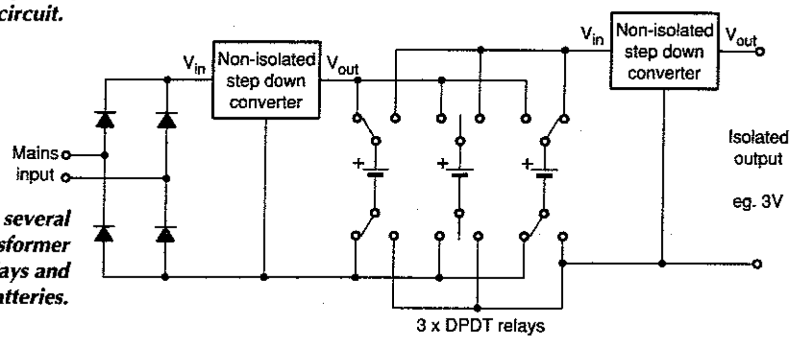


Fig. 13. Isolation to several kilovolts without a transformer is possible using three relays and switching batteries.

rise to 60V plus ripple filtering of the reservoir capacitor on this bus. I hope to present design details in a later article.

Mains to 50V conversion

The preferred arrangement for the mains converter is the circuit in Fig. 9, but with eight stages.

This arrangement uses fewer components and it allows the ratio to be varied in small increments for voltage pre-regulation and power factor improvement. It does this by varying the ratio over the mains half cycle. Current flows over most of the cycle in a sinewave like fashion.

At 50V output, this converter can use standard diodes giving about 2V in 50V loss and an efficiency above 95%. This is well above the target of 85% for the mains converter discussed earlier. It should be possible to get 80% overall for mains to 3V without using unrealistic amounts of silicon.

Isolating the output

The flying capacitor technique has been proposed as a way of 'isolating' the converter common from the mains[13]. Although this does not provide the same degree of isolation from the mains as a transformer it does allow the converter's low-voltage output common to be connected to an earth and it allows the mains to be rectified using a full-wave bridge rectifier.

Normally, the output of a bridge rectifier must be floating relative to the ac input terminals. The flying capacitor technique, **Fig. 12**, can be implemented with six mosfets or igbts switched at a high frequency.

Although this technique may be acceptable in some installations, it has safety limitations where people can touch the equipment being supplied. Even with a reliable earth leakage relay and the usual output over-voltage crowbar there are some people such as those with a pre-existing heart problem are at risk if exposed to a short duration mains shock while the earth leakage relay trips.

Using three batteries and three relays, **Fig. 13**, the output can be isolated to several kilovolts without a transformer,* with uninterruptible power supply as a bonus. With a 30V intermediate bus, three 30V batteries can be rotated in sequence, one being charged from the mains converter while the second supplies the load. The third ensures no-break changeover. A cycle life of around a second allows small batteries to be used, while giving an acceptable relay life.

This can match transformer isolation given suitable relays and sufficient tracking clearance of circuit board tracks. The batteries and their connections need to be suitably isolated from each other and the output side. Where the relays are exposed to humidity etc they must be protected or sealed.

In summary

Hopefully I have demonstrated that it is not impracticable to use switched capacitor dividers to convert mains to 3V without a transformer – and do it efficiently without huge amounts of silicon.

Certainly the number of components is greater than a conventional switch-mode power supplies, but when power factor correction and distributed converters are added to a switch-mode psu, the component difference reduces. ■

Anyone interested in communicating with the author regarding the concepts described here can do so by contacting Electronics World's editorial offices.

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