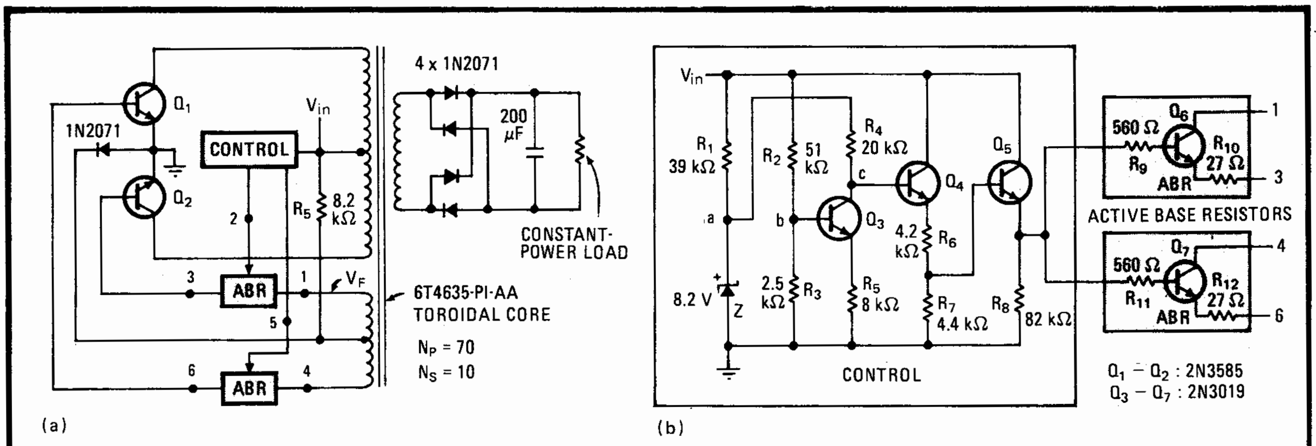
Dc-dc converter maintains high efficiency

by P. R. K. Chetty, *Department of Electrical Engineering, California Institute of Technology, Pasadena*

A simple control circuit enables this design to overcome the major drawback of the conventional dc-dc converter—its inability to maintain high efficiency over a wide range of input voltages. Varying the base drive to the converter's power-switching transistors as the inverse square of input voltage in order to achieve a near-constant ratio of output power to circuit losses, this unit attains efficiencies of 78% to 80% for $20 < V_{in} < 40$.



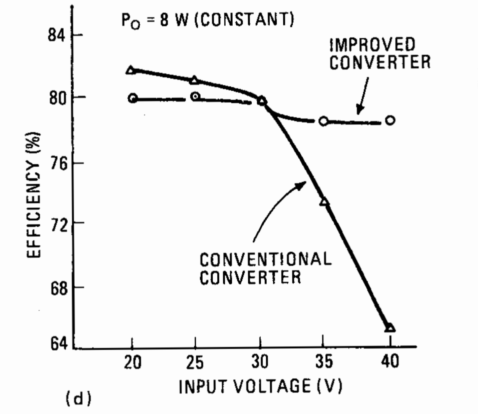
DESIGN EQUATIONS

$V_a = 8.2$ nom for $n \approx 80\%$
$V_b = (R_3/R_2 + R_3)\, V_{in}$
$V_{E_{Q3}} = V_b - V_{BE_{Q3}}$
$I_{E_{Q3}} = V_{E_{Q3}}/R_5$
$V_c = V_a - R_4 I_{E_{Q3}}$

$V_{E_{Q4}} = V_c - V_{BE_{Q4}}$
$V_{B_{Q5}} = (R_7/R_6 + R_7)V_{E_{Q4}}$
$V_{E_{Q5}} = V_{B_{Q5}} - V_{BE_{Q5}}$
$V_{E_{Q3}} = (R_3/R_2 + R_3)V_{in} - V_{BE_{Q3}}$
$I_{E_{Q3}} = V_{in}/R_5(R_3/R_2 + R_3) - V_{BE_{Q3}}/R_5$

$$V_c = V_a - V_{in}(R_4/R_5)(R_3/R_2 + R_3) + (R_4/R_5)V_{BE_{Q3}}$$
$$V_{E_{Q4}} = V_a - V_{in}(R_4/R_5)(R_2/R_2 + R_3) + (R_4/R_5)V_{BE_{Q3}} - V_{BE_{Q4}}$$
$$V_{BE_{Q5}} = (R_7/R_6 + R_7)[V_a - V_{in}(R_4/R_5)(R_3/R_2 + R_3) + (R_4/R_5)V_{BE_{Q3}} - V_{BE_{Q4}}]$$
$$V_{E_{Q5}} = (R_7/R_6 + R_7)[V_a - V_{in}(R_4/R_5)(R_3/R_2 + R_3) + (R_4/R_5)V_{BE_{Q3}} - V_{E_{Q4}}] - V_{BE_{Q5}}$$

(c)



Leveled. High efficiency of conventional dc-dc converter (a) is maintained with control circuit (b) that generates base drive to switching transistors Q_1–Q_2 in inverse proportion to input voltage. Equations (c) aid design. Typical performance (d) is plotted.

In the ordinary converter (a), which would not include the control block shown and where a fixed resistor, R_b, would be substituted for the active-base-resistor block, an increase in V_{in} causes efficiency, n, to drop off as the inverse of the square of input voltage. As may be seen, this loss results because:

$$R_b = V_F/2I_b = [V_F\, n(V_{in} - V_{ce\,sat})hfe_{min}]/2P_o$$

and $V_F = K\, V_{in}$, where R_b is selected to drop half of the feedback voltage, V_F, P_O is the desired output power, $V_{cc(sat)}$ and hfe_{min} are the collector-to-emitter drop and current gain, respectively, of either power transistor, and k is a constant dependent on the turns ratio. Thus, combining the two equations above, it is realized that n is approximately equal to $1/V_{in}^2$, keeping other variables constant.

It can be further shown that if R_b is made to vary as approximately V_{in}^2, the efficiency will be a maximum at any given input voltage. Equivalently, efficiency will be maximum if the base drive to the switching transistors Q_1–Q_2 is made inversely proportional to V_{in}.

Although the circuit required to exactly satisfy Eq. 1 would be complex, a relatively simple configuration (b) will provide acceptable performance when it is added to the basic converter. Here, a three-transistor controller (component values given for P_o = 8 W) and two active-base-resistor networks drive Q_1–Q_2.

As V_{in} increases, the voltage at point b increases. R_2 and R_3 are selected so that V_b is about 1 volt at $V_{in\,min}$, enabling Q_3 to operate in active region.

Because the collector of Q_3 is biased from a reference (point a), the drive signal applied to Q_4 is a function only of the voltage applied to Q_3's base. When the voltage at point B increases, the potential at point C decreases. Thus Q_4–Q_5, biased in its nonlinear i_b–e_c region, drives switching transistors Q_1–Q_2 through Q_6–Q_7 with less base current. As a result, the resistance between points 1–3 (and 4–6) will vary approximately as V_{in}^2.

Only one operating variable must be determined empirically, the voltage at the base of Q_5, $V_{B\,Q5}$. Breaking the circuit at this point to connect a variable-voltage source, the user sets V_{in} to its minimum expected value. The variable-voltage source is then set to saturate Q_1 and Q_2 for a constant P_o, and its value (V_{b1}) noted. The procedure is repeated to find V_{b2} for $V_{in\,max}$.

Now the design procedure may be initiated using the equations in (c) to determine R_6 and R_7, given that $V_{E\,Q5}$ equals V_{b1} at $V_{in\,min}$ and is equal to V_{b2} at $V_{in\,max}$. Experimental results for P_o = 8, $15 < V_{out} < 35$ and $20 < V_{in} < 40$ are tabulated (d) versus the performance of a conventional converter. □

Low voltage d.c.-to-d.c. converter

A CONVERTER operating from 2 or 3V is not easily made efficient because the voltage drop across the switching device in the on state can be a large fraction of the supply voltage. With a bipolar transistor as the switching device, V_{CEsat} can be made tolerable by providing a large base current, but this normally leads to extra loss in the base drive circuit. It is possible, however, to place the load in series with the transistor base-emitter junction so that the load acts as a high current base drive circuit and the transistor acts as a rectifier for load current. Both functions are performed without any extra power dissipation. The circuit shows this principle applied to an otherwise conventional self-oscillating converter, with the saving of two diodes. The usual separate base driver winding on the transformer is eliminated, which leaves space for thicker primary and secondary windings. Also, the base current is large and varies with loading. These two factors provide good regulation and high efficiency over a wide range of load currents.

An added advantage of this circuit is that the oscillation ceases if the load is open circuit, which reduces battery drain to a few microamps. The converter can therefore be controlled by a switch in series with the output. The prototype circuit, supplied from 2 nickel-cadmium cells, delivered 1A at 6V with an efficiency of 75 to 80%. The oscillation frequency was about 10kHz.

It is possible to modify the circuit by changing the transformer windings. If an audible whistle is undesirable, the frequency can be increased by reducing the number of primary turns. This will, however, reduce efficiency because commutation will take a larger fraction of the time, and the peak I_c will be higher. The output voltage can be raised by increasing the number of secondary turns. Above about 10V, diodes must be used to protect the base-emitter junctions against reverse breakdown.

A wide range of transistors can be used but they should be selected for low V_{CEsat} at the design current, and the BV_{EBO} must be adequate for the design output voltage.

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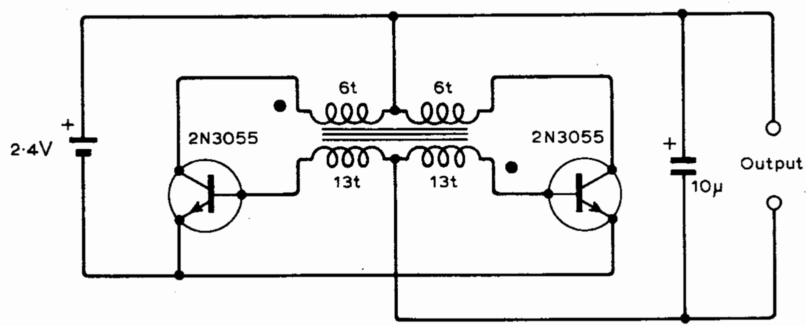


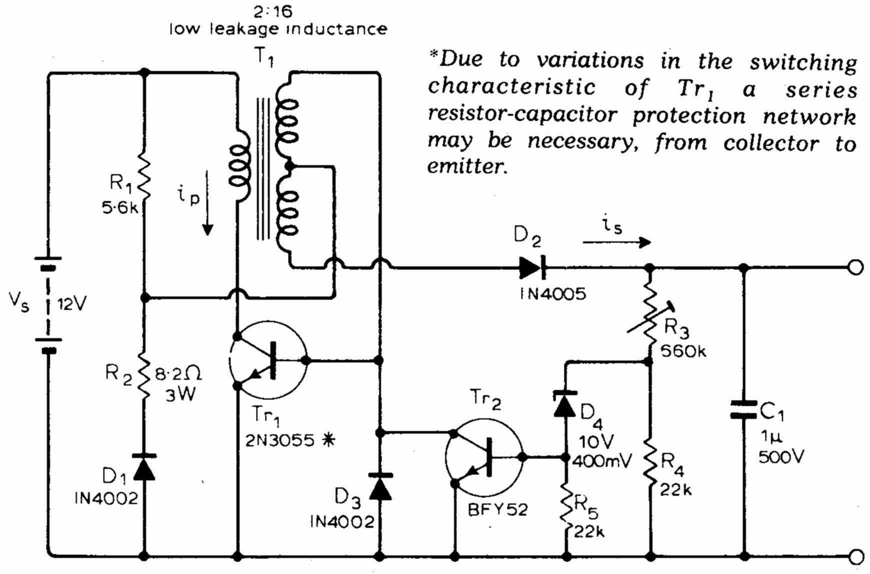
Transformer core: FX2239

Circuit Ideas

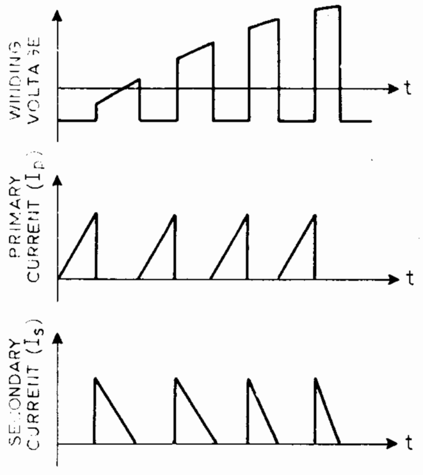
Variable voltage-ratio transistor converter

Conventional square wave inverter circuits give an output voltage proportional to the d.c. supply. The efficiency of such a system can never exceed 50% due to voltage limiting and component losses. The circuit described can charge a capacitor from zero volts at efficiencies of over 80%. When a conventional series stabilizer is used to drop the voltage of a 9V battery to 5V the average efficiency, taking the end point voltage of the battery as 6V, is only 67%. The use of this inverter extends a battery life in two ways: firstly by the higher efficiency and secondly by allowing the battery to be used to a lower end point voltage.

The primary current in T_1 is controlled by Tr_1 and oscillation is started by a low current passing through R_1 and the feedback winding of T_1 to the base of Tr_1. The transistor switches on due to the positive feedback action of the transformer and the main base drive is provided by the feedback winding through R_2 and D_1 While Tr_1 is conducting, D_2 is reverse biased and there is no current in the secondary winding. The collector current is therefore the sum of the referred base drive current and the



**Due to variations in the switching characteristic of Tr_1 a series resistor-capacitor protection network may be necessary, from collector to emitter.*



magnetizing current which rises linearly according to the equation

$$\frac{di_p}{dt} = \frac{V_{cc} - V_{ces}}{L_p}$$

where V_{ces} is the transistor saturation voltage, L_p the primary inductance of transformer and i_p the primary current. Thus the collector current rises until the base current is insufficient to saturate T_1, the positive feedback action then causes it to switch off. Reverse voltage on the transformer windings rises until D_2 and D_3 conduct and secondary current passes into the reservoir capacitor C_1. Diode D_1 is then reverse biased. Energy stored in the magnetic field of the transformer is transferred to C_1 and, when the current has dropped to zero, the winding voltages collapse and the oscillation is repeated until C_1 has charged up to a voltage which causes D_4 to conduct through the resistor R_3 and the base of Tr_2. This transistor then diverts the base current from Tr_1 and stops the oscillator until C_1 has discharged through the load circuit. Resistor R_5 carries the leakage current of D_4 and so prevents Tr_2 from conducting before the zener voltage has been reached. Waveforms are shown as C_1 is charged from zero. For fast switching it is essential that the leakage inductance between primary and feedback winding is very low. The leakage inductance between primary and secondary must also be low because the energy stored in the leakage flux cannot be transferred to the secondary when the transistor switches off. Air-gapped cores have been found most successful. There are two types of application where the circuit is of particular interest. For pulse generators where repetitive charging of a capacitor in a pulse-forming network is required or for generating one or more stabilized voltages from a dry battery. The circuit enables voltages above or below the battery voltage to be generated and the efficiency remains high throughout battery life. In the case of capacitor-discharge ignition systems the power conversion efficiency is so high that a heatsink is not required and only one power transistor is needed. The full output voltage is obtainable when the battery voltage is less than half its nominal value which results in an improvement in cold starting.

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Transformerless d.c. to d.c. converter

This converter operates by charging a bank of capacitors and connecting them in series periodically to give a higher voltage at the output. Transistors Tr_1 and Tr_2 form a multivibrator which produces a square wave. When the output of Tr_2 is positive C_1, C_2 and C_3 are charged through Tr_3, Tr_5, Tr_6, Tr_{10} and Tr_{11}. When the output of Tr_2 is zero, these transistors are cut off and Tr_4, Tr_7, Tr_8 are switched on, which connects the capacitors in series. In the prototype circuit the output voltage was 30V at 1mA although, using high current/voltage switches, an improved performance is possible. This system is also lighter in weight compared to transformer type converters.

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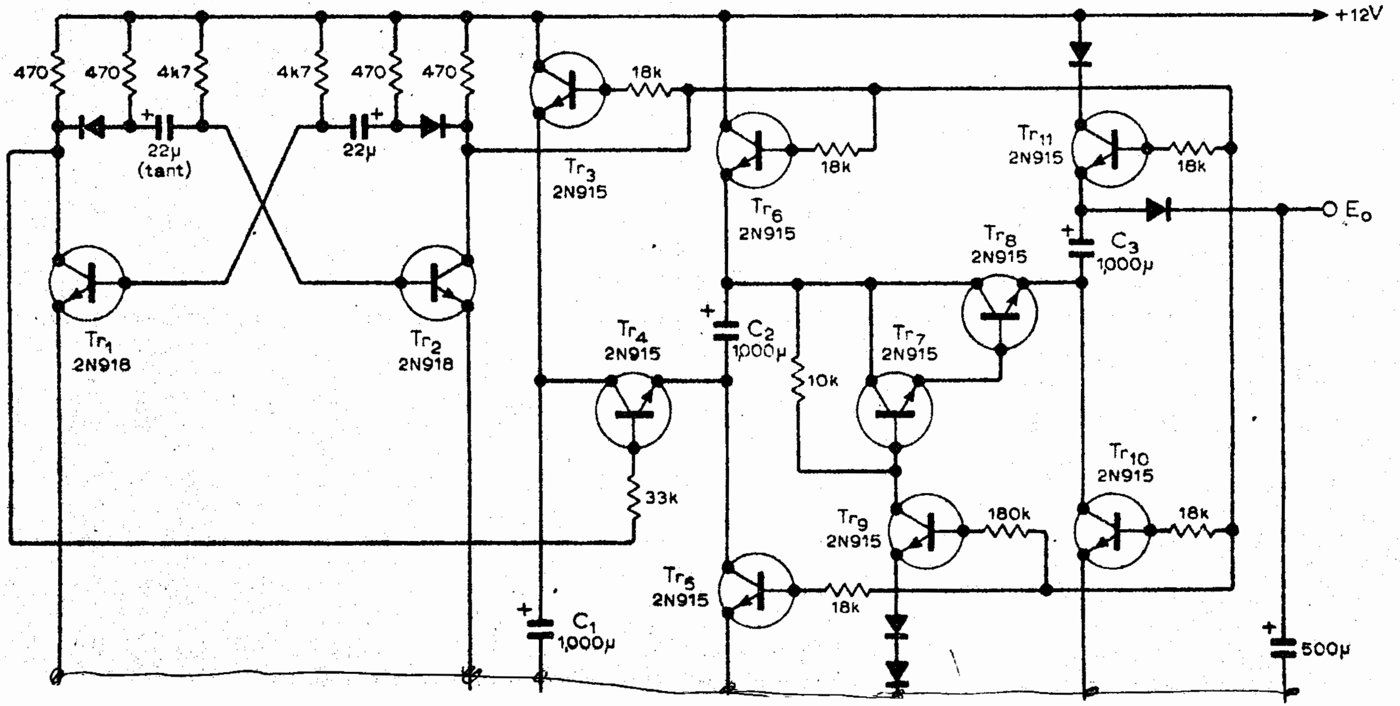

Proper filter choice eliminates oscillations in dc-dc converters

by Joseph Perkinson
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Designers often call upon a low-pass LC filter to reduce system noise entering a high-efficiency dc-dc converter. But often as not, they end up with an oscillator instead, if the converter has a negative input impedance and the filter's values are inappropriate. The oscillatory condition can, however, be avoided if the converter's input impedance is known, because then it becomes possible to select components for the filter that will prevent it from exchanging energy with the converter, in effect preventing the converter from turning into a negative-resistance voltage generator.

Dc-dc converters that have an efficiency of 70% or higher and that accept input-voltage variations of 2:1 or greater are the most vulnerable to oscillation when there is a negative input impedance. Generally, this happens if the current measured at the input of a converter decreases as the input voltage increases. Although this impedance can be determined graphically from data sheets, it is better to determine it analytically since this yields a greater understanding of the problem. It can be done if the converter's efficiency is known and, according to the data sheet, remains fairly independent of input-voltage variations.

Such an analytical solution is less time-consuming and involves less lab work than using data sheets to determine the relation between the converter's input resistance and the LC filter's constants. (Note that, at low frequencies, the terms input impedance and input resistance may be used interchangeably.) Shown in (a) of the figure is a typical input characteristic of a typical dc-dc converter. The curve is a hyperbola, meaning that:

$$V \cdot I = P_{in} = \text{constant}$$

or:

$$dV/dI = -P_{in}/I^2 = -V^2/P_{in} = R \quad (1)$$

Now assume that an LC filter is placed at the input of the converter. As shown in (b), the network the converter sees when looking back from its input is a parallel-resonant LC filter having an impedance at resonance of:

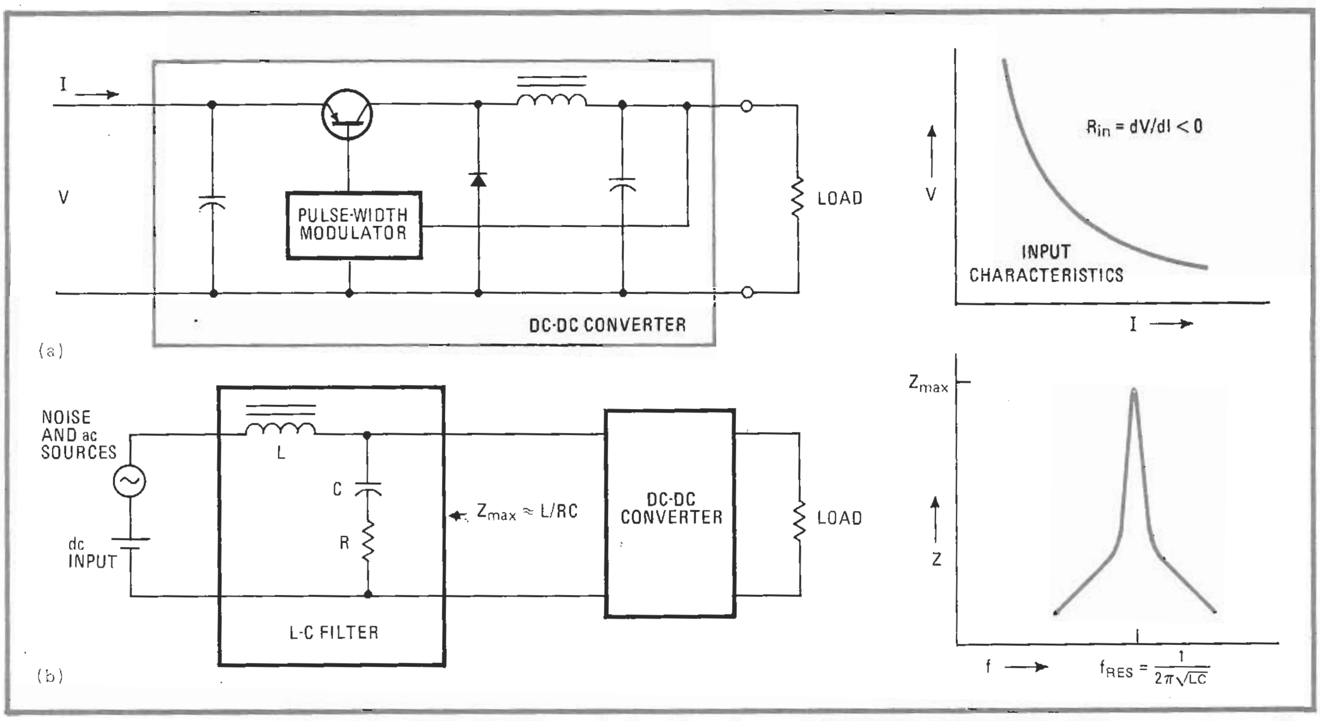
$$|Z_{max}| = L/CR \quad (2)$$

where R is the equivalent series resistance of the capacitor and $|Z_{max}|$ represents the impedance of an undamped (medium Q) circuit.

The converter-filter impedance of the (parallel) is thus expressed by:

$$R_T = 1/(1/Z + 1/R_{in}) \quad (3)$$

Oscillation results when $R_T < 0$. In other words, unless $L/CR < R_{in}$, the converter-filter combination will oscil-



Filter interaction. Switching dc-dc converter (a) exhibits negative input impedance, causes converter/noise-filter combination (b) to oscillate at f_{res} if $Z_{max} > R_{in}$. For proper filtering, and to eliminate the possibility of oscillations, filter's L-to-C ratio must be reduced.

late at the resonant frequency of the filter, and this condition must be avoided.

An example will underscore the usefulness of Eq. 3. Assume that a dc-dc converter with an input power of 50 watts and an input voltage range of 10 to 40 volts has a negative input resistance. Also, assume that the dc power-distribution system limits the input ac currents to the converter to less than 20 milliamperes peak to peak at 20 kilohertz, but that the actual input current waveform has a 5-ampere peak amplitude at 20 kHz.

If the LC filter values are arbitrarily selected for L = 75 microhenries and C = 470 microfarads, the filter cutoff frequency will be about 850 hertz, which offhand would seem a suitable value for amplitude and noise reduction. But from Eq. 1, given V = 10:

$$R_{in} = -10^2/50 = -2 \text{ ohms}$$

and from Eq. 2:

$$|Z_{max}| = 5.33\Omega$$

for R = 30 mΩ. Then $5.33 > |R_{in}| = 2$, and oscillation occurs. If the converter's input characteristic cannot be altered, then other values for L and C must be selected independently of the resonant frequency. □

Modulating the flyback inverter reduces supply's bulk

by Vladimir Brunstein
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The cost, size, and weight of an inverter will be reduced if a flyback transformer, modulated by a high-frequency carrier, is configured in the inverter's output stage. Modulating the flyback transformer is more efficient than attempting to modulate push-pull or bridge circuits, as is sometimes done.

The two subassemblies that are most difficult to miniaturize in the standard inverter are the output filter and transformer. But while the output filter can easily be simplified in a number of ways, the dimensions of the power transformer are determined mainly by the operating frequency.

The obvious solution to reducing the volume and weight of the transformer and the number of filter components, then, is to convert the dc input to a high frequency, such as 20 kilohertz. Here the harmonics of the output signal will be significantly higher than the demodulated frequency, and so a simple low-pass filter can be used to recover the desired 60-Hz waveform.

Using the flyback scheme brings an additional simplification, compared to a push-pull or bridge configuration. The recovered voltage on the secondary of the transformer will be amplitude-modulated, and only a capacitor will be required to obtain the low-frequency component.

As shown, a 60-Hz sinewave voltage is required to drive the flyback transformer via an error amplifier and pulse modulator and also through a power-driver connected to Q_2. Details of these blocks vary with individual requirements and so are not shown in detail here. Picking the right ferrite core is a subject in itself and has been discussed in various papers.

In general, coils T_{2a} and T_{2b} charge during the time interval t_1, and discharge through the load resistance R_L during t_2. Note that an inexpensive optical coupler may be used to replace T_{2b}. The voltage on the transformer secondary will be:

$$V_L = \eta \frac{n_S}{n_p} \frac{\tau}{1-\tau} E \qquad (1)$$

where:
η = efficiency of the power stage
$\tau = t_1/T$
t_1 = turn-on time for transistor Q_1
T = commutation period $t_1 + t_2$
E = dc input voltage

It is seen that the equation lends itself to achieving the end solution, for if the duty cycle is varied by the sine wave while keeping the turn-off time of Q_1 constant, then the peak value of the secondary will follow the sinewave reference as shown in the timing diagram.

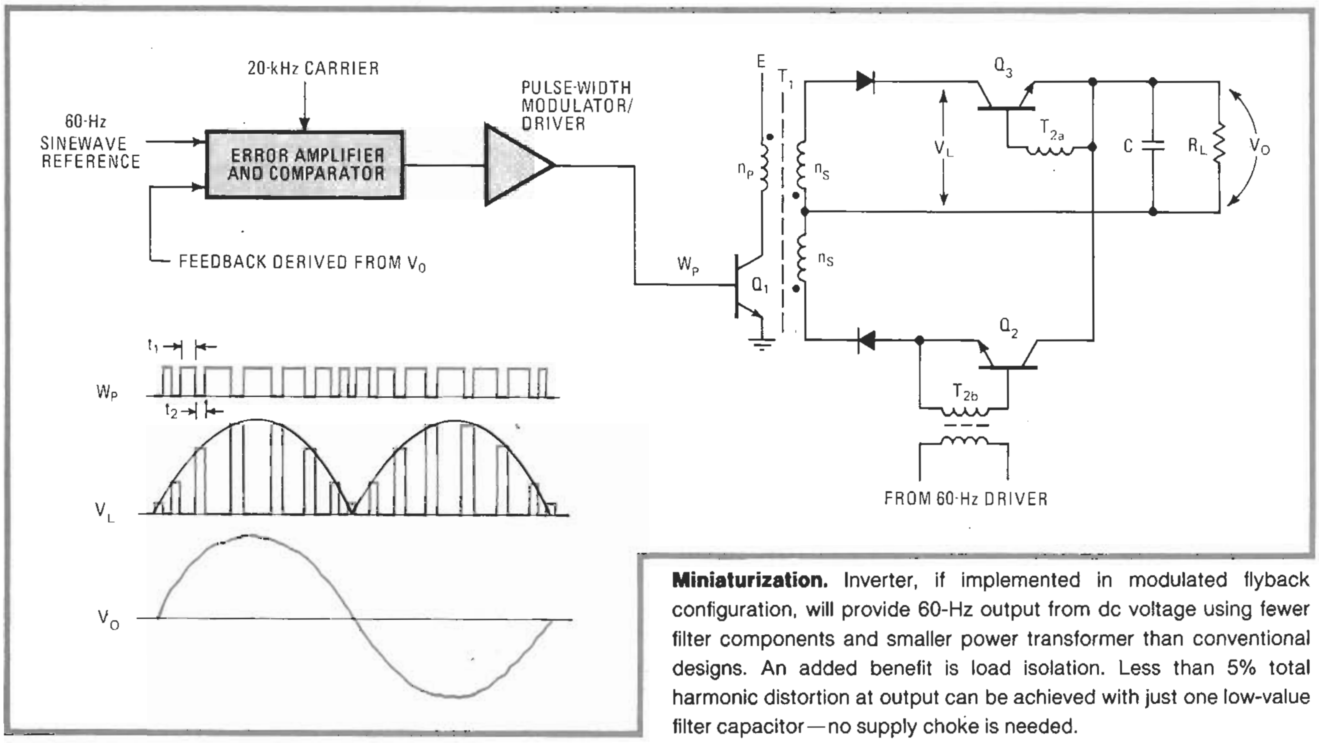
This action results if $t_1 = k_1 \sin \omega t$, for in that case Eq. 1 becomes:

$$V_L = K_2 \sin \omega t \qquad (2)$$

where K_1 and K_2 are constants.

In reality, one half of the sine wave would normally be inverted, as shown for V_L. Q_2 and Q_3 act as 60-Hz synchronous commutators serving to restore the original wave shape.

Note that only a relatively small output capacitor, C, is required for filtering—no choke is necessary. Total harmonic distortion at the output is less than 5%. □



Miniaturization. Inverter, if implemented in modulated flyback configuration, will provide 60-Hz output from dc voltage using fewer filter components and smaller power transformer than conventional designs. An added benefit is load isolation. Less than 5% total harmonic distortion at output can be achieved with just one low-value filter capacitor—no supply choke is needed.

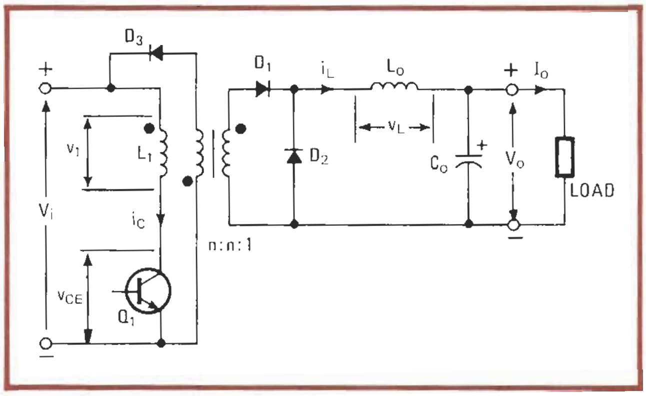
Low-cost forward converters ease switching supply design

Simplified circuitry built around novel transformer converts dc power up to kilowatt range

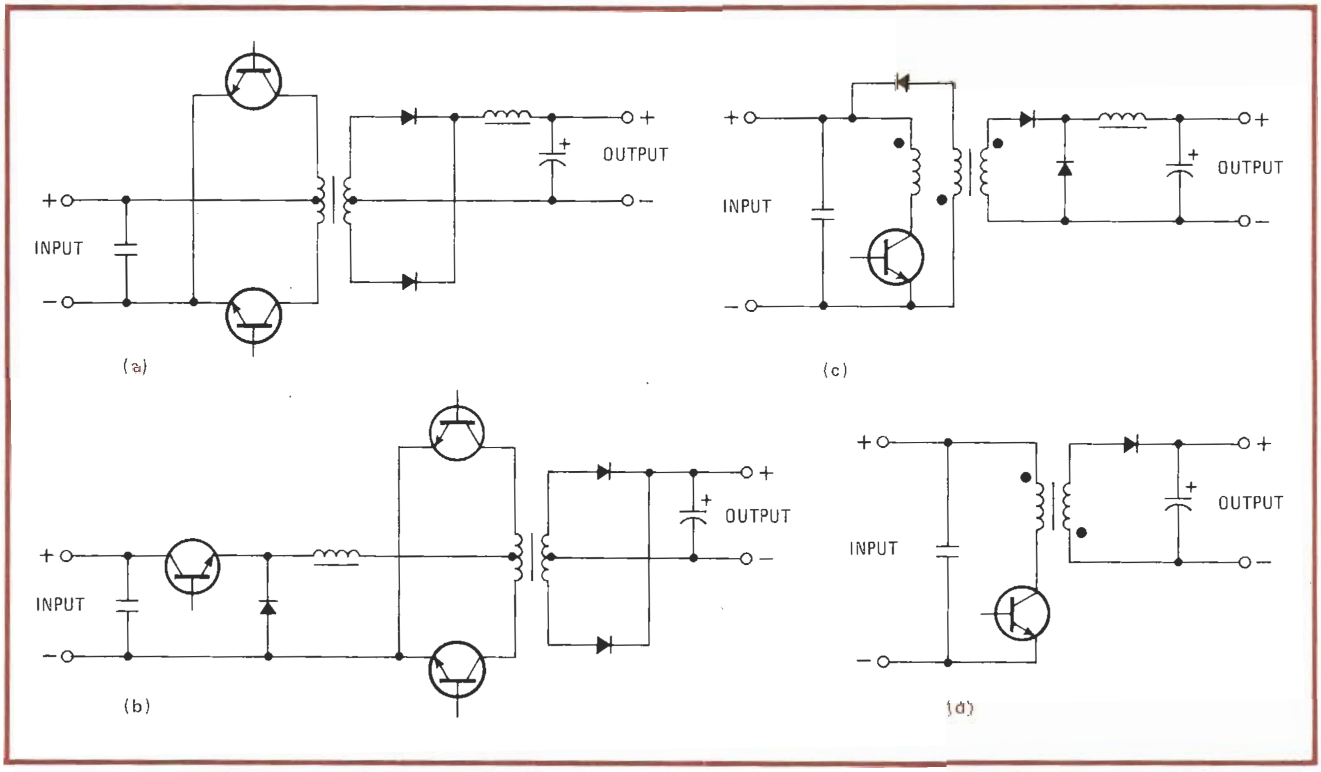
by Kees van Velthooven and Hugo Koppe, *N V Philips Semiconductor Laboratory, Nijmegen, the Netherlands*

☐ Switching power supplies are winning out over linear types because of their higher conversion efficiency, which also makes them smaller and cooler. And as the variable switching element of the supply, transformer-isolated, base-controlled dc-dc converters are winning out over series pass transistors. Transformer isolation of the input and output of course makes for safer operation, while use of a converter allows the output voltage to be greater than the input. In addition, such an approach reduces noise.

Now, the development of the forward and the double forward converter makes this approach even more attractive. These two circuits cost less than other transformer-isolated versions and are inherently easier to design. Furthermore, double forward converters, which have been built with output power in the kilowatt range, have a faster transient response and use smaller compo-



2. Forward converter. A special three-winding transformer is the key to this highly efficient one-transistor converter. A demagnetizing winding tightly coupled to the primary, along with diode D_3, prevents the transformer core from saturating when Q_1 is cut off.



1. Isolated. The push-pull converter (a), switching regulator followed by an unregulated converter (b), forward converter (c), and ringing-choke converter (d) isolate the input dc from the regulated output voltage. A forward converter resembles an isolated switching regulator.

nents than the other types.

Figure 1 shows the various transformer-isolated converter circuits. The push-pull converter (Fig. 1a) is the most widely used, although it suffers from the disadvantage of collector current peaking if its transformer saturates when there is a dc unbalance or a sudden rise in load. This drawback becomes even more serious with lower-loss core materials, which promote saturation-limited rather than loss-limited transformer design.

One approach to preventing collector current peaking employs a switching regulator combined with an unregulated converter (Fig. 1b). The switching regulator effectively limits any voltage surges that could unbalance the switching transformer. However, the circuit is complex and needs separate base drives for the regulating transistor and the unregulated transistor converter.

The forward converter (Fig. 1c), first designed in 1974, is now finding its way into power supplies. It is almost as simple as the low-efficiency, one-transistor ringing-choke supply (Fig. 1d), yet its ripple and its output capability are comparable to those of the push-pull converter. In addition, the forward converter does not have the push-pull converter's problem of dc unbalance in the transformer core, because transistor conduction occurs only once per converter cycle. Moreover, there is less flux peaking in its output transformer during load transients, and there is no interaction between the magnetizing and load currents.

Even more recent is the double forward converter. This type, which will be discussed later, is suitable for high-power applications.

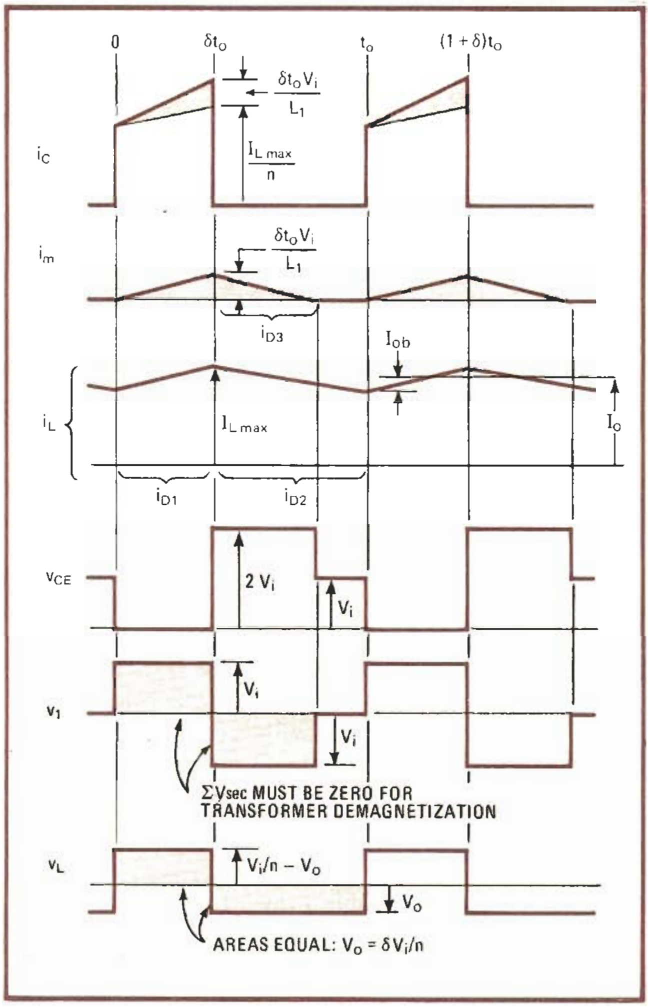
The table gives the behavior of the forward, double forward, and push-pull converters and the size of their components, assuming identical outputs.

The forward converter

The forward converter, whose schematic and waveforms are shown in Figs. 2 and 3 respectively, needs only one transistor for load switching. This is a tremendous advantage in view of the high cost of high-voltage switching transistors.

When Q_1 is switched on by a positive pulse at its base, rectifier D_1 starts to conduct, and energy is passed to output choke L_o and the load. During this stage, choke current i_L is rising. At the same time, magnetizing current begins to build up in the transformer primary. When the pulse at the base of Q_1 goes to zero, Q_1 is switched off, i_L falls, and part of the energy stored in L_o is transferred to the load through flywheel diode D_2. Meanwhile, magnetizing current continues to flow through the demagnetizing winding and D_3, a fast soft-recovery diode that doubles as a collector voltage clamp. Because choke current flows continuously, output ripple is low.

Before Q_1 is switched on again, the magnetizing current must have reached zero, or transformer saturation will occur; that is, with the primary-to-demagnetizing-winding turns ratio assumed to be unity, the transistor duty cycle, δ (the on time of the transistor divided by the cycle time of the converter), must not exceed 0.5. As the V_{CE} waveform shows (Fig. 3), the peak collector voltage is $2V_i$ (ignoring some additional voltage due to



3. Forward-converter waveforms. These waveforms are valid for ideally coupled windings and zero transistor switching times. Magnetizing current is shown cross-hatched. Peak collector voltage is $2V_i$ for primary-to-demagnetizing-winding turns ratio of unity.

ringing). To ensure smooth transfer of the magnetizing current, the primary and demagnetizing windings must be tightly coupled.

In comparison, in push-pull converter transformers the magnetizing current flows alternately in the primary and the secondary. Tight coupling between these windings leads to complex split-winding constructions owing to the necessity for interleaving and isolating windings. Forward-converter transformer design, on the other hand, is relatively simple because tight coupling between the primary and secondary is not necessary.

As seen from the V_L waveform (Fig. 3), the converter's dc output voltage is:

$$V_o = \frac{\delta V_i}{n}$$

where n is the primary-to-secondary turns ratio of the transformer. This relation holds for an uninterrupted choke current. At light load, an interrupted choke current occurs and V_o tends to rise, as is the case with all other converter types.

The boundary value of output current below which the choke current becomes interrupted is:

$$I_{ob} = \frac{V_i t_o \delta}{nL_o 2}(1-\delta)$$

where t_o is the converter cycle time. Below this current level, a large change in δ must occur to maintain regulation. This requires a high open-loop gain from the regulating circuit that controls the base drive of the switching transistor. The value of I_{ob} (which in turn determines the inductance of the output choke) is, therefore, an important design parameter. Normally, I_{ob} is 0.05 to 0.1 times the rated load current.

As with any converter type, the base-drive circuit for the switching transistor must be carefully designed to minimize conduction and switching losses.

To repeat, the merits of the forward converter are:

- Better transistor utilization—because energy still flows into the load while the transistor is off—and lower output ripple than with the ringing-choke converter.
- Simpler circuitry and easier transformer design than with the push-pull converter.

Output ripple

In most switching supplies, high-value output capacitors are used to reduce to acceptable levels output voltage transients caused by abrupt load changes. Therefore the output ripple depends almost entirely on the equivalent capacitor series resistance, ESR, and inductance, ESL, of the filter capacitance, rather than on the value of the filter capacitance itself. Figure 4 shows an equivalent circuit of an output filter for a switching regulator.

For the forward converter, the peak-to-peak output ripple, $V_{r(pp)}$, divided by the output voltage is:

$$\frac{V_{r(pp)}}{V_o} = \frac{ESL}{\delta L_o} + \frac{ESR(1-\delta)t_o}{L_o}$$

In comparison, the ripple equation for the push-pull converter is:

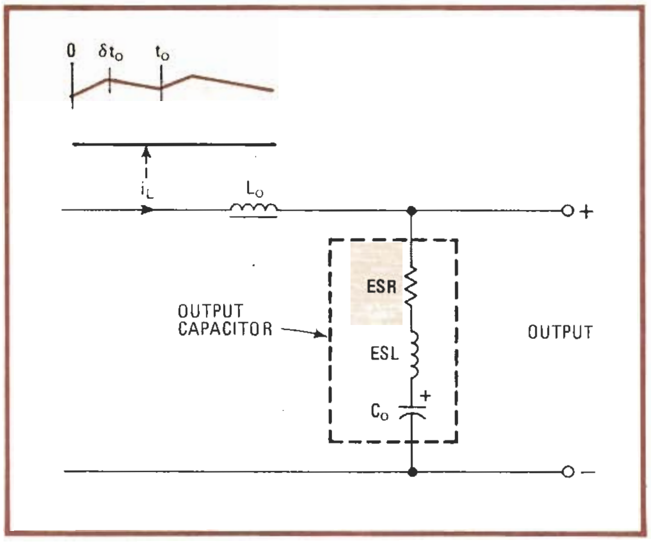
$$\frac{V_{r(pp)}}{V_o} = \frac{ESL}{2\delta L_o} + \frac{ESR(1-2\delta)t_o}{2L_o}$$

If identical output capacitors are taken for both cases, and if the output voltage and power are assumed to be equal, the ripple voltages will not greatly differ. With present-day, low-impedance electrolyte capacitors, a ripple of less than 1% is readily attainable for both converter types.

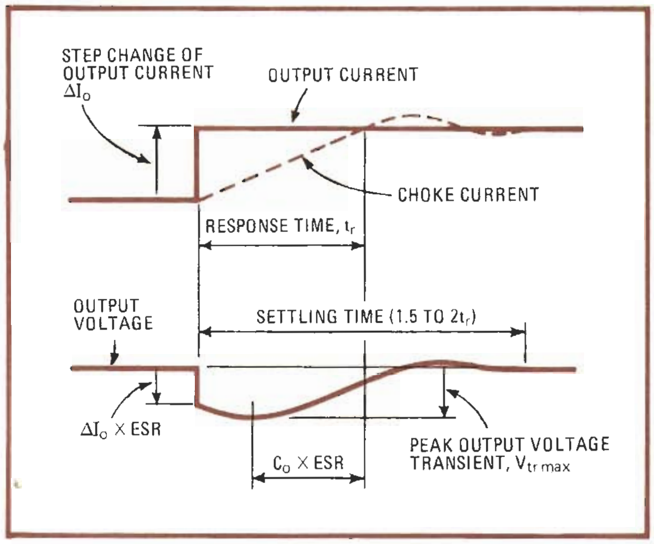
Switching power supplies respond much more slowly to load transients than linear supplies because the current through the choke in series with the output (Fig. 4) cannot change instantaneously. The difference between the choke and load currents must flow through the output capacitor, generating the voltage transient shown in Fig. 5. Output voltage settling time is 1.5 to 2 times the response time, t_r, which is the time required for the choke current to adjust to the new value of the load current. The peak output voltage transient, $V_{tr\,max}$, depends on the values of C_o and ESR, its minimum value being equal to $\Delta I_o \times ESR$. High-value, low-impedance capacitors are therefore essential.

The magnetizing current in the output transformer of the forward converter flows in only one direction (see the i_m waveform in Fig. 3), in contrast to that in the push-pull converter, and thus its core is unilaterally rather than bilaterally magnetized. With the low-loss ferrite cores now available, transformer design is based mostly on the magnetic flux permitted in the core.

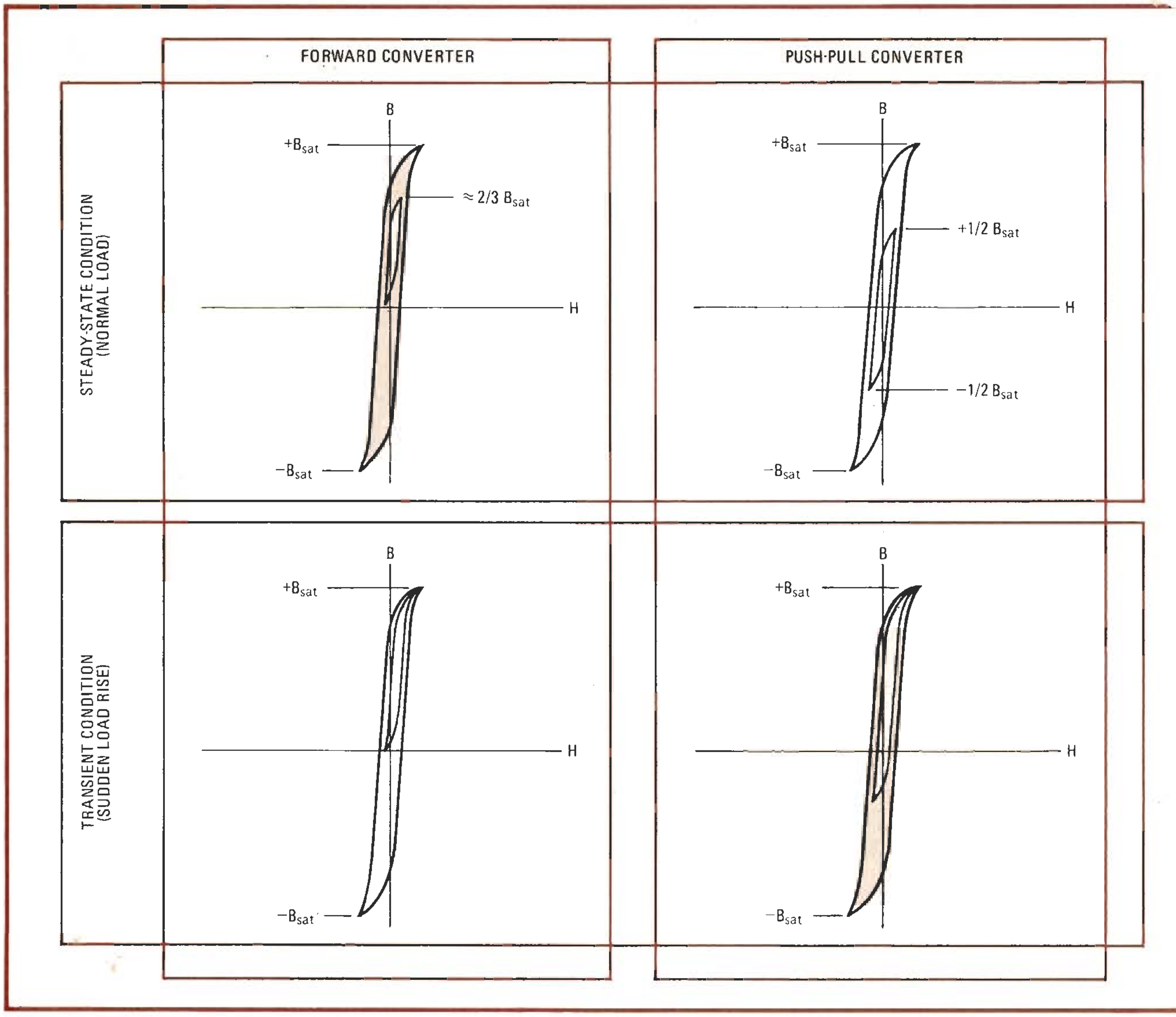
If the push-pull converter's operation could be perfectly balanced, then its core would work close to saturation and its size could be half that of the forward converter. But because of flux peaking during transients (Fig. 6) and the possibility of unbalanced operation, no great reduction in core size is obtained. However, although it can be shown that the peak flux increase due to transients is approximately 100% in the push-pull converter, compared with 50% in the forward converter,



4. Output ripple. Shown is the equivalent ac circuit of the output filter for either a forward or a push-pull converter. Since the output capacitor must be large to meet output voltage transient specifications, output ripple depends almost entirely on the equivalent series resistance and inductance, ESR and ESI, respectively.



5. Transient behavior. In forward and push-pull converters, the current through the filter choke cannot instantly follow a step change in output current. The difference between choke and output currents flows into the output capacitor, causing a slowly decaying transient whose amplitude is $\Delta I_o \times ESR$ or larger.



6. Hysteresis. In a forward converter, flux density, B, of the transformer swings between 0 and 2/3 B_{SAT}, unlike the case of a push-pull converter, where flux density operates between $\pm \frac{1}{2} B_{SAT}$. As a result, the forward converter is less susceptible to load transients than the push-pull converter, which can be driven into core saturation, causing collector current peaking.

COMPARISON OF FORWARD AND PUSH-PULL CONVERTERS (EQUAL OUTPUT VOLTAGE AND POWER)

	Forward converter	Double forward converter	Push-pull converter
Output voltage equation	$\delta V_i/n$	$2\delta V_i/n$	$2\delta V_i/n$
Transformer size	1	2 × 0.5	1
Choke size (equal boundary output currents)	1	0.3	0.6
Output capacitor size (equal output voltage transients)	1	0.3	0.6
Number of switching transistors	1	2*	2*
Number of output diodes	2	3†	2
Response time	1	0.3	0.6

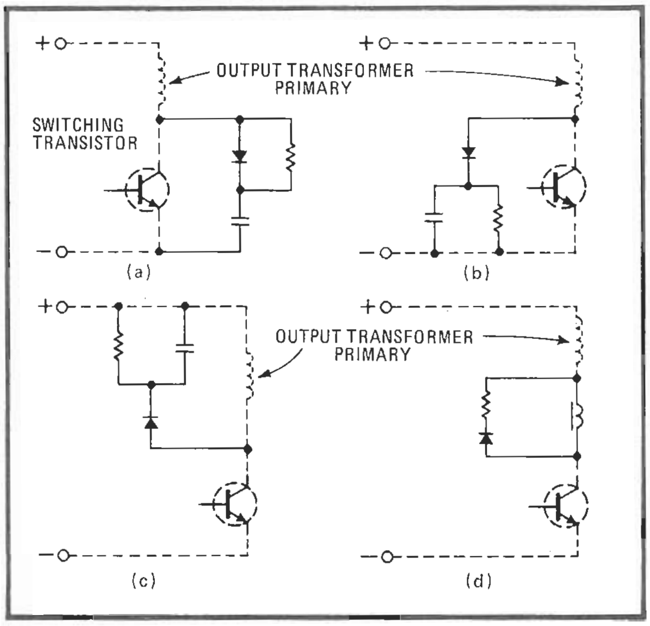
*Required collector current rating is half that for forward converter.
†Required flywheel diode reverse voltage rating is half that for forward converter.

transformer size is about the same for both types.

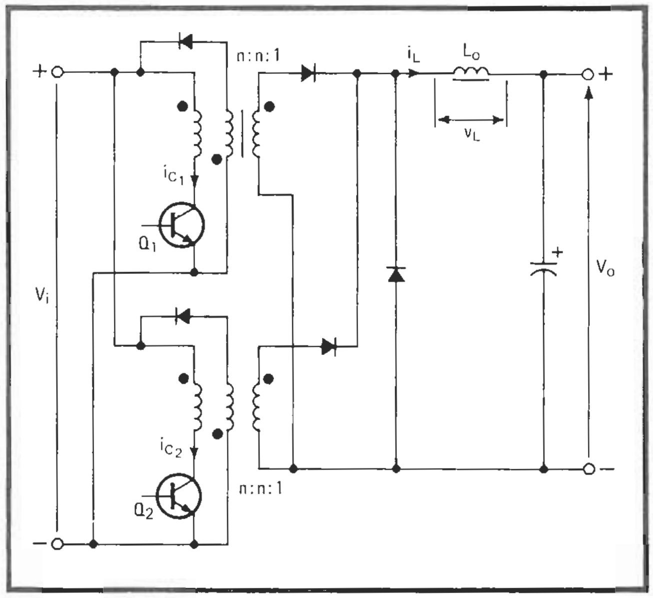
Since switching supplies run at high frequencies, fast output diodes are necessary to avoid large diode reverse-recovery currents, which could cause high diode turn-off power losses and high turn-on peak collector currents. Diode reverse-recovery times should be at most approximately 100 nanoseconds, and preferably less. Soft-recovery properties are necessary to minimize electromagnetic interference and diode turn-off voltage surges, both of which can jeopardize the diodes themselves.

Epitaxial diodes or, for low output voltages, Schottky barrier diodes should be used for switching supplies. The reverse-voltage rating required varies between $4V_o$ and $5V_o$, depending on output voltage, diode voltage drop, and ringing suppression.

Protection networks are indispensable for keeping a supply's switching transistor within its safe operating area. In push-pull converters, RC networks across the switching transistors or the transformer primary provide protection. A low-value resistor (to keep the network



7. Protection. Four protective networks used with forward converters to keep the power transistor within its safe operating area are dV_{CE}/dt limiting (a), peak-rectified clamping (b), clamping with reduced losses (c), and turn-on di_C/dt limiting (d).



8. Doubling up. The double forward converter consists of two forward converters switched alternately. The waveforms produced are similar to those of the single forward converter. However, output power and ripple frequency are doubled and transient response is improved over the one-transistor version.

time constant low) will limit the turn-on peak collector current to some extent, but high power losses will result because of the large capacitor charge and discharge currents. Forward converters, on the other hand, offer much greater freedom in network design, and the protection networks possible ensure safe turn-on and turn-off collector current with acceptable loss.

One of the most successful types, the dv_{CE}/dt limiter (Fig. 7a), ensures that collector current has dropped to the cut-off level before collector voltage exceeds the rated V_{CEO}. It does so by decreasing the collector voltage rise rate to between 400 and 800 volts per microsecond. For effective action, the capacitor should discharge almost fully during transistor conduction.

The peak-rectifier clamper (Fig. 7b, 7c) limits the peak collector voltage to about twice the converter dc input voltage by suppressing voltage surges due to inductive switching. A third kind, the di_C/dt limiting network (Fig. 7d), prevents turn-on dissipation. It is used where the transformer has low leakage inductance, which together with the collector current would otherwise rise abruptly because of the winding capacitance charge current and the output-diode reverse recovery.

The diodes in these protective networks are high-speed, soft-recovery types. They cause negligible surge voltage and electromagnetic interference.

Switching transistor choice is governed by the dc input voltage and the transformer throughput power. With any of the recommended networks, the required V_{CE} rating is twice the converter dc input voltage plus some additional overvoltage (approximately 50 v) for ringing. The peak collector current is the sum of the reflected peak choke current, which equals $(I_o + I_{ob})/n$, and the peak magnetizing current; depending on transformer size, the peak magnetizing current is 0.01 to 0.1 times the reflected peak choke current.

With present-day state-of-the-art components, forward converters can be built to handle up to 500-watt outputs. Higher power can be obtained by paralleling transistors or transistor-transformer pairs. Where the power requirement exceeds 1,000 w, the double forward converter (Fig. 8) becomes attractive. This circuit consists of two forward converters with a common choke, the converters being switched alternately by base drives to Q_1 and Q_2 that are 180° out of phase.

Doubling output power

Total output power of the double forward converter is twice that delivered by the individual converters. Because ripple frequency is doubled, lower values of filter capacitance and inductance are needed. Also, load transient response is much faster than that of the forward converter. These improvements, however, are bought at the cost of somewhat greater circuit complexity and a more intricate base drive. The latter remains simpler than that of push-pull circuits, since no measures are needed to counter unbalance.

Performance is similar to that of the push-pull converter, but with two advantages: there is no interaction between magnetizing current and output current at low output power, and the circuit is hardly sensitive to dc unbalance. As in the case of the push-pull converter, the output voltage is:

$$V_o = \frac{2\delta V_i}{n}$$

As in the case of the forward converter, a boundary value of load current exists:

$$I_{ob} = \frac{V_i t_o \delta}{n L_o 2}(1 - 2\delta)$$

below which output regulation will worsen. □

14

two-TUN voltage doubler

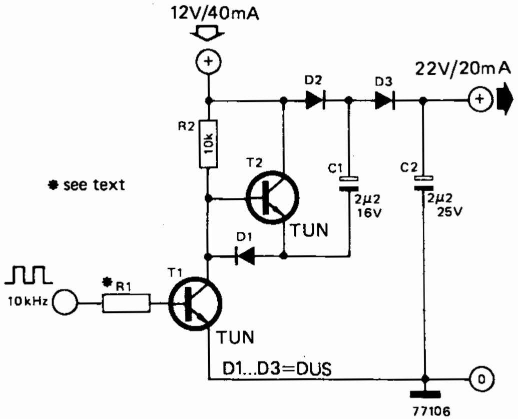
elektor july/august 1977

This little circuit will produce a DC output that is almost twice the supply voltage. A square-wave input is required of sufficient level to turn T1 fully on and off. When T1 is conducting, C2 is charged to just under the supply voltage. When T1 is cut off, T2 starts to conduct and raises the voltage at the negative end of C2 to just under the positive supply level. This implies that the voltage at the positive end of C2 is raised to almost twice the supply voltage, so that C3 will ultimately charge to this level.

The circuit is remarkably efficient: the current drawn from the main supply is only marginally greater than twice the output current. In the example shown here, the efficiency is approximately 90%.

The value for R1 depends on the amplitude of the square-wave input: T1 will require a base current of 0.5 . . . 1 mA.

(RCA application note)



On-chip transistors extend audio amp's design flexibility

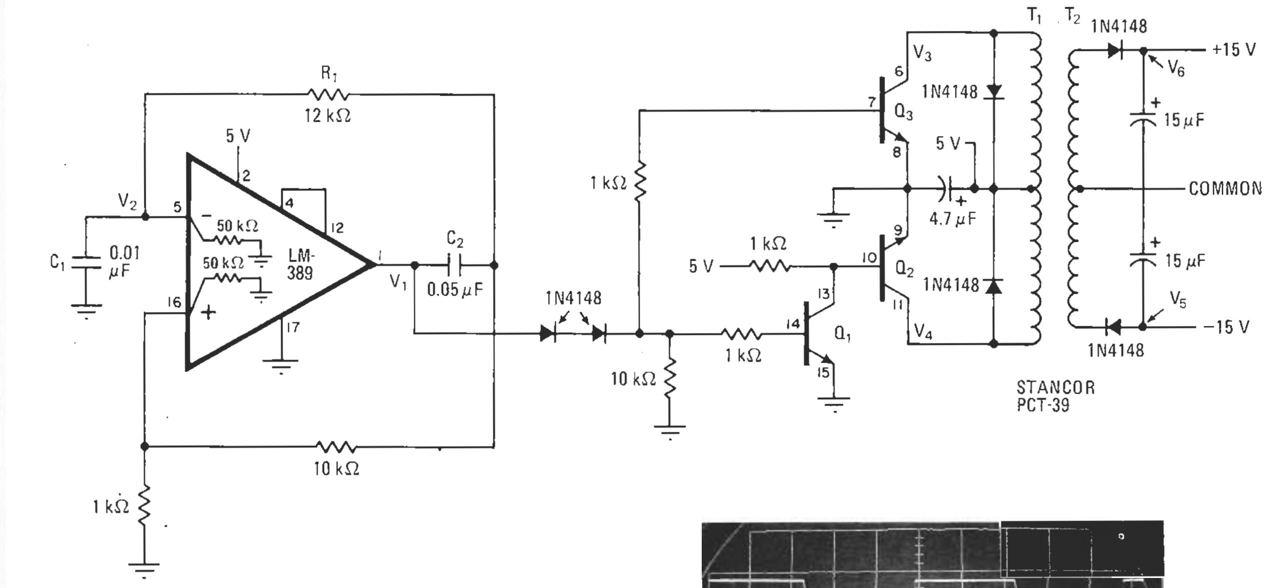
by Jim Williams
National Semiconductor Corp., Santa Clara, Calif.

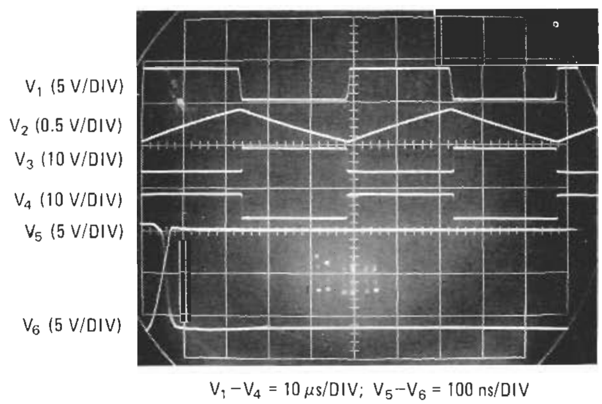
The availability of extremely low-cost audio-amplifier integrated circuits with on-chip transistor arrays, such as National's LM389, gives designers a great deal of flexibility in designing audio circuits. They make it much easier to develop low-cost versions of circuits unrelated to basic audio amplification, such as dc-dc converters, touch switches, stabilized frequency standards, scope calibrators, low-distortion oscillators, and logarithmic amplifiers. The designs of the often-needed converter, a bistable touch switch, and a tuning-fork frequency standard are discussed here in the first part of this two-part presentation.

The LM389 contains a 250-milliwatt audio amplifier and an array of three npn transistors, each of which is uncommitted. The amp has differential inputs and separate pins for setting its gain (from 20 to 200) via a resistor and runs off a single supply that may range from 4 to 15 volts. The three transistors have a minimum current-handling capability of 25 milliamperes and a minimum current gain of 100 for $V_{ce\ max}$ = 12 V and for a wide range of collector currents. The chip is therefore ideal for general use.

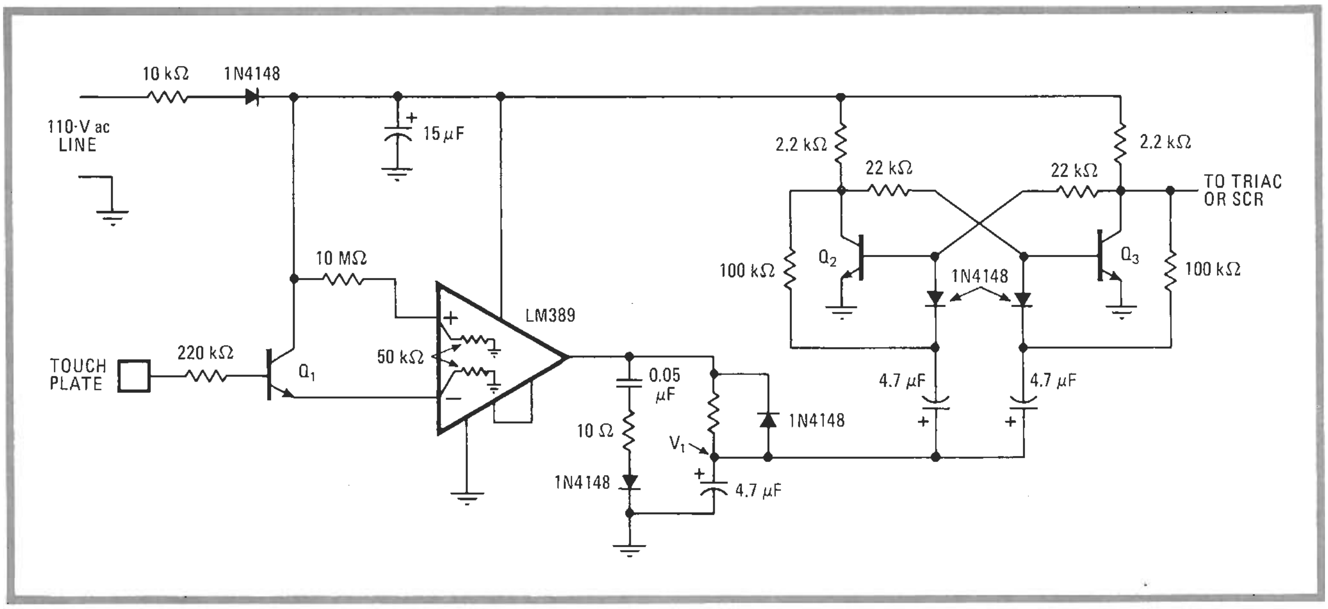
One area in which the chip will be useful is in dc-dc switching conversion. The device in Fig. 1 is intended for use as a power supply in a digital system where it is necessary to supply ±15 V to a low-power load. As can be seen from the oscilloscope photograph, the LM389 switches at 20 kilohertz. That rate is determined by the triangular-wave feedback signal, whose time constant is set by $R_1C_1C_2$, and its square-wave output is applied to transistors Q_1 and Q_3. The series diodes ensure clean turn-off for Q_1 and Q_3.

Q_1's inverted output drives one half of the transformer primary through Q_2, while Q_3 drives the other half. The diodes across Q_2 and Q_3 suppress spikes. Thus there is an

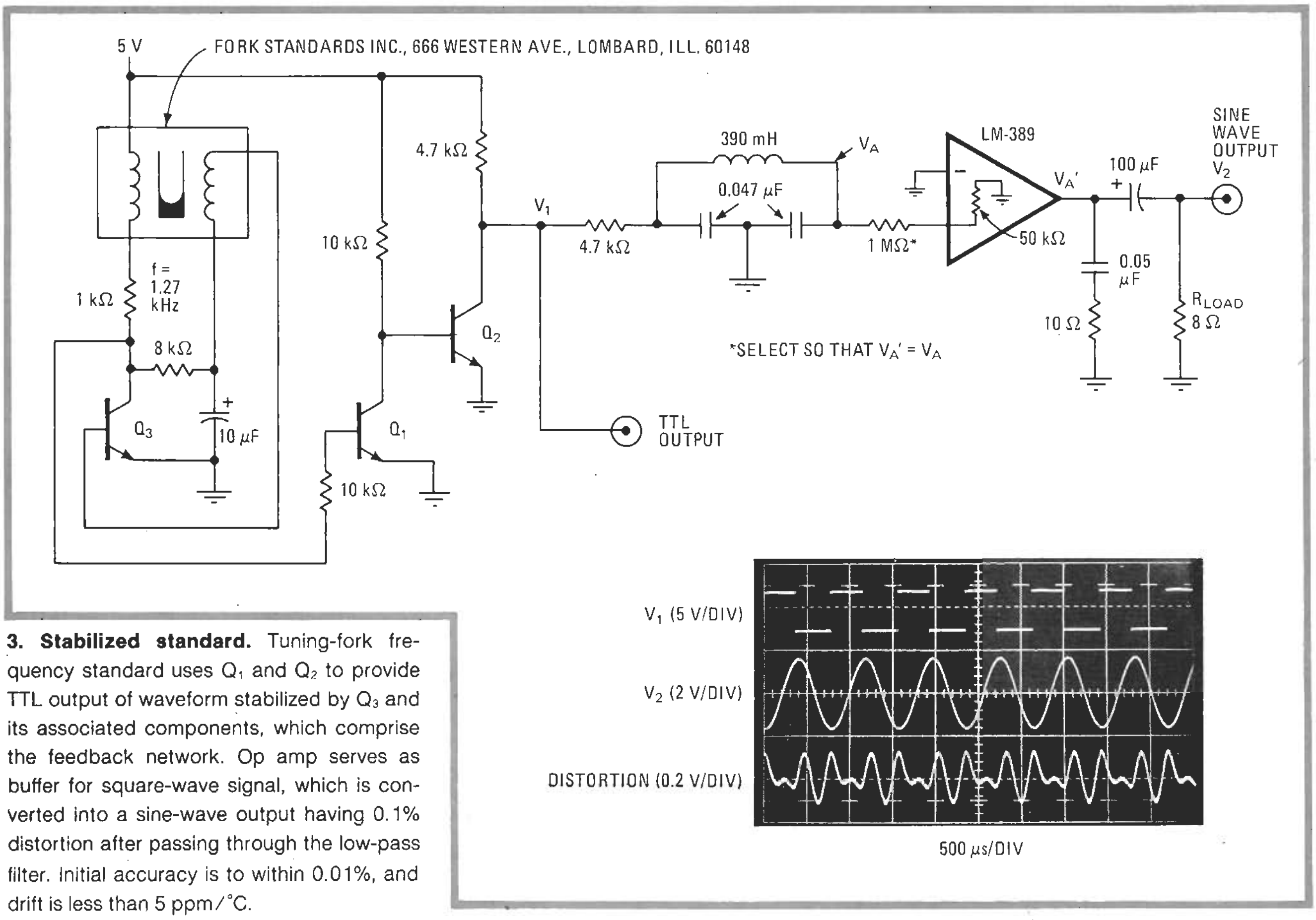




1. Condensed converter. Op amps with an on-chip transistor array, such as National's LM389, make inexpensive one-chip switching converters. Costing under $10, unit shown provides ±1.5 mA at ±15 V for instrumentation amplifier. Scope photo displays waveforms at various points in circuit. Use of diodes in driver side of circuit allows spike suppression, plus clean switching.



2. Simple switching. Simple bistable touch switch may be similarly constructed. Op amp works as comparator and trigger for flip-flop Q_2–Q_3, which changes state each time plate is contacted. Thus, SCR at output may be alternately fired and switched off on command.



3. Stabilized standard. Tuning-fork frequency standard uses Q_1 and Q_2 to provide TTL output of waveform stabilized by Q_3 and its associated components, which comprise the feedback network. Op amp serves as buffer for square-wave signal, which is converted into a sine-wave output having 0.1% distortion after passing through the low-pass filter. Initial accuracy is to within 0.01%, and drift is less than 5 ppm/°C.

efficient step-up of voltage across T_2. This ac signal is rectified and filtered to produce complementary output voltages that may be used to power the desired linear components, in this case delivering ±1.5 mA, enough to power an operational or instrumentation amplifier.

The bistable touch switch (Fig. 2) allows a line-powered load to be controlled from a touch plate by means of a thyristor. Each time the plate is contacted, emitter-follower Q_1 conducts, permitting a fraction of the 60-hertz input signal to be applied to the inverting input of the amplifier. Consequently, the normally high output of the op amp follows the 60-Hz line input, causing V_1 to drop sharply.

This negative transition triggers a toggling flip-flop formed by Q_2 and Q_3. In this manner, the output of the flip-flop changes state each time the touch plate is contacted, prompting the firing of the silicon controlled rectifier or triac that switches ac power to the load.

Figure 3 shows a tuning-fork frequency standard that is stabilized by appropriate feedback. Both sine-wave and TTL-compatible outputs are available. As the circuit needs only 5 V, it can run off a battery.

The tuning fork proper supplies a low-frequency output that is very stable (typically to within 5 ppm/°C) and has an initial accuracy of within 0.01%. Moreover, it will withstand vibration and shock that would fracture a quartz crystal. Here, Q_3 is set up in a feedback configuration that forces the fork to oscillate at its resonant frequency. Q_3's output is squared up by Q_1 and Q_2, which provide a TTL-compatible output. When passed through an LC filter and the op amp, which provides a low-impedance (8-ohm) output, the signal is converted into a sine wave having less than 1% distortion, as shown in the figure.

Several other useful circuits also can be built. The second part of this article will deal with the chip's use in a portable scope calibrator, a low-distortion oscillator, and as a logarithmic amplifier. □

BUILD A TRANSFORMERLESS DC-TO-DC VOLTAGE DOUBLER

High-efficiency, high-current solid-state circuit doubles the input dc level.

BY MARLOWE J. BUCHANAN

FOR MANY years, the only reasonable nonmechanical means of generating high dc potentials from low voltages in the medium-to-high-power range has been the oscillator-driven transformer. This type of converter has two major drawbacks—the average experimenter may not have the knowledge or the materials to design and build his own transformer/converters, and the power consumption is relatively high, making the converters inefficient at currents much below their ratings.

The capacitive voltage doubler described here is superior to normal converters in many applications. For example, it can be used to extend the range of a low-voltage power supply or to run a medium-power audio system in a vehicle. Capable of very high efficiencies, the converter can easily be adapted to the voltage and power needed.

Circuit Operation. Basically, the circuit shown in Fig. 1 acts as a set of high-speed electronic switches that alternately charge *C3* and *C4* to the supply voltage and then connect the capacitors in series with the supply and load. The output load in effect "sees" the sum of the voltages across *C3* and *C4* and the supply voltage. Since *C3* and *C4* are driven 180° out-of-phase, this is a true full-wave voltage doubler.

Transistors *Q7* and *Q8* and their associated components form a multivibrator that has an operating frequency of roughly 6000 Hz with a 12-volt dc input supply. This oscillator produces two square waves that are 180° out-of-phase with each other and are used to drive *Q5, Q6, Q9,* and *Q10.* Resistors *R1* through *R4* are selected to limit base current to 3 mA. Drive transistor pairs *Q5/Q9* and *Q6/Q10* produce equal and opposite square waves that have extremely fast rise and fall times and are capable of sinking or sourcing a minimum of 100 mA to the output transistor pairs (*Q1/Q2* and *Q3/Q4*).

When *Q7* conducts, *Q2, Q3, Q5,* and *Q10* are driven into saturation. Capacitor *C3* charges through *Q2* and *D3,* while *C4* discharges through *Q3, D2* and

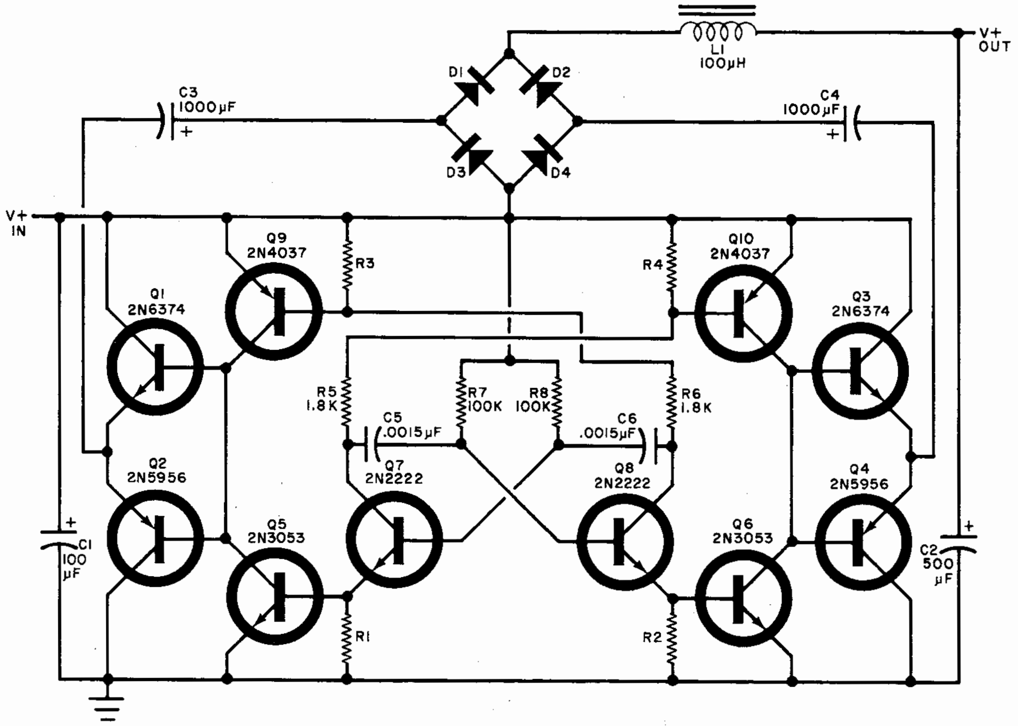


Fig. 1. The boosted dc output voltage is the sum of input dc plus charge on C3 and C4. The two capacitors are oppositely charged by multivibrator.

PARTS LIST

C1—100-µF, 25-V electrolytic capacitor
C2—500-µF, 50-V electrolytic capacitor
C3, C4—1000-µF, 50-V electrolytic capacitor
C5,C6—0.0015-µF, 50-V capacitor
D1 through D4—50-V, 6-A rectifier diode (or bridge)
L1—100-µH, 4-A, 0.1-ohm dc resistance (or less) choke
Q1,Q3—2N6374 or similar transistor
Q2,Q4—2N5956 or similar transistor
Q5,Q6—2N3053 or similar transistor
Q7,Q8—2N2222 or similar transistor
Q9,Q10—2N4037 or similar transistor
Following resistors are ¼ or ½ watt, 10%:
R1 through R4—For input of 6 volts, omit; for 9 V, use 1000 ohms; for 12 V, use 330 ohms; for 15 V, use 180 ohms; for 18 V, use 120 ohms; for 21 V, use 100 ohms; for 24 V, use 82 ohms.
R5,R6—1800 ohms
R7,R8—100,000 ohms
Misc.—10-watt heat sink, suitable enclosure, heavy-gauge wire, mounting hardware, etc.

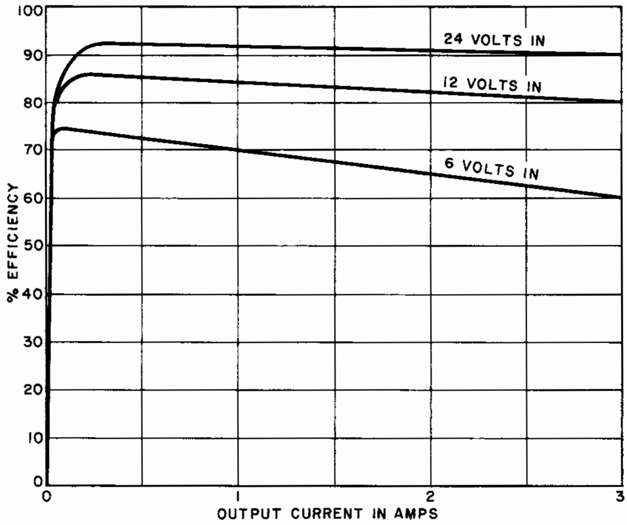


Fig. 2. With 12-volt power source, the efficiency remains above 80% even with full load. The higher the input voltage, the higher the efficiency.

the output load. When *Q8* conducts, the process reverses itself, with *C3* discharging through the output load, *Q1*, and *D1*, while *C4* charges through *Q4* and *D4*. Capacitor *C1* helps prevent high-frequency pulses from entering the supply, while *L1* and *C2* filter the output.

Construction. Since circuit operation is not critical, any convenient means of construction can be used to build the converter. However, for maximum efficiency and minimum ripple, the following should be observed:

Keep all leads, especially those to the output devices, as short as possible. Even the minor inductance of long wires can cause ripple in the output at the high switching frequency used in this voltage converter.

Use 12- or 14-gauge wire for power lines and making connections to output transistors *Q1/Q2* and *Q3/Q4*. If you elect to assemble the circuit on a printed circuit board, use at least ¼″ (6.4-mm) wide copper traces to interconnect the output devices. These measures will eliminate resistive voltage losses.

Connect *C1* as close as possible to the emitter leads of *Q5/Q6* and *Q9/Q10* to minimize ripple feedback into the voltage source.

No heat sinking is required for outputs up to 1 ampere. At higher currents, *Q1* to *Q4* must be on a heat sink that has a thermal resistance of 5°C/watt.

Use. The major loss of efficiency in a capacitive voltage doubler is the inherent voltage drop across the diodes and output transistors. Hence, it is not possible to exactly double the supply voltage, the difference being 1.4 volts with no load to 4.8 volts with a 3-ampere load. Since these losses are a fixed function of the output current, overall efficiency will increase dramatically with higher operating potentials, as shown in Fig. 2. Efficiencies in excess of 98% at several hundred watts can be achieved by a capacitive voltage doubler adapted for 100 volts input. These doublers also have very low ripple, typically less than 200 mV at a 3-ampere output.

The other loss of efficiency is the power consumed by the multivibrator and drive transistors. This is generally less than 150 mW at a 12-volt dc input, which accounts for the circuit's ability to achieve high efficiencies over 98% of its operating range. This sharply contrasts with the much lower efficiencies obtainable with transformer-based converters.

If the circuit shown in Fig. 1 does not satisfy your needs, there are a number of adaptations you may wish to try. For example, higher currents or potentials can be handled by substituting appropriately rated devices and slightly altering the multivibrator. Higher efficiencies can be obtained by substituting germanium power transistors for *Q1* through *Q4* and replacing the rectifiers with four suitable germanium power rectifiers (transistors with the base tied to the collector or emitter). For currents less than 100 mA, *Q1* through *Q4* can be omitted; *C2, C3, C4* need be rated at only 50 μF; and the rectifier diodes (*D1* through *D4*) can be rated at 1 ampere each.

A voltage tripler can be made by adding two 1000-μF capacitors and another bridge rectifier between ground and the common emitters of the output transistors. However, triplers work at lower efficiencies and with considerably reduced output current capability.

It is possible to use the output of one doubler as the input voltage source for another doubler. Two doublers connected in this manner provide about 36 volts output from a 12-volt battery. ◇

dard multimeter with 'GR' inputs.

To mount the plugs, first remove the plastic handles and cut them down to 20 mm so they can be fitted inside the box. Solder about 100 mm of insulated wire to each plug and feed the end through the holes in the box. Grab each plug with pliers and push them through the holes from the bottom. Now slip the handles over the wires and tighten up the plugs.

If all that seems too much, you may want to simply bring a couple of wires out to the multimeter with the banana plugs soldered to the ends.

The spring terminals I used had mounting holes about 45 mm apart which allowed screws to tap into the plastic pillars in the corners of the box. You may also mount the terminals on the long side of the box. In any case, a couple of holes will be necessary under the spring terminals to allow the solder lugs to pass through.

The pc board is straightforward to assemble, watch the orientation of the electrolytic capacitors, the two diodes (note: D1 is the smaller) and the transistor. There are several types of transformer available and some may require R1 to be mounted on the copper side of the board in order to fit properly. The pc board may slot into a groove inside the box, or simply lay alongside the battery as in our prototype.

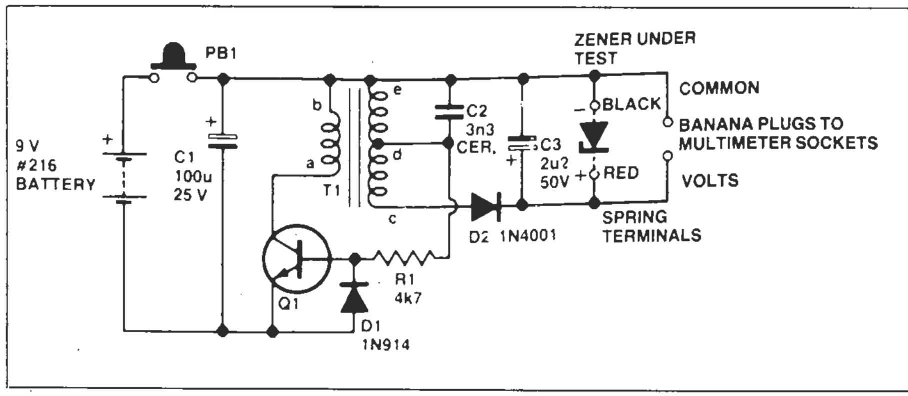
TABLE 1 Performance of prototype.

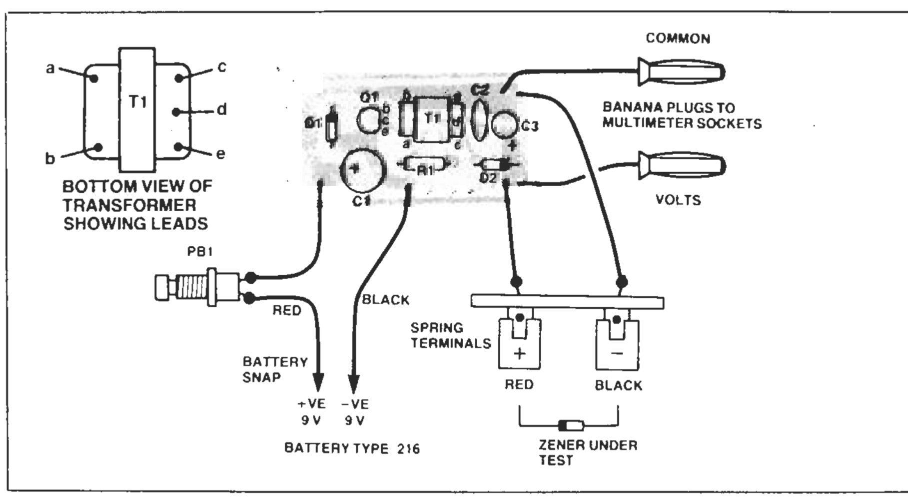
OUTPUT VOLTAGE volts	OUTPUT CURRENT mA	BATTERY DRAIN mA
0	5	140
5	8	160
15	9	190
24	9	190
48	5	160
60	1.5	130
72	0	120

PARTS LIST

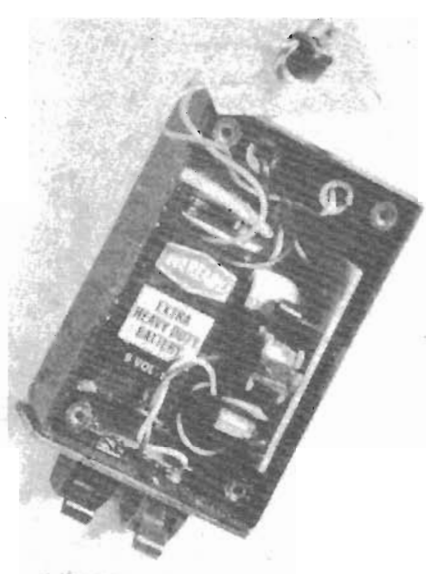
Resistor	¼W, 5%
R1	4k7
Capacitors	
C1	100u/25 V (or 16 V) electro
C2	3n3 ceramic
C3	2u2/50 V electro (see text).
Semiconductors	
D1	1N914
D2	1N4001, 1N4002, etc.
Q1	MPS A06
Miscellaneous	
PB1	momentary action pushbutton
T1	transistor audio transformer, '1k CT to 8 ohm', Hammond 146K or equiv.

pc board; box spring terminals; wire etc.





Diode-LED test. When testing diodes and LEDs, the anode goes to the red terminal, the opposite way to zeners.



Inside story. Internal view of the zener tester. Note the cut-down banana plug handles at the top of the box.

HOW IT WORKS

The operation of apparently simple inverter circuits is usually exceedingly complex, so the following is a simplified explanation.

After PB1 is closed, current flows through terminals 'e' and 'd' of the transformer (and C2) to the base of Q1 starts to conduct and causes current flow through transformer terminals 'b' and 'a' (the primary winding) which causes the magnetic field to build up in the transformer. This field increases the base current to Q1 because of the phasing of the windings.

The magnetic field increases until the transformer core saturates, when the transistor base current reverses, turning the transistor off.

Diode D1 protects the base-emitter junction against excessive reverse bias voltage.

The energy in the transformer's magnetic field is dissipated via several mechanisms, one being to charge C3 via D2.

The whole cycle repeats at a rate of a few kilohertz.

Capacitor C1 provides a low impedance source to ac signals and improves operation with a battery supply.

ETI

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Edited by Bill Travis

5V power supply teams low-dropout regulator, charge pump

Jim Christensen, Maxim Integrated Products, Sunnyvale, CA

Auxiliary power of 3.3V is replacing the 5V auxiliary power that "silver boxes" supply in most computer systems, but some circuits still require a 5V supply. Such systems impose the messy task of creating a central 5V auxiliary supply from the 3.3V auxiliary supply and then routing 5V power throughout the motherboard. An alternative exists, however, for systems in which only a few ICs need 5V: Employ inexpensive charge pumps as low-power 3.3V-to-5V converters and place them directly at the 5V loads. Regulated charge pumps do this job, but they are uncommon, and they often command a premium price. You can build a regulated charge pump by combining an unregulated charge pump with a low-dropout regulator that reduces the voltage to 5V. Unfortunately, that method requires a low-dropout regulator rated for at least 7V, because an unregulated charge pump can deliver 7V when its 3.3V input goes to the upper limit of tolerance. That fact eliminates the possibility of using the latest low-cost, low-dropout regulators, whose small geometry limits their maximum input to 6.5V.

You can reverse the order by placing the low-dropout regulator in front of the charge pump, thereby reducing the 3.3V to 2.5V before doubling it. That approach allows the use of a low-cost, low-voltage, low-dropout regulator, but the charge-pump output impedance then becomes an issue. A low-cost charge pump, such as the MAX-1683, operating with 1-µF capacitors exhibits a typical output impedance of 35Ω, making it unusable at currents above a few milliamps. The circuit of **Figure 1** shows a better way to cascade the charge pump with a voltage regulator. The low-dropout regulator, IC_1, reduces the 3.3V input to a lower value, and the unregulated charge pump, IC_2, doubles that value to 5V. To cancel the voltage drop that charge-pump output impedance causes, the circuit feeds the 5V output back to the low-dropout regulator, which alters its output to maintain output regulation. The available headroom of at least 1V allows output currents to approximately 30 mA or even higher with larger capacitors. Although it requires two ICs instead of a single regulated charge pump, this approach can be cheaper because high-vol-

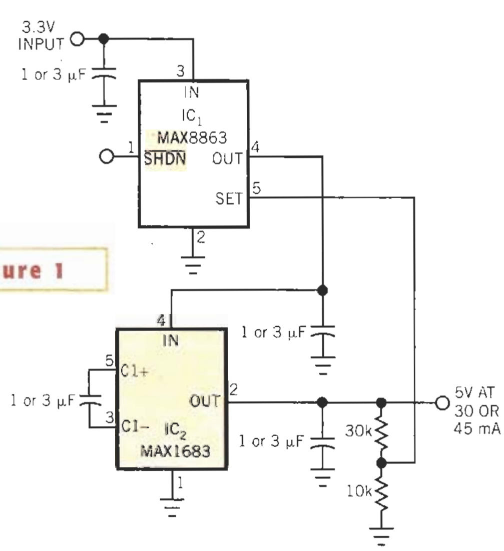
TABLE 1—POWER-SUPPLY PARAMETERS WITH ALL 1-µF CAPACITORS

V_{OUT} (V)	I_{OUT} (mA)	P_{OUT} (mW)	I_{IN} at V_{IN}=3.3V (mA)	P_{IN} (mW)	Efficiency (%)	V_{OUT} low-dropout regulator (V)	V_{RIPPLE} (mV p-p)
5.06	10	50.6	20.9	68.8	73.5	2.71	358
5.01	20	100.2	41.1	135.6	73.9	2.86	312
4.9	30	147	62.2	205.3	71.6	3.02	420

TABLE 2—POWER-SUPPLY PARAMETERS WITH ALL 3-µF CAPACITORS

V_{OUT} (V)	I_{OUT} (mA)	P_{OUT} (mW)	I_{IN} at V_{IN}=3.3V (mA)	P_{IN} (mW)	Efficiency (%)	V_{OUT} low-dropout regulator (V)	V_{RIPPLE} (mV p-p)
4.99	10	49.9	20.37	68.8	74.2	2.63	154
4.99	20	99.8	40.4	133.3	74.9	2.76	104
4.98	30	149.4	60.6	200	74.7	2.89	154
4.93	40	197.2	80.5	265.7	74.2	3.02	192
4.9	45	220.5	90.5	298.7	73.8	3.09	214

Figure 1



This 5V supply, which you obtain by reducing the 3.3V input with a low-dropout regulator and doubling that output with a charge pump, minimizes the charge-pump output impedance by feeding 5V back to the regulator.

5V power supply teams low-dropout regulator, charge pump 59
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ume applications use unregulated charge pumps and low-current, low-voltage, low-dropout regulators. Furthermore, because the low-dropout regulator and charge pump are available in SOT-23 packages, the overall footprint of the circuit in **Figure 1** is comparable to that of a regulated-charge-pump circuit.

Table 1 demonstrates the circuit's ability to maintain output-voltage regulation and deliver currents as high as 30 mA; the input, output, and flying capacitors are all 1 μF. Similarly, **Table 2** shows the regulation for output currents to 45 mA; all capacitors are 3 μF. As you can see, load current does not affect efficiency, which is approximately equal to the output voltage divided by twice the input voltage. Capacitor values affect the ripple voltage and available output current but have little effect on efficiency.□

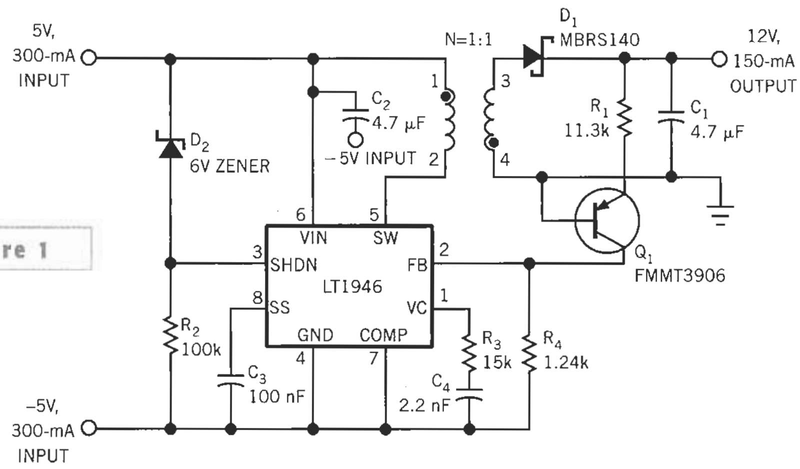
Simple dc/dc converter increases available power in dual-voltage system

David Kim, Linear Technology, Milpitas, CA

THE SCHEMATIC in **Figure 1** shows a way to increase the power available from a current-limited 5V supply by adding power from a −5V supply. The dc/dc converter generates a single 12V, 150-mA (1.8W) output from two regulated and current-limited input sources at 5V, 300 mA (1.5W) and −5V, 300 mA (1.5W). Because the input uses different-polarity voltage sources, the design uses a flyback dc/dc converter to avoid a system-grounding problem. Level-shifted feedback sensing using a pnp transistor, Q_1, references the feedback signal to the negative input voltage. You calculate the feedback-resistor divider by using the formula $R_1=R_4(V_{OUT}-V_{BE})/V_{REF}$, where R_1 connects to the emitter of Q_1, R_4 connects to the collector of Q_1, V_{BE} is the base-emitter voltage of Q_1, and V_{REF} is the feedback reference voltage of the switching regulator.

To simplify the circuit, the flyback converter in **Figure 1** uses an LT1946 monolithic switching regulator. The voltage rating of the monolithic regulator has to be greater than the maximum switching voltage of the flyback converter, calculated by $[(V_{IN1}+|-V_{IN2}|)_{MAX}+V_{OUT(MAX)}/(T_1 \text{ turns ratio})]+V_{SPIKE}$. The maximum switching voltage is approximately 25V for the circuit in **Figure 1**. Note also that the input capacitor and dc/dc regulator input must be able to handle a maximum input voltage of 10V, resulting from the calculation $+V_{IN1(MAX)}+|-V_{IN2(MAX)}|$. In an event of fault-current conditions, such as shorted input or output, a zener diode, D_2, creates the undervoltage-lockout threshold to turn off the LT1946 whenever either input source is in current limit or the input voltage $(+V_{IN1}+|-V_{IN2}|)$ drops below 6V to help the input supply recover when the fault condition is removed. In a system with two available current-limited power supplies, you can convert the two supplies into a single supply that has more power-handling capability than either of the two inputs. A flyback topology based on an LT1946 monolithic converter offers a simple approach to the grounding problem and the feedback-sensing problem inherent in a dual-input power supply. Sharing the power between two input sources not only adds output-power capability, but also increases the overall flexibility of the system.□

Figure 1



Combining two opposite-polarity power supplies can increase the available power from a flyback regulator.

LINEAR TECHNOLOGY

DESIGN NOTES

Direct Efficient DC/DC Conversion of 100V Inputs for Telecom/Automotive Supplies – Design Note 398

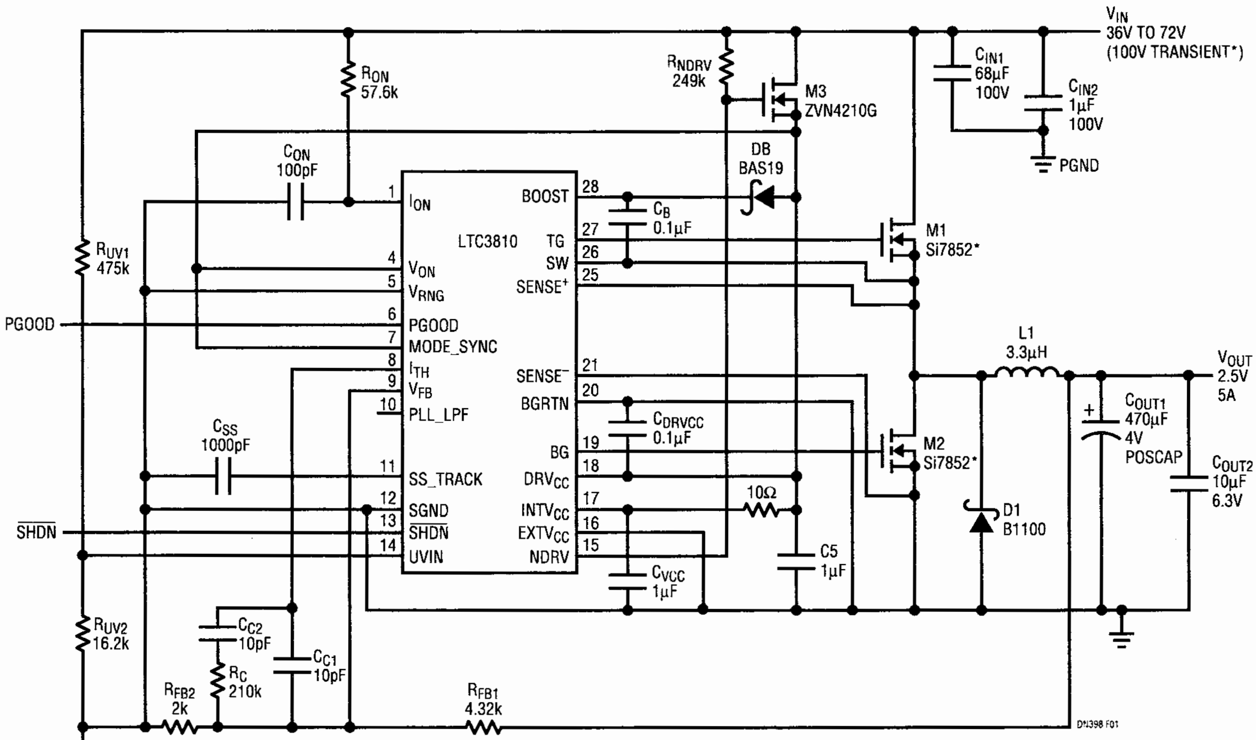
Greg Dittmer

Introduction

Automotive, telecom and industrial systems have harsh, unforgiving environments that demand robust electronic systems. In telecom systems the input rail can vary from 36V to 72V, with transients as high as 100V. In automotive systems the DC battery voltage may be 12V, 24V or 42V with load dump conditions causing transients up to 60V or more. The LTC®3810 is a current mode synchronous switching regulator controller that can directly step down input voltages up to 100V, making it ideal for these harsh environments. The ability to step down the high input voltage directly allows a simple single inductor topology, resulting in a compact high performance power supply—in contrast to the low side drive topologies that require bulky, expensive transformers.

Feature-Rich Controller

The LTC3810 drives two external N-channel MOSFETs using a synchronizable constant on-time, valley current mode architecture. A high bandwidth error amplifier provides fast line and load transient response. Strong 1Ω gate drivers minimize switching losses—often the dominant loss component in high voltage supplies—even when multiple MOSFETs are used for high current applications. The LTC3810 includes an internal linear regulator controller to generate a 10V IC/driver supply from the high voltage input supply with a single external SOT23 MOSFET. When the output voltage is above 6.7V, the 10V supply can be generated from the output, instead of the



*USE Si7456 FOR 100V OPERATION

Figure 1. Compact 36V–72V to 2.5V/6A Synchronous Step-Down Converter

input, for higher efficiency. Other features include:

- Programmable cycle-by-cycle current limit, with tight tolerances, provides control of the inductor current during a short-circuit condition. No R_{SENSE}™ current sensing utilizes the voltage drop across the synchronous MOSFET to eliminate the need for a current sense resistor.
- Low minimum on-time (<100ns) for low duty cycle applications. The on-time is programmable with an external resistor and is compensated for changes in input voltage to keep switching frequency relatively constant over a wide input supply range.
- Precise 0.8V, ±0.5% reference over the operating temperature range of 0°C to 85°C.
- Phase-locked loop for external clock synchronization, selectable pulse-skip mode operation, tracking, programmable undervoltage lockout and power good output voltage monitor.
- 28-pin SSOP package with high voltage pin spacing.

High Efficiency 36V–72V to 2.5V/6A Power Supply

The circuit shown in Figure 1 provides direct step-down conversion of a typical 48V telecom input rail to 2.5V at 5A. With the 100V maximum DC rating of the LTC3810 and 80V for the MOSFETs, the circuit can handle input voltages of up to 80V without requiring protection devices (up to 100V if appropriate MOSFETs are used). This circuit demonstrates how the low minimum on-time of the LTC3810 enables high step-down ratio applications: 2.5V output from a 72V input at 250kHz is a 140ns on-time.

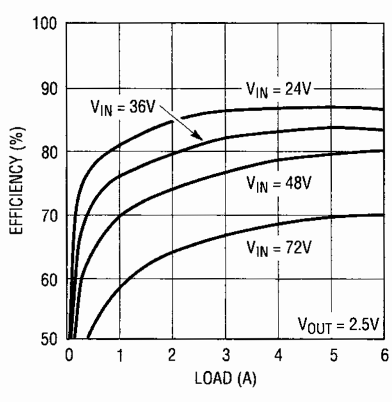


Figure 2. Efficiency of the Circuit in Figure 1

The frequency is set to 250kHz with the R_{ON} resistor to optimize efficiency while minimizing output ripple. Figure 2 shows mid-range efficiencies of 80% to 84% at 36V input and 65% to 70% at 72V input. Type II compensation is used to set the loop bandwidth to about 75kHz, which provides a 20µs response time to load transients (see Figure 3).

The V_{RNG} pin is set to 0V to set the current limit to about 8A (3A after foldback) during a short-circuit condition (see Figure 4). The resistor divider (R_{UV1}, R_{UV2}) sets the input supply undervoltage lockout to 24V, keeping the LTC3810 shut-down until the V_{IN} > 24V.

The LTC3810's internal linear regulator controller generates the 10V IC/driver supply ($INTV_{CC}$, DRV_{CC} pins) from the input supply with a single external MOSFET, M3. For continuous operation the power rating of M3 must be at least (72V – 10V) • (0.02A) = 1.2W. If another low voltage supply (between 6.2V and 14V) capable of supplying the ~20mA IC/driver current is available, this supply could be connected to $INTV_{CC}$/DRV_{CC} pins to increase efficiency by up to 10% at loads above 1A.

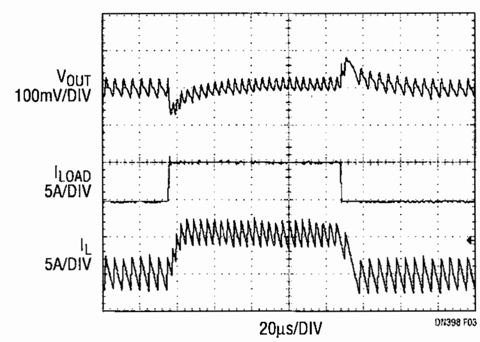


Figure 3. Load Transient Performance of Figure 1 Circuit Shows 20µs Response Time to a 5A Load Step

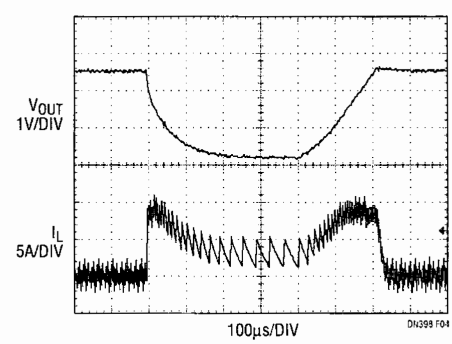


Figure 4. Short-Circuit Condition in Figure 1 Circuit Shows Tight Control of Inductor Current and Foldback

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