18. Switch-Mode Electronic Power Conditioning

These days, hardly any low-voltage dc power-supply designs are not "switch-mode" (or "switched-mode," as grammar sticklers insist on calling them). From our previous discussions of pulse modulators we should be prepared to expect that the active devices in a switch-mode power supply are operated as electronic switches and not as linear or proportional amplifiers, and this is indeed the case. Furthermore, in modern usage the term switch-mode has become all but synonymous with solid-state, which is not always the case.

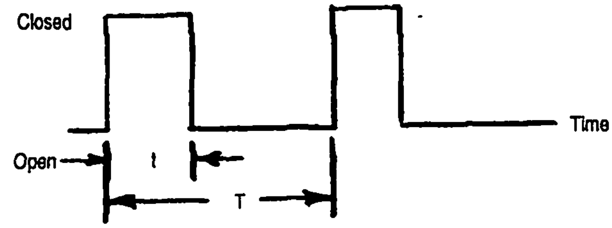
Why, we might ask, put a switchlike device in a dc power supply in the first place (other than to turn it on or off)? Consider the problem of converting the output of a source of primary power that can produce only dc, such as a battery or a photovoltaic solar cell, to a different voltage level. What we need, of course, is a dc transformer. Unfortunately, as we all know, such a thing does not exist. Or does it?

18.1 Switch-mode dc variable-voltage circuits (dc-to-dc converters)

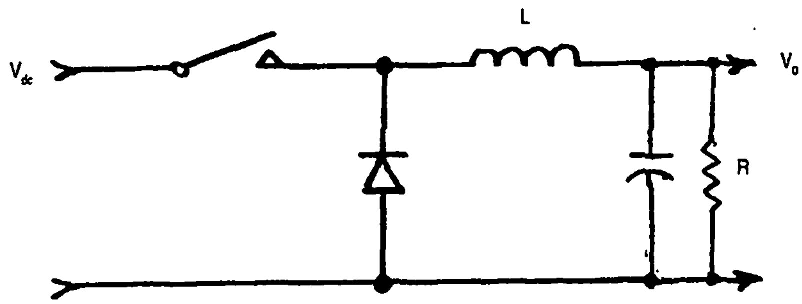
Faced with the problem of converting the voltage of a dc source to a lower voltage at a load, a very literal technologist might reason that if a switch were placed between the source and the load, and that switch was repetitively turned on and off, the time-averaged voltage measured at the load would be smaller than that of the source simply because the source voltage was not being applied to the load all of the time. When the switch is closed, the load voltage is equal to the source voltage. When the switch is open, the load voltage is zero, as shown in Fig. 18-1a. Even though this could hardly be called dc transformation, the literalist is, of course, basically correct.

Suppose, however, that some additional components were added to the simple series-switch circuit—an inductor, a diode, and a capacitor—and they were arranged as shown in Fig. 18-1b. When the switch is closed, current will first start to build up linearly with time at an initial rate of V_{dc}/L and will exponentially approach a value of V_{dc}/R if the switch remains closed for a long-enough time interval. However, if the switch is repetitively opened and closed at a repetition rate that is high enough so that the interpulse interval, T, as shown in Fig. 18-1a, is small with respect to the circuit time-constant, L/R, then the time-averaged voltage at the load side of the switch will be $V_{dc} \times t/T$, where t is the length of time in each switching interval that the switch is closed. The current from the source will be discontinuous because it is being chopped by the switch. The current in the load, however, will be continuous, because the current in the inductor has no time to change from one switch-conduction interval to the next. When inductor current is not being supplied by the source through the closed switch, it is flowing through the shunt diode, which is often called a free-wheeling diode.

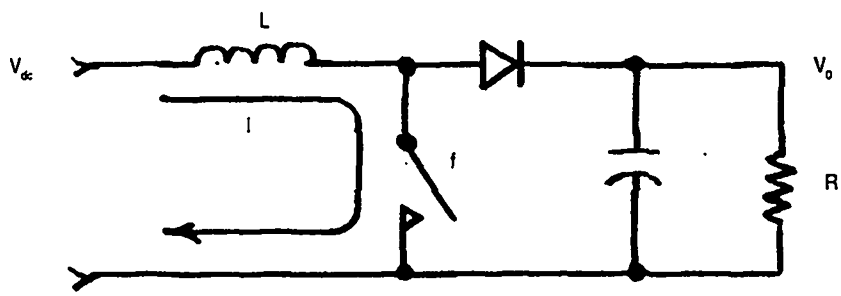
The circuit of Fig. 18-1b is called a single-quadrant chopper, or a buck regulator, because its output voltage is always smaller than the source voltage (not because they cost a dollar, which almost none do). When $t = T$, the switch is



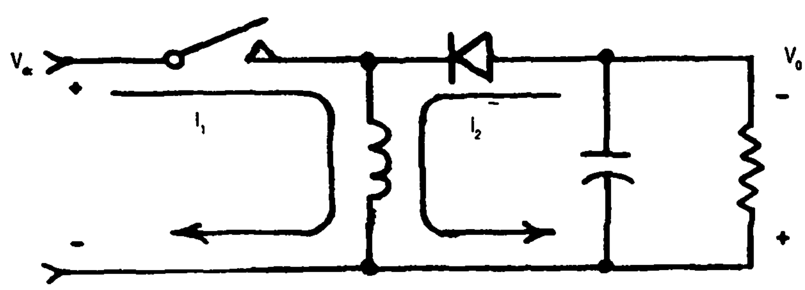
A. Timing of switch conducting and blocking intervals.



B. "Buck" type of switch-mode dc variable-voltage circuit.



C "Boost" type of switch-mode dc variable-voltage circuit.



D. "Buck/boost" type of switch-mode dc variable-voltage circuit.

Figure 18-1. Different types of switch-mode dc variable-voltage circuits.

closed all of the time and $V_0 = V_{dc}$. When $t = 0$, the switch is never closed and $V_0 = 0$. When t equals $1/2\ T$, $V_0 = 1/2\ V_{dc}$. If the output voltage is sampled and used as an input to a feedback circuit that can modulate the switch-conduction interval, t, the output voltage can be precisely regulated. This is called a pulse-width-modulated (PWM) regulator, which usually runs at a precise clock rate ($1/T$). The higher this clock rate, the smaller the value of L can be and the smaller its core can be, if one is needed.

For switch-mode power conditioning, a full-control electronic switch is required, just as in the hard-tube pulse modulator. In general, the faster the switch can be turned on and off, the better. The efficiency of the circuit is primarily determined by the conduction-voltage drop of the switch, although there are "switching" losses as well that relate to circuit shunt capacitance. These losses are proportional to the clock rate. If the conduction-voltage drop of the switch approaches zero, the efficiency of the circuit approaches 100%. There are no intentional losses in the circuit. All of the power that leaves the source is delivered to the load. The peak current from the source is always equal to the average current in the load. If, for instance, the regulator is operating with $t = T/2$, the load voltage will be $V_{dc}/2$ and the load current $V_{dc}/2R$, which is also the peak source current. The load power is $V_{dc}^2/4R$. The peak source power is $V_{dc}^2/2R$, but for only half of the time. So the average source power is $V_{dc}^2/4R$, which is the same as the load average power, even though the source voltage is twice the load voltage.

The transient response of the circuit, which is related to its closed-loop bandwidth, also depends on the value of L/R, which determines how rapidly current can change in the load circuit. Note that none of the currents in the circuit ever change direction, or alternate. They are all dc. Is this, then, a dc transformer?

The buck circuit is not the only configuration of the key components that will produce the equivalent of dc transformation. The circuit of Fig. 18-1c is the dual of the buck circuit. It can only boost. When the switch is open all of the time, the output voltage is equal to the input voltage. When the switch is repetitively opened and closed, the output voltage is greater than the input voltage. Note that unlike the buck regulator, this circuit cannot operate with a unity switch-conduction duty factor. With the switch closed all of the time, the source will eventually become short-circuited by the dc resistance of the inductor and there will be no output at all.

However, at some switch duty factor less than unity, closing the switch momentarily will cause current to build up in the inductor at a rate of V_{dc}/L. When the switch is opened again, there will be energy stored in the magnetic field of the inductor of $W = 1/2\ LI^2$, and the voltage at the load end of the inductor will rise to whatever value is required to cause conduction through the diode and the load circuit. (So far, this circuit is the same as the inductive-storage control-grid modulator shown back in Fig. 12-10.) If the switch remains open long enough for all of the inductor's stored energy to be delivered to the load circuit during each switch cycle, the average power in the load will be $P_0 = W \times f$, where f is the recurrence, or clock rate, of the switch closures. Assuming that V_0 is considerably greater than V_{dc}, the load power is $V_0^2/R = W \times f = 1/2\ LI^2 \times f$, therefore

$$V_O = \sqrt{P_O \times R} = \sqrt{\frac{1}{2} L I^2 \times f \times R}\ .$$

Note that the output voltage is proportional to $\sqrt{R}$, other conditions being equal.

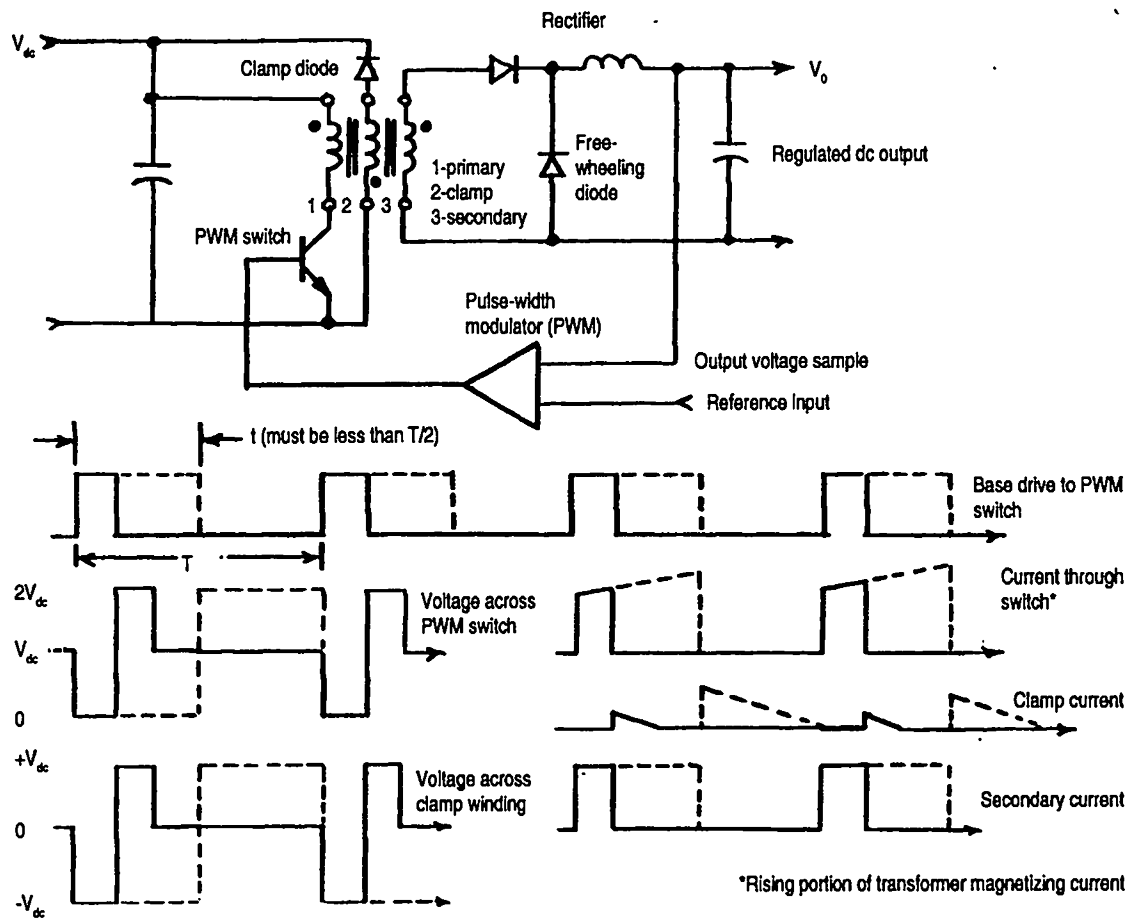


Figure 18-2. The single-quadrant dc-ac inverter with rectifier and pulse-width modulation regulation.

The inductor serves as a "charge pump," lifting charge and energy from the source to a load whose voltage can be many times greater.

Figure 18-1d shows yet another arrangement of the key components. This one yields a buck/boost converter circuit that is capable of producing output voltage from zero to infinity—theoretically, anyway. The output voltage is zero at zero switch-conduction duty factor. As the switch-conduction duty factor increases, output voltage will increase in a fashion similar to that of the boost converter, except that the polarity of the load voltage is the opposite of that of the source. Unity switch-conduction duty factor cannot be reached because it will short-circuit the source through the dc resistance of the inductor. Note that even though the direction of current flow through the load is the opposite of that through the source, neither change direction. They are both dc, one intermittent and the other continuous.

18.2 DC-AC inverters

Although the circuits of Fig. 18-1, much like a transformer, are capable of converting one dc voltage level to another with theoretically no loss in power transferred, they are actually like the single-winding autotransformer in that one terminal is common to source and load. To realize nearly complete power transfer and the isolation of a true four-terminal network, there is no substitute for the two-winding transformer.

The simplest dc-dc converter circuit with output-circuit isolation is the type shown in Fig. 18-2. This circuit uses a single full-control switch in a role similar to the switch tube in the transformer-coupled hard-tube modulator shown in Fig. 10-3. When the PWM switch is gated into conduction, current flows from the source, V_{dc}, through the primary winding of the transformer. If the switch is nearly perfect—that is, if it has low collector-emitter voltage drop during conduction—the voltage across the transformer winding will be nearly V_{dc}. During the conduction interval, current will also flow through the rectifier diode and load circuit connected to the secondary winding. The value of V_{dc}, the transformer primary/secondary turns ratio, and the resistance of the load will all determine the magnitude of the flat-top portion of primary current. Magnetizing current will also linearly rise throughout the conduction interval, producing an upwardly sloping total primary current.

The currents in the primary and secondary windings are unidirectional, or dc, but they are not continuous. However, there can be no average value of voltage across any transformer winding because the windings are, in the steady-state, simply lengths of wire. The size of the core they are wound around makes no difference. In order to satisfy the criterion that there is no average voltage, the volt-time products must be the same in both polarities. The energy stored in the transformer magnetizing inductance will cause the transformer voltage to reverse the instant the PWM power switch is turned off. The current that had built up in the magnetizing inductance during the conduction interval wants to continue flowing, just as it did in the transformer-coupled hard-tube modulator with a diode load. Rather than being dissipated, the energy stored can be returned to the power source by means of an additional transformer winding, a clamp winding, and a clamp diode. If the clamp winding has the same number of turns as the primary but is of opposite polarity as shown, voltage will reverse across the clamp winding at the same time it reverses across the primary. The clamp diode will conduct as soon as the clamp winding voltage becomes more positive than the input voltage, V_{dc}, and the magnetizing current will continue to flow back into the source, linearly decreasing with time, until the energy has been restored to the source. The voltage across the switch during the non-conducting interval will be clamped at twice V_{dc}. The switch duty factor cannot exceed 50%, because that would produce the limiting case of equal volt-time products. If the clamp winding had a different number of turns, higher peak voltage on the switch could be exchanged for a higher duty factor, or vice versa.

The load voltage will be a function of the switch duty factor. At constant clock frequency, or constant value of interpulse interval, T, the duty factor will be proportional to the on time, t, or pulse width. Modulation of pulse width in response to an error signal generated by the difference between the sampled output voltage and a reference voltage can result in a regulated output voltage. The greater the clock rate, the smaller the transformer volt-time products will be and the smaller the transformer core can be made before saturation occurs. When attempting to produce high-voltage output, however, there are limits to the optimum switching rate.

The circuit just described is called a single-quadrant forward converter because current flow in the load circuit and in the switch occurs at the same time. A circuit similar to it, called a flyback converter, operates in the opposite mode. The polarity of the secondary is reversed. During switch conduction, energy is deliberately stored in the transformer magnetizing inductance because only magnetizing current flows. When the switch is turned off, the energy stored in the magnetizing inductance is transferred to the load (in similar fashion to the boost and buck/boost converters, which also operate in the flyback mode). No clamp winding is required, of course. The most common example of the flyback converter is the high-voltage power supply of a television receiver, which makes use of the energy stored in the horizontal-deflection coils during the retrace intervals at the ends of each horizontal trace.

In transmitter circuits, even low-power ones, the capabilities of the single-switch converter are usually inadequate. Moreover, the current flow in both primary and secondary windings is unidirectional, so full use is not made of core magnetization, meaning that larger cores are needed for a given power level for this kind of circuit than for others. A commonly used alternative is the push-pull converter, shown in Fig. 18-3. It uses two switches that are connected to opposite ends of a center-tapped primary winding. The switches conduct alternately for equal times. The currents in the two halves of the primary winding are unidirectional, but they magnetize the core alternately in opposite directions, so there is no dc component of magnetization. The rectifier circuit shown is a full-wave center-tapped type. It also conducts alternately in synch with switch conduction. (Note that any type of full-wave, single-phase rectifier would be appropriate.) No free-wheeling diode is required for the filter inductor, because total rectifier conduction is continuous, so there is always a path for filter-inductor current.

In the circuit shown, the transformer-core magnetization capability is fully utilized, but the ohmic capacity of the windings is not. Each winding conducts half of the average load current. But it is in the form of a square wave whose peak current is equal to the average load current of the 50% duty factor (normalized to the transformer turns ratio). Remember from Fig. 6-2 that the average value of this waveform is $I_{PK}/2$, but the RMS value is $I_{PK}/\sqrt{2}$. Therefore, the

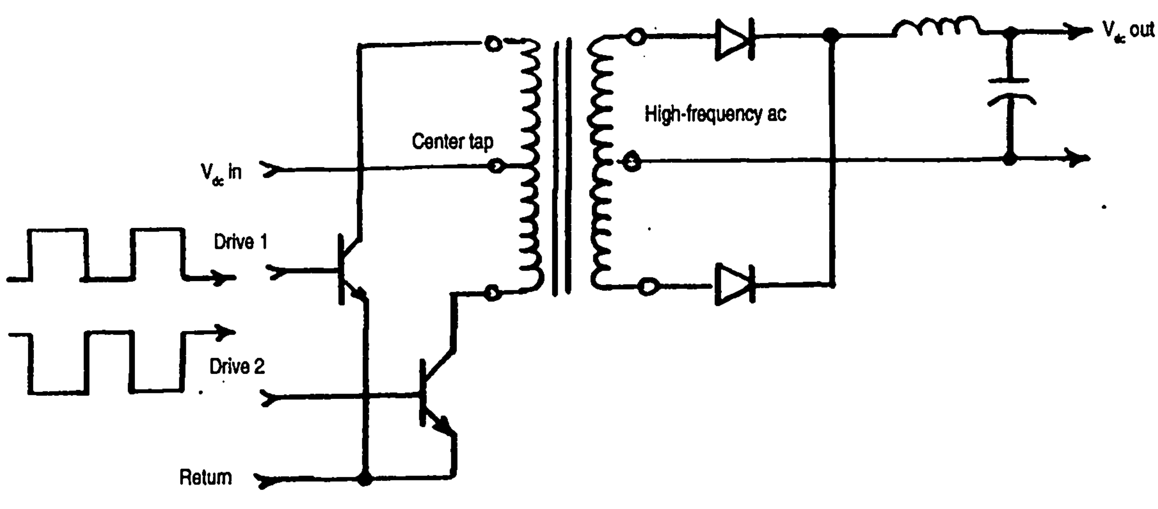


Figure 18-3. The push-pull switch-mode dc-ac inverter with full-wave, center-tapped rectifier.

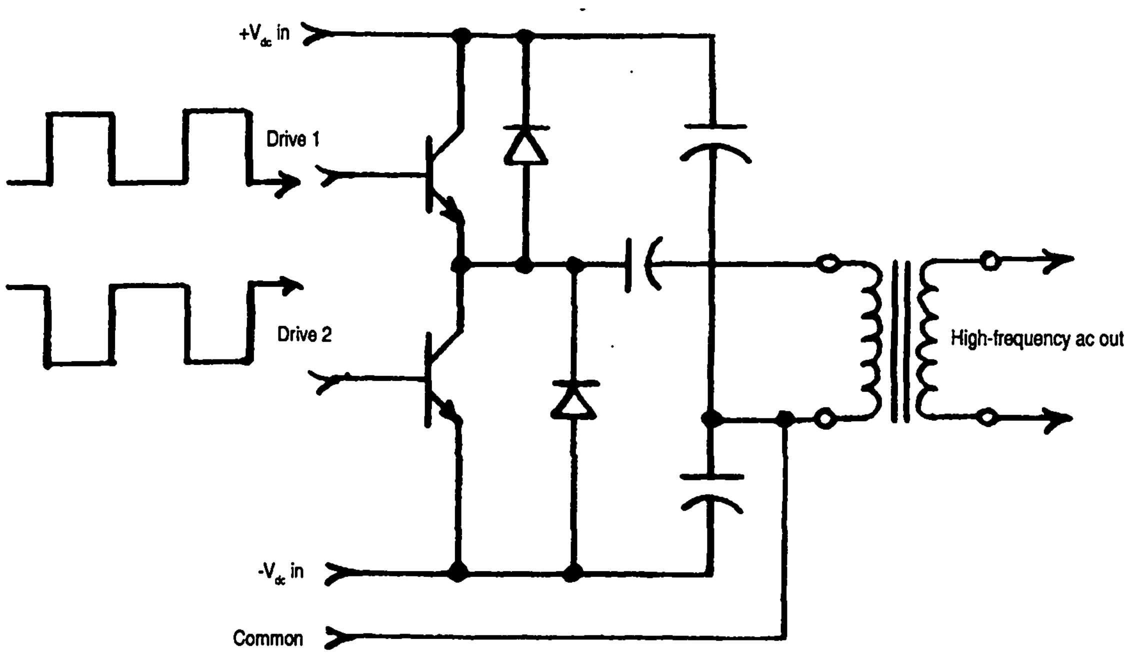


Figure 18-4. The half-bridge, or "totem-pole," switch-mode dc-ac inverter.

dissipation in the winding is twice the minimum achievable value. The same is true of the secondary winding for the type of rectifier shown. Despite this drawback, however, such a circuit may prove useful in certain circumstances. The push-pull inverter produces a transformer input of ac—or, more correctly, *av* (alternating voltage)—that has a peak-to-peak value of 2 x V_{dc}.

An alternative type of inverter, shown in Fig. 18-4, produces true ac input to the rectifier-transformer primary. It is called the half-bridge, or "totem-pole," inverter, and it is a circuit topology that we have certainly seen before. Like the push-pull inverter, it uses two switches that are alternately gated into conduction. In order to achieve an input voltage to the transformer of 2 x V_{dc}, the dc source must be $2V_{dc}$ or, as shown, plus and minus $1V_{dc}$. In the push-pull inverter, the only anomaly capable of producing a dc component of transformer magnetization (assuming a perfectly center-tapped primary winding) is to time the push-pull switch conduction at some time interval other than 50/50. In the half-bridge inverter, a coupling capacitor in series with the primary winding will assure that there is no dc component. Without the capacitor, differences between the absolute values of $+V_{dc}$ and $-V_{dc}$, and a switch timing other than 50/50 can both produce a dc offset. If a source voltage of $2V_{dc}$ is used, there is no alternative to the coupling capacitor. It will automatically assume an average voltage drop nominally equal to V_{dc}, but it will differ from that amount by whatever increment is required to assure that the inverter output is true ac with precisely equal positive and negative volt-time products.

Although the half-bridge circuit has been shown as a generic inverter, it would not necessarily be a component of a dc-dc converter that might be used to drive an arbitrary load. Unlike a rectifier load, which has a unity power factor with square-wave input, the current and voltage of an arbitrary load cannot be expected to have the same wave shape, let alone change direction at the same

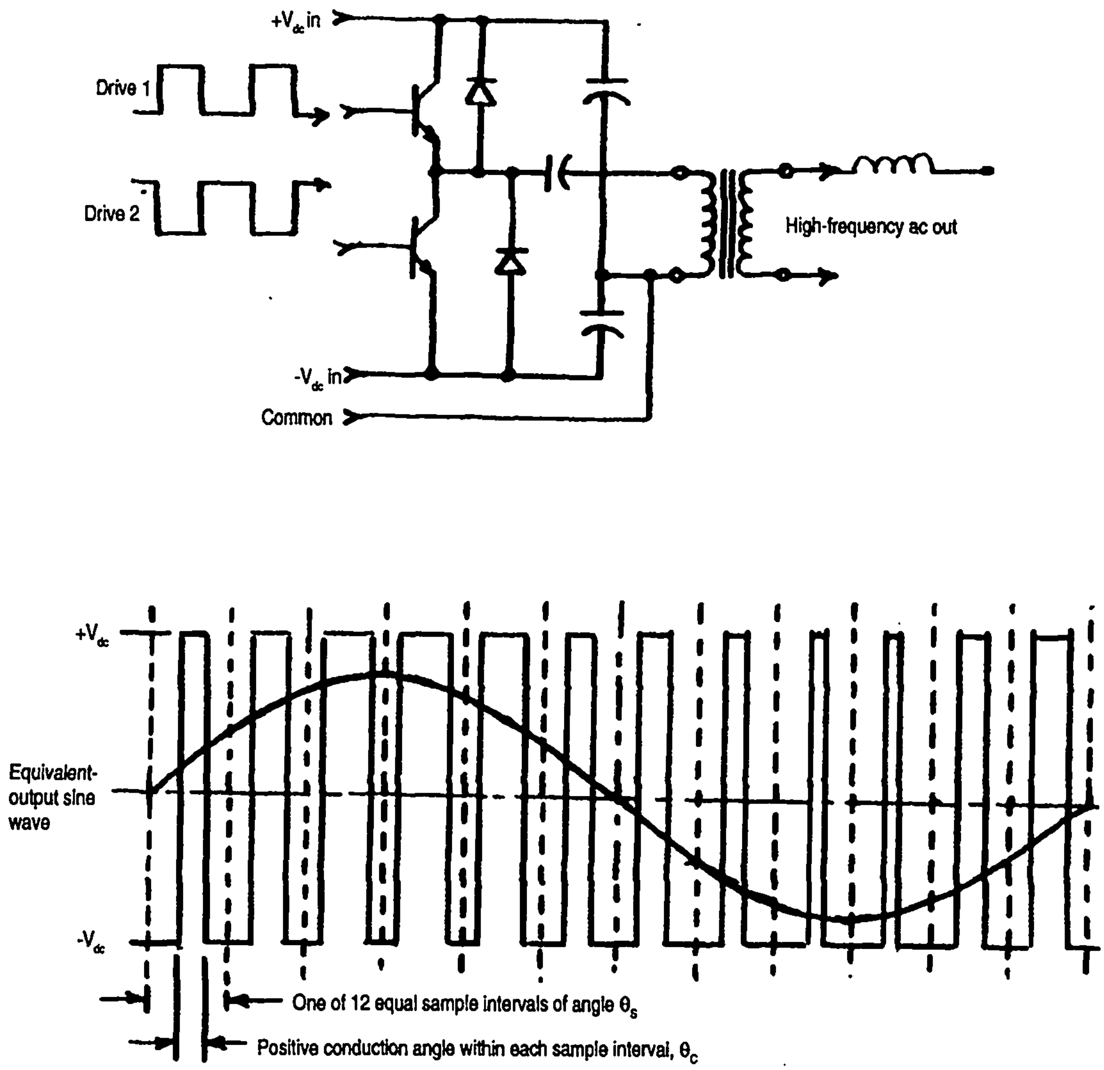


Figure 18-5. Pulse-width-modulation strategy for the bridge inverter to synthesize sine-wave output.

instant. For that reason, the generalized inverter must have the internal capability to pass currents in either direction at all times. This is the function of the anti-parallel diodes that shunt the inverter switches.

Simultaneous conduction, or shoot-through, of the switches in the half-bridge is always a possibility and can cause serious stress on the switching devices. A gating strategy is required to assure that one switch has ceased conduction before the other one begins. The power MOSFET requires less attention in this regard than the power bipolar transistor, which has inherent turn-off delay due to minority-carrier cleanup. (In the push-pull inverter, the magnetizing inductance of the transformer will at least slow the rate-of-rise of shoot-through current, making the problem less severe.)

Figure 18-5 shows a half-bridge inverter being used to synthesize a sinusoidal output waveform by means of a pulse-width-modulated bipolar output. The inverter operates at a clock frequency that is many times greater than the output frequency. In the example shown, it is 12 times as great, dividing up the output sinusoid into 12 sample intervals, each one dealing with the appropriate 30°

segment of the output wave. The load has an inductor-input band-pass filter. The rail-to-rail bipolar input voltage is developed across this inductor. The proper half-angle of conduction is different for each sample interval. It is determined by solving the following equation

$$\frac{\sin\theta_C}{2} = \frac{1}{4\sin\left(\theta_1 + \frac{\theta_S}{2}\right)}$$

$$\left\{\cos\theta_1 - \cos(\theta 1 + \theta_S) + \frac{K}{2}\left[\theta_S + \sin\theta_1\cos\theta_1 - \sin(\theta_1+\theta_S)\cos(\theta_1+\theta_S)\right]\right\},$$

where θ_S is the angle of the equal-sample interval, θ_C is the positive conduction angle within each sample interval, θ_1 is the angle (referred to the output sine wave) at the beginning of each sample interval, and K is the ratio of peak output sine wave to input V_{dc}.

Just such an inverter was designed and built, but it used triode vacuum tubes instead of transistors as switches. The triodes were driven from a computer program that approximated the solution of the equation to produce frequency-shift-keyed, extremely low-frequency (ELF) signals. This application was the prototype for an early version of the ELF submarine communication system.

With more switches, even more can be done with an inverter. The full-bridge inverter shown in Fig. 18-6 uses four switches, two of which conduct simultaneously (1A and 1B, or 2A and 2B). In return for the additional complexity and switch dissipation, the full-bridge inverter produces a peak-to-peak input to the transformer primary of 2 x V_{dc} for a source voltage of V_{dc}. A coupling capacitor in series with the transformer primary once again can insure that there is no dc component by automatically assuming whatever average voltage across it is re-

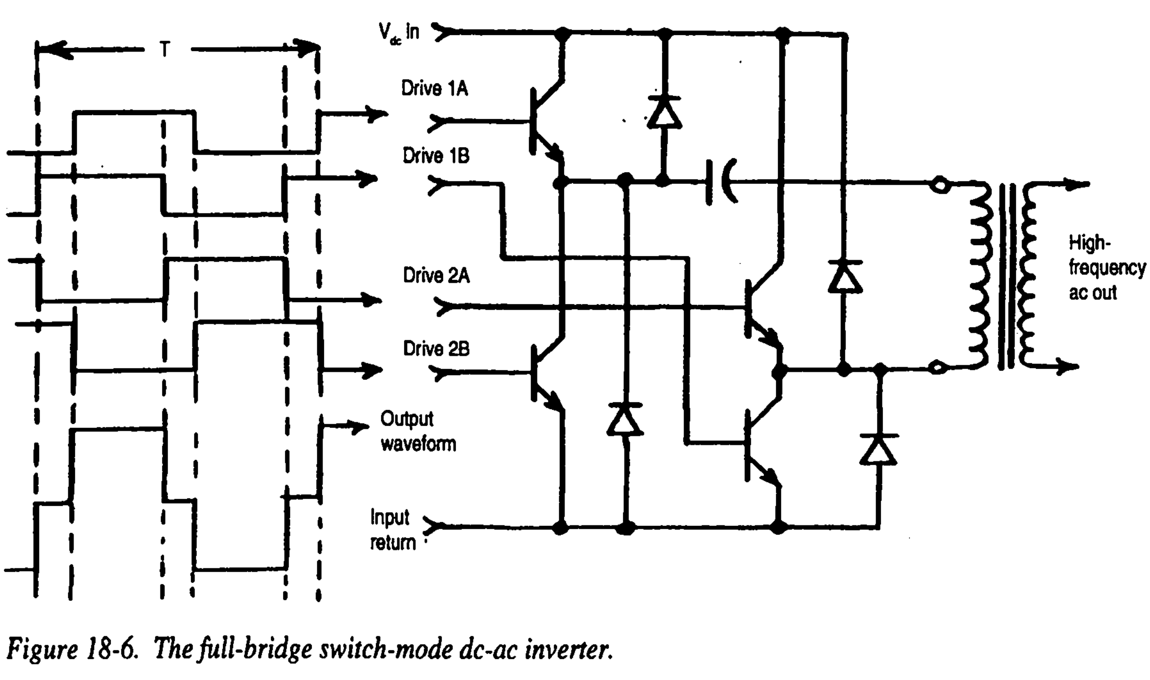


Figure 18-6. The full-bridge switch-mode dc-ac inverter.

quired to nullify any timing asymmetry. Primary voltage asymmetry is unlikely because the dc source is first connected across the primary in one direction (1A and 1B conducting) and then in the other (2A and 2B conducting). A three-state, pulse-width-modulated output voltage is achieved by varying the relative timing of square-wave gating signals, as shown. During the zero, or neutral, state, both transformer primary terminals are clamped either to the source high-side or to the return (either 1A and 2A conducting or 1B and 2B). This clamping short-circuits the primary winding during the neutral state, minimizing possible ringing or other transient effects when zero output is desired.

The switch-mode converter and inverter circuit topologies discussed are by no means the only ones. Resonant- and quasi-resonant-type inverters have been mentioned before (see Fig. 9-45) and will be again, but even their inclusion does not round out the complete list. The most modern switch-mode circuits have been made practical by the development of powerful solid-state, full-control switches such as the power MOSFET. (In order for the switch function to be performed by a single device, it must have full-control properties. In the past, large-power circuits were switched with half-control devices such as SCRs and even Ignitrons, that had to be shut off by "forced" commutation, which required additional components and imposed a serious limit to the switching rate.)

As an interesting side note, let me say that had MOSFETs been in existence in the early days of electricity, it is conceivable that the point of view of Thomas Edison, who was an influential proponent of dc power distribution, might have prevailed over that of George Westinghouse, the most powerful proponent of ac power distribution. (At least the controversy would have been more interesting, especially considering that the highest-voltage power transmission system in this country is the Pacific Intertie, which operates at 1 MVdc and uses power rectifiers and power inverters at its terminals.)

Even before the advent of modern super-power solid-state switches and rectifiers, dc-ac inversion and dc-dc conversion was common. Literally millions of automobile radio receivers used vacuum-tube amplifiers with anode supply voltages up to 300 Vdc. They were powered from automotive storage batteries whose voltage was as low as 6.3 V at one time. The repetitive switch function in these radios was performed by a device called a vibrator, in which contacts of an actual mechanical switch were magnetically opened and closed at a low audio frequency (low enough that one could tell whether or not it was vibrating by the sound it made). The most sophisticated of the vibrators was a "synchronous vibrator." It had two sets of contacts, one in the primary circuit of a step-up transformer and the other in the secondary to directly rectify the output. Switch-mode power conversion, therefore, predates the solid-state revolution by a number of decades.

Up until now, the discussion of converters and inverters has been limited to those that operate from a dc power source. Conversion to a different voltage level with no power loss or inversion to ac requires a controllable switch or switches. There is no alternative. The faster the switch or switches can be made to turn on and off, the smaller the other components can usually be made, especially the iron-core components. For transmitter applications, however, these circuits may be required to produce relatively high voltages, or at least voltages that are many times greater than that of a typical dc source. As output-voltage levels increase, step-up transformers require more insulation between primary and secondary windings. More insulation means greater physical separation between windings and greater leakage inductance. The reactance of this leakage inductance increases with increasing frequency, thus limiting the amount of voltage step-up that can be achieved for a given output current. For each specific application, there will exist an optimum transformer design and inverter operating frequency that will be defined by the source voltage and the requirements for output voltage and power. Although optimization studies are rarely performed, it is true that few high-power and high-voltage systems using switch-mode, high-frequency technology, operate at more than 10 kHz. And the true "optimum" frequency in many cases may be considerably less. As one representative of a transformer and power-supply vendor succinctly put it, "High frequency and high voltage don't mix."

When the source of primary power is ac to begin with, or, as is sometimes the case, when the prime mover is something as non-electrical as a rotating shaft, the choice of switch-mode power conditioning should deserve even more thought. The switch-mode power supply operating from an ac source is often called an off-line switcher, meaning that it operates "off" a power line, not that it is broken and therefore "off-line." The first thing that must be done in an "off-line" power conditioner is rectification to produce the internal dc rail for the high-frequency inverter, whose output is then transformed in voltage and eventually rectified again to get back to dc. At low voltages and relatively low power levels, all of these steps may be justifiable for a high-frequency inverter in terms of overall size, weight, and efficiency, especially if output-voltage variability and regulation are required. For this kind of device, "high" efficiency is usually in the 65-85% range, which is not high at all when compared with the overall efficiency of a high-voltage, high-power transformer-rectifier system. These systems can easily exceed 95% efficiency. But the size and weight of such a high-voltage power supply operating from commercial 60-Hz ac source can be immense when compared to one that operates from a much higher-frequency ac source.

Techniques have been developed that offer dramatic reduction in the size and weight of iron-core components operating even from 60-Hz ac. They involve the use of wire that may be one-tenth the diameter of that used to wind a transformer of conventional size and efficiency. Obviously, such a ploy increases winding losses, but the windings must then be cooled effectively to keep them from overheating. The transformer efficiency will also be degraded by using smaller wires, perhaps to 90%. In return, however, the size and weight of the core can be drastically reduced because the same number of winding turns can fit in a much smaller core window. The overall efficiency may be not be much greater than that of a complete off-line high-frequency, switch-mode power supply, but the circuit complexity might be two orders of magnitude less. And it might be less expensive and more reliable as well. Such techniques are only rarely investigated now because of the "technological correctness" of the electronic alternative, but high-frequency, switch-mode power conditioners are neither simple nor cheap. If a designer has the time, all options should be studied.

When the source of system input power is the prime-mover itself, and the choice of mechanical-to-electrical power conversion is at the system architect's discretion, even more options are available. High-frequency three-phase alternators operating up to 1 kHz are practical. It goes without saying that 400-Hz alternators are practical because they are standard for most large-power aircraft auxiliary power units. Iron-core components can be made quite small and lightweight to operate from 400 Hz—even without resorting to the ultra-miniaturization techniques mentioned above—and they can be made even smaller to operate at 1000 Hz. The problems of leakage reactance and winding skin-effect are only moderate. If an alternator is dedicated to a single power supply, voltage regulation and variability can be accomplished by modulation of the alternator-field excitation. The alternator can also perform the function of high-speed electrical disconnect.

18.3 Voltage-multiplier rectifier circuits

If, even after visiting all of the aforementioned considerations, the designer finds the lure of the high-frequency, switch-mode conditioner still irresistible, then he or she should know that there are ways to overcome the limited transformer step-up ratio. One of them is the voltage-multiplier rectifier circuit. A basic form of a voltage multiplier rectifier is shown in Fig. 18-7. It is an iterative circuit that can be expanded to almost any level of voltage multiplication. Illustrated are the two-pulse voltage doubler, the three-pulse voltage tripler, and the four-pulse voltage quadrupler.

The "pulses" referred to are, in the cases illustrated, polarity reversals of the input alternating voltage. During the first "pulse," the high side of the transformer secondary will be positive with respect to the grounded return. Current will flow clockwise through the first capacitor and the first diode. If the source impedance is low enough, the capacitor will be charged to the peak value of the input voltage, V, especially if the wave shape is rectangular. When the polarity of the input reverses, assuming that the second capacitor has no charge as yet, the total voltage in the loop, which includes the first two capacitors and the second diode, will be $2V$. Current will flow counterclockwise. If the second capacitor is many times smaller than the first, it will be charged to a voltage of $2V$ with the polarity as shown. During the third "pulse," the source voltage reverses polarity again. The first and the third diodes will now conduct. Current I_1 flows through the first diode and replenishes the charge on the first capacitor. If the first capacitor has lost no charge, I_1 will be zero because there will be no net loop voltage. The net voltage in the loop, including the first three capacitors and diodes, will be $2V$ again and be supplied by the voltage across the second capacitor. The source voltage and the voltage across the first capacitor are of opposite polarity and cancel each other. Assuming that the third capacitor is much smaller than the second, current I_2 will charge the third capacitor to $2V$ with the polarity shown. The fourth "pulse" will produce current through the first and fourth diodes and all of the capacitors. If, once again, the fourth capacitor is much smaller than the third, it will be charged to $2V$, which is the initial loop voltage ($2V + V + V - 2V = 2V$). As described, the doubler will have an output voltage of $2V$ after two "pulses," the tripler will have an output of $3V$ after three "pulses," and the quadrupler will have an output of $4V$ after four "pulses."

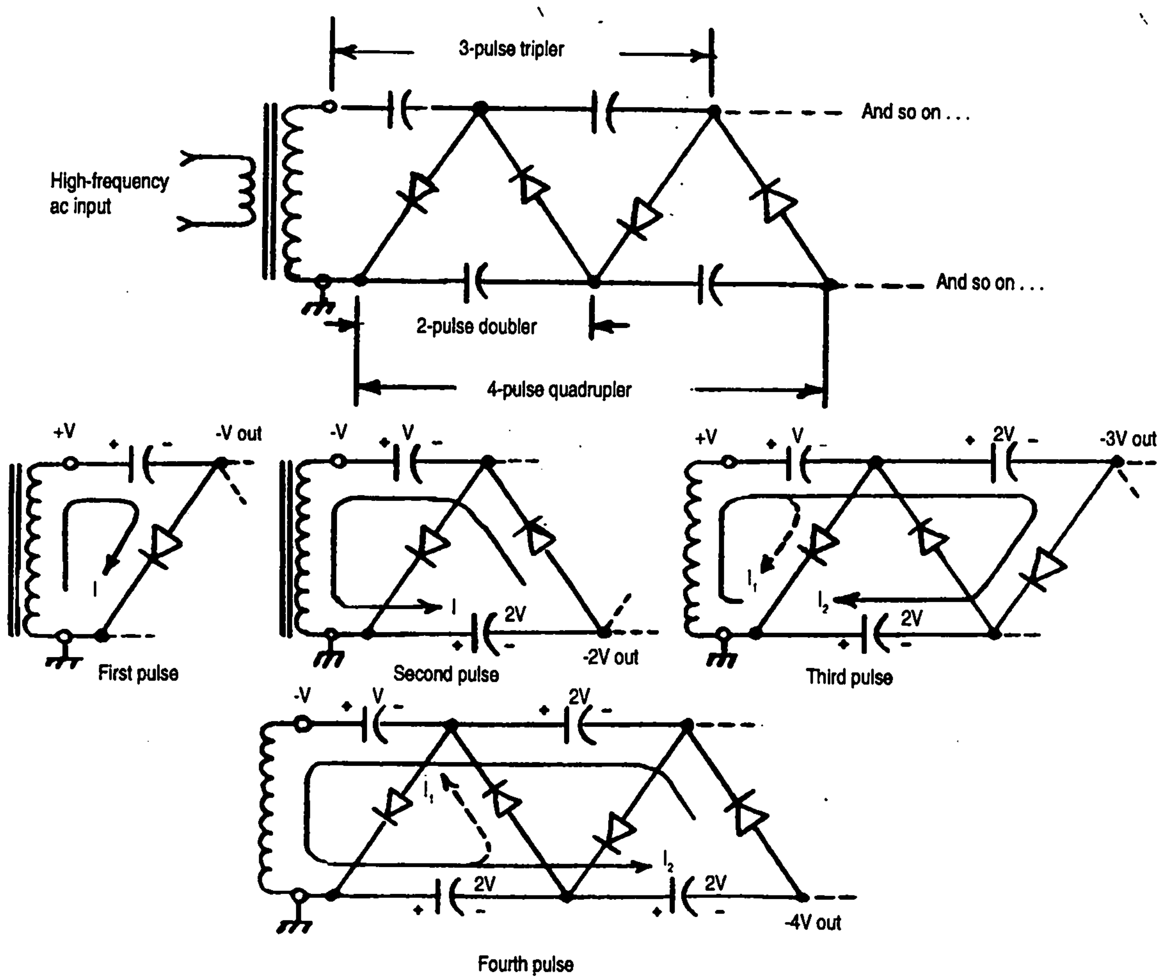


Figure 18-7. Multiple-phase voltage-multiplier type of rectifier circuits.

The theoretical multipliers described are less than practical because the successive capacitors must be much smaller than the preceding ones. Practical multipliers use nearly identical capacitors in each stage. Many more input "pulses" are required to fully charge all of the capacitors because of the voltage lost by early capacitors in supplying charge to later capacitors. In addition, there is a significant time lag between the time when the input voltage changes and when the output changes. This lag affects the bandwidth and transient response of a closed-loop regulation system. There are virtually as many types of voltage-multiplier rectifier circuits as there are conventional rectifiers, including full-wave center-tap, full-wave bridge, etc.

Voltage multipliers need not be restricted to single-phase input. Figure 18-8 shows a three-phase voltage-multiplier rectifier that is driven by three square-wave half-bridge inverters whose timing is such that the outputs of each are delayed by 1/3 of a master-clock period, thus producing the effect of three-phase ac. (There is no reason to be limited to three phases either, or multiples thereof; four- or five-phase systems are no less practical.) As previously noted, one of the advantages of polyphase rectification of ac having a sinusoidal waveform is that conduction is restricted to the peaks of the sinusoids, and the smaller the conduc-

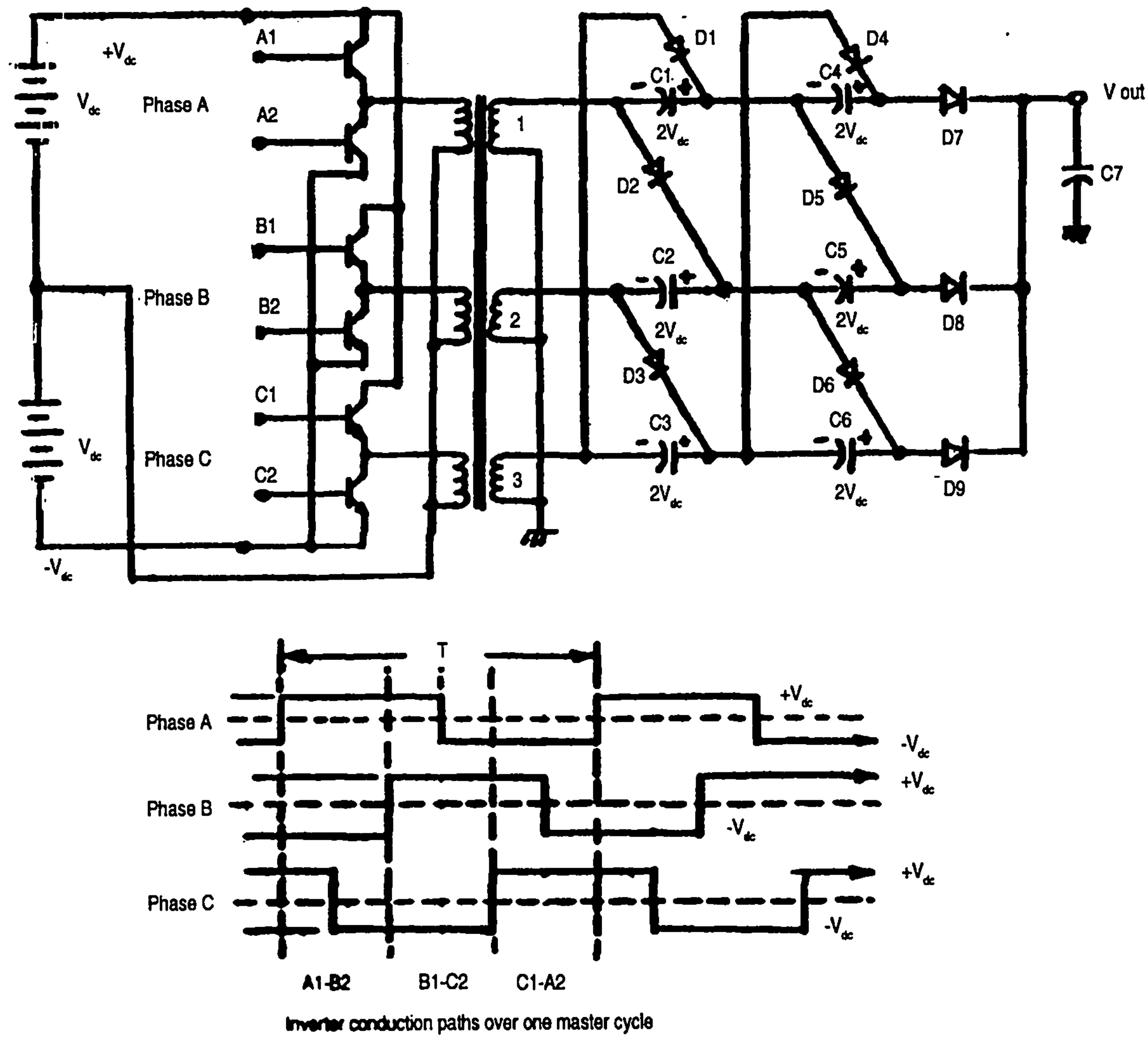


Figure 18-8. Three-phase voltage-multiplier rectifier circuit.

tion angle, the closer the unfiltered rectifier output is to pure non-varying dc. When the ac is produced by square-wave inverters, however, there is no such advantage to a polyphase system, because the tops of the ac input are nominally flat. The polyphase advantage, therefore, results from dividing up the total average current into multiple parallel, but time-sequential paths, that time-share the load.

In the practical example shown, the outputs of the inverters are connected to the primary windings of a wye-connected transformer, the "neutral" of which is connected to the common point, or ground, of the bipolar dc source, $+V_{dc}$ and $-V_{dc}$. Given the timing relationships shown, and starting with phase *A* positive (switch *A1* conducting), we see that phase *B* is negative (switch *B2* conducting). Secondary winding 1 will be positive by an amount proportional to V_{dc}, and secondary winding 2 will be negative, also by an amount proportional to V_{dc}. The actual voltages will depend upon the transformer turns ratio and other transformer and load-current factors. Conduction through diode *D2* will charge capacitor *C2* to an amount proportional to $2V_{dc}$, in the polarity shown. The conduction interval will have a duration of 1/3 of a master-clock period. For the next 1/3 period, phase *B* will be positive and phase *C* negative (switches *B1* and *C2* conducting). Capacitor C3 will charge to a voltage proportional to $2V_{dc}$ through *D3*. For the final 1/3 period, phase *C* is positive and phase *A* negative (switches *C1* and *A2* conducting), and *C1* will charge through *D1*. As the cycle continues, charge accumulated by *C1* will be shared with *C5* through *D5;* charge on *C2* will be shared with *C6* through *D6;* and charge on *C3* will be shared with *C4* through *D4* until all capacitors are charged to voltages proportional to $2V_{dc}$. The output capacitor *C7* will eventually charge to a voltage proportional to V_{dc} $(2n + 1)$, where n is the number of multiplier stages, or series capacitors, through summing diodes *D7, D8,* and *D9*. The summing diodes charge the output capacitor to a voltage that is greater than the sum of the series-capacitor voltages by an amount proportional to V_{dc}. In the case illustrated, $n = 2$, so the output voltage will be proportional to $5V_{dc}$.

The input-output characteristics of the voltage-multiplier rectifier are not unlike those of a transformer-rectifier system having a high transformer step-up ratio. The dc output voltage of the multiplier-rectifier can be many times the peak voltage of the ac source. In return, we can expect that the source current will be higher than the load current by the same factor—or more, if there are circuit power losses—simply by law of the conservation of energy. The same result is reached by recognizing that the capacitors are in series with respect to the load but in parallel with respect to the source. Charge delivered to the load over any period of time is removed from each capacitor simultaneously, but the charge must be replaced individually from the source. If charge Q is removed and there are n capacitor stages, a total charge of nQ must be supplied by the source over the same period of time to maintain steady-state equilibrium.

One particular form of the voltage-multiplier rectifier is often used to produce very high voltages in particle-accelerator applications. It is called the Cockroft-Walton generator, a type of which is shown in Fig. 18-9. This rectifier uses a full-wave center-tapped voltage multiplier that is formed by superimposing a mirror image on the half-wave type shown in Fig. 18-7 to produce a voltage step-up of 10. For this type of voltage-multiplier rectifier, output voltages up to one megavolt are common. At these high voltage levels, such generators have an inherent physical superiority over simple, high-voltage transformer-rectifier assemblies: insulation. Both types use about the same number of individual rectifier elements, but in the Cockroft-Walton generator the stages can be spaced and graded (using deck corona shields of adequate radius of curvature) so that air can provide sufficient dielectric strength even at such high voltages. The type illustrated can be of prodigious size. Even so, with a voltage step-up of 10, the peak input voltage must be 100 kV for an output voltage of 1 MV. Such generators typically operate at source frequencies of up to 10 kHz, with the input supplied by high-power audio amplifiers that use power vacuum tubes as output devices. For an average-output current of 10 mA at 1 MV, the average power input must be at least 10 kW (and the reactive VA may even be higher).

18.4 High-voltage dc power supply using a quasi-resonant inverter

The inverters discussed so far have had square-wave outputs, or rectangular-pulse outputs if it is pulse-width modulated. These waveforms require the abrupt "chopping" of a dc input. This activity introduces significant repetitive transient pollution that must be suppressed, and it produces corresponding stress on the electronic switches that must do the chopping. The switches, moreover, must either have full-control capability or they must be turned off by "forced-commutation" involving additional components and switching-speed limitations. The quasi-resonant inverter, which was introduced in Fig. 9-45, does not share these problems. A high-voltage, off-line, dc power supply that uses such an inverter along with a high-step-up multiplier-rectifier, is shown in Fig. 18-10.

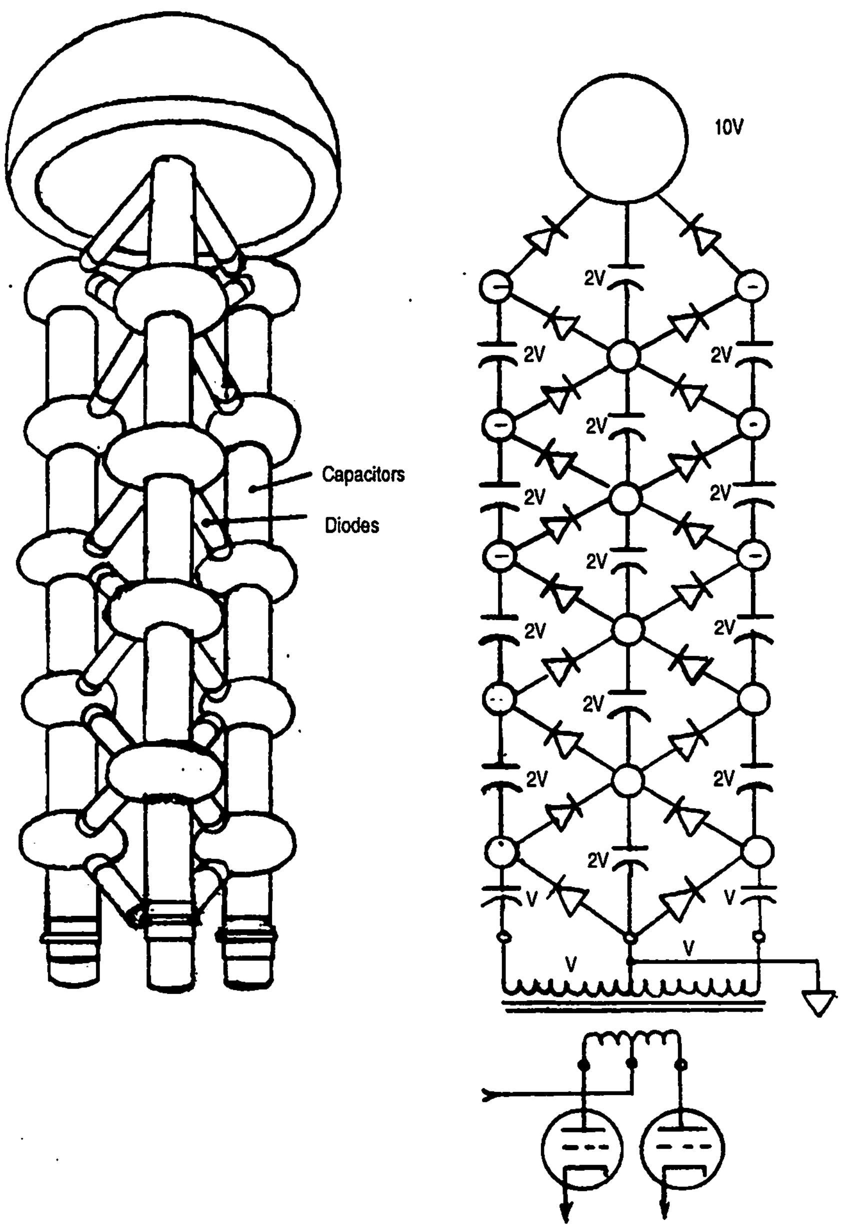


Figure 18-9. The Cockroft-Walton generator for very high voltages.

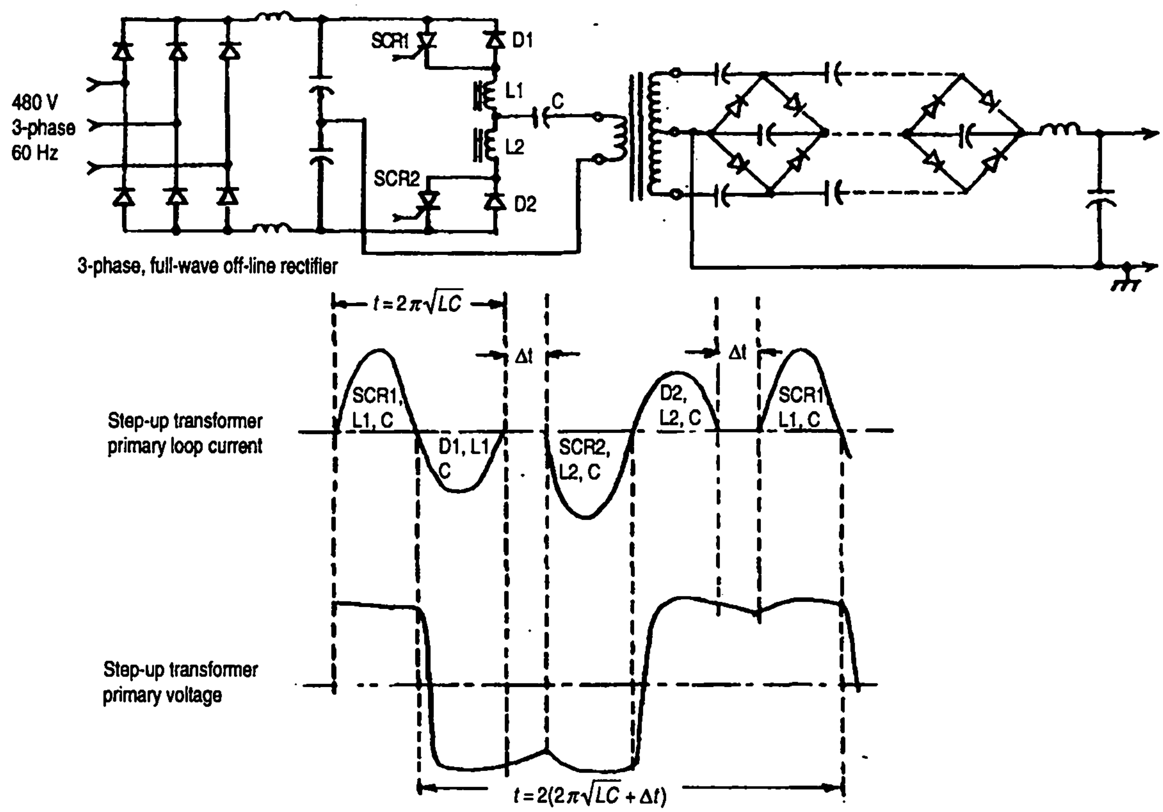


Figure 18-10. High-voltage switch-mode power supply using resonant-type inverter and full-wave bridge-type voltage-multiplier rectifier.

A dc source is created by direct three-phase, full-wave (six-pulse) rectification of the line, producing two balanced dc rails. A neutral return for the half-bridge inverter is provided by a center-tapped capacitor filter at the rectifier output. Power transfer from the source to the load occurs during the half-cycle conduction intervals of the two SCR power switches, *SCR1* and *SCR2*. The conduction path for *SCR1* is in the positive direction through *L1*, *C*, and the rectifier-transformer primary winding. The conduction path for *SCR2* is in the negative direction through *L2*, *C*, and the transformer primary winding. Inductors *L1* and *L2* are of equal inductance. The values of *L1*, *L2*, and *C* are chosen to provide two important properties of the overall circuit. First, the characteristic impedance of the discharge circuit ($\sqrt{[L/C]}$), where *L* is either *L1* or *L2*, is chosen so that it is considerably smaller than the value of resistance that is reflected back from the rectifier and load into the transformer primary. Therefore, the series-resonant circuit is underdamped and will ring. The current will have a damped sine-wave shape. When the current through *SCR1* passes through zero, *SCR1* will be automatically commutated off. The negative-going half-cycle following *SCR1* conduction will flow through *D1*. When current reaches zero again, it will stop until *SCR1* is gated on again. The same sequence occurs in the opposite direction with *SCR2* and *D2*. The values of *L* and *C* also determine the period, or oscillation frequency, of the discharge current. The parameter that allows variability (or regulation) of output voltage is the controllable time interval, Δt, between each power-transfer cycle. Even though the transformer primary currents are sinusoidal segments, the primary voltage is "rectangularish" because of the hold-up produced by the reflected load-circuit capacitance.

Two significant differences are apparent between the square-wave and the quasi-resonant inverter. The first is that the peak switch current for the quasi-resonant inverter is several times greater than the load current (typically 5 to 6 times as great). This is necessary for the primary circuit to perform in underdamped, self-commutating fashion. This circulating current is not provided by the primary power source. It contributes to inefficiency only in that the ohmic losses in the inductor and transformer primary will be proportional to the square of the circulating current. On the other hand, the switch only needs to have half-control capability. It plays no role in determining either the amplitude or duration of the current, which, by virtue of its sinusoidal wave shape, has no sharp discontinuities, either. The second difference is that the quasi-resonant inverter, in order to have control over load voltage, operates at variable frequency rather than constant frequency and variable pulse duration.

The rectifier shown is the same full-wave, center-tapped type used in the Cockroft-Walton generator. At least one practical version of this circuit uses a multiplier with a step-up ratio of 20, giving an output voltage of 75 kVdc with a transformer turns ratio of 21 and 480-V, three-phase input. Like the Cockroft-Walton generator, the power supply is entirely air-insulated, thus taking advan-

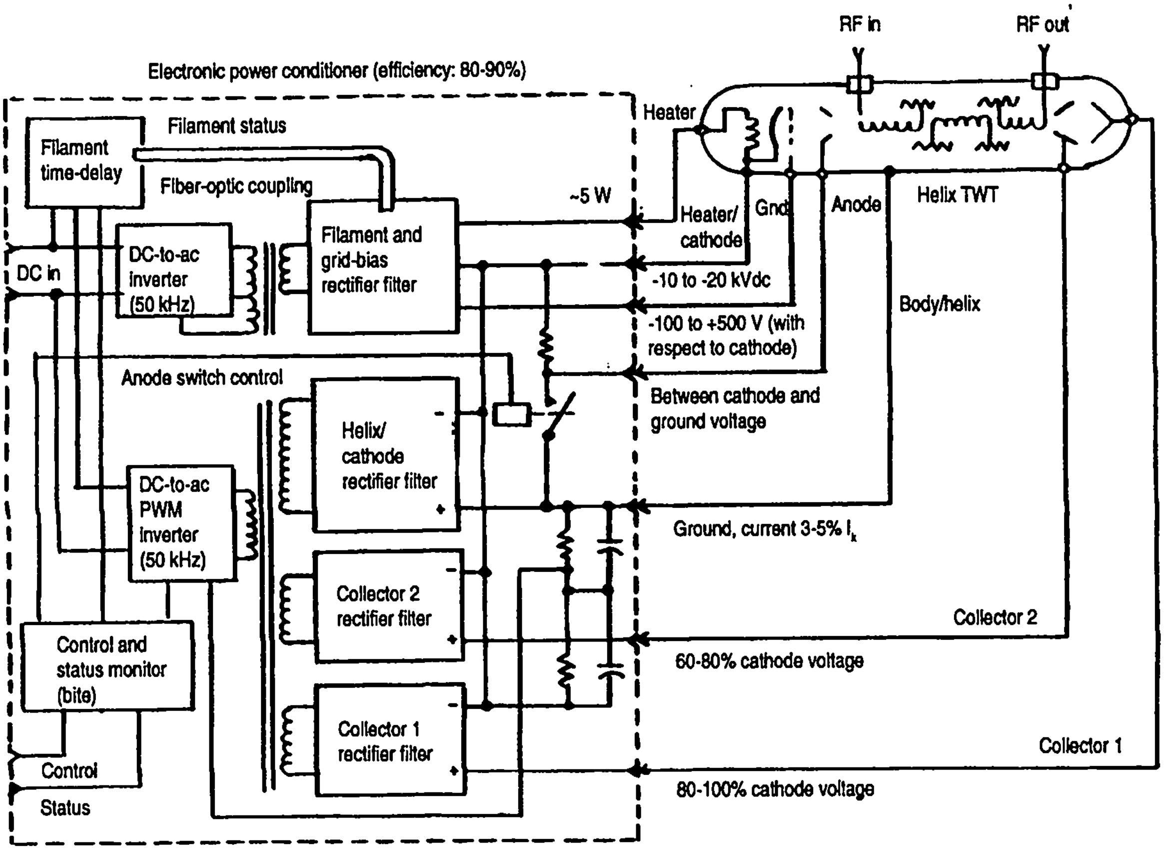


Figure 18-11. Block diagram of traveling-wave-tube amplifier (TWTA) using high-frequency switch-mode electronic power conditioning (EPC).

tage of the large dimensional spacings possible with voltage-multiplier rectifiers.

18.5. The electronic high-frequency power conditioner as applied to a microwave tube

The most common microwave-tube application of the high-frequency electronic power conditioner (EPC), is in the traveling-wave-tube amplifier (TWTA), a generic block diagram of which is shown in Fig. 18-11. Such TWTAs abound in aircraft and satellite environments where size and weight are of primary concern. Many such EPCs operate from so-called 28-Vdc aircraft primary power (which has a specified minimum value of 19 Vdc). There is a practical upper limit to the amount of power that can be converted from such an input bus, somewhere around 1 kW. (Higher-power can be extracted if the input bus is 208-V, three-phase, 400-Hz because direct rectification will yield internal dc rail voltage of almost 300 Vdc, from which MOSFET inverter switches will operate more optimally, inverting higher power at higher efficiency.)

A TWTA will have no fewer than two independent, high-frequency inverters that operate at the highest practical clock rate, usually between 10 kHz and 100 kHz. The higher the voltage and power levels, the lower the clock rate is likely to be for overall optimum performance. The first inverter supplies the input to the filament and grid-bias transformer-rectifier circuit (and to the grid modulator, if it is the direct-coupled type). After this inverter has been turned on, there will often be a read-back of filament voltage and current and grid bias voltage. These values are usually summed as a single optically coupled status signal that is sent to the ground-level control electronics, which will include a filament time-delay circuit. The filament and grid-bias circuits float at the TWT cathode voltage, which is the output of the helix/cathode transformer-rectifier. This voltage can be anywhere from a few kilovolts to a few tens of kilovolts, but it is most usually in the 3-kV to 5-kV region. Its input comes from the second high-frequency inverter, which is turned on only after the requisite cathode-heating time-delay has elapsed. The helix/cathode or cathode/ground voltage is the highest and must be the most precisely controlled voltage in the system. A sample of this voltage is ordinarily used as the input to the PWM regulator for this inverter. The TWT used in such amplifiers will almost always have a depressed collector or several depressed collectors (as many as five stages of depressed collector have been featured in some high-power space-based TWTs). Each collector will have a separate transformer-rectifier, or at least a rectifier operated from a separate secondary winding that shares the same primary and core. In a two-stage depressed-collector TWT, as illustrated, the Collector-1 voltage may be from 80% to 100% of cathode-ground voltage (0 to 20% depression) and the Collector-2 voltage may be from 60% to 80% of cathode-ground voltage (20% to 40% depression). The cathode-ground rectifier may handle as little as 3% to 5% of the total beam current, with the remainder being split between the grid and the multiple collectors.

Some TWTs, especially those with coupled-cavity circuits, may be endangered if beam current flows before beam voltage reaches the nominal operating value during turn-on. (If the tube is pulse modulated, this is not an additional problem. Pulses to the modulator can be simply inhibited until beam voltage is at the

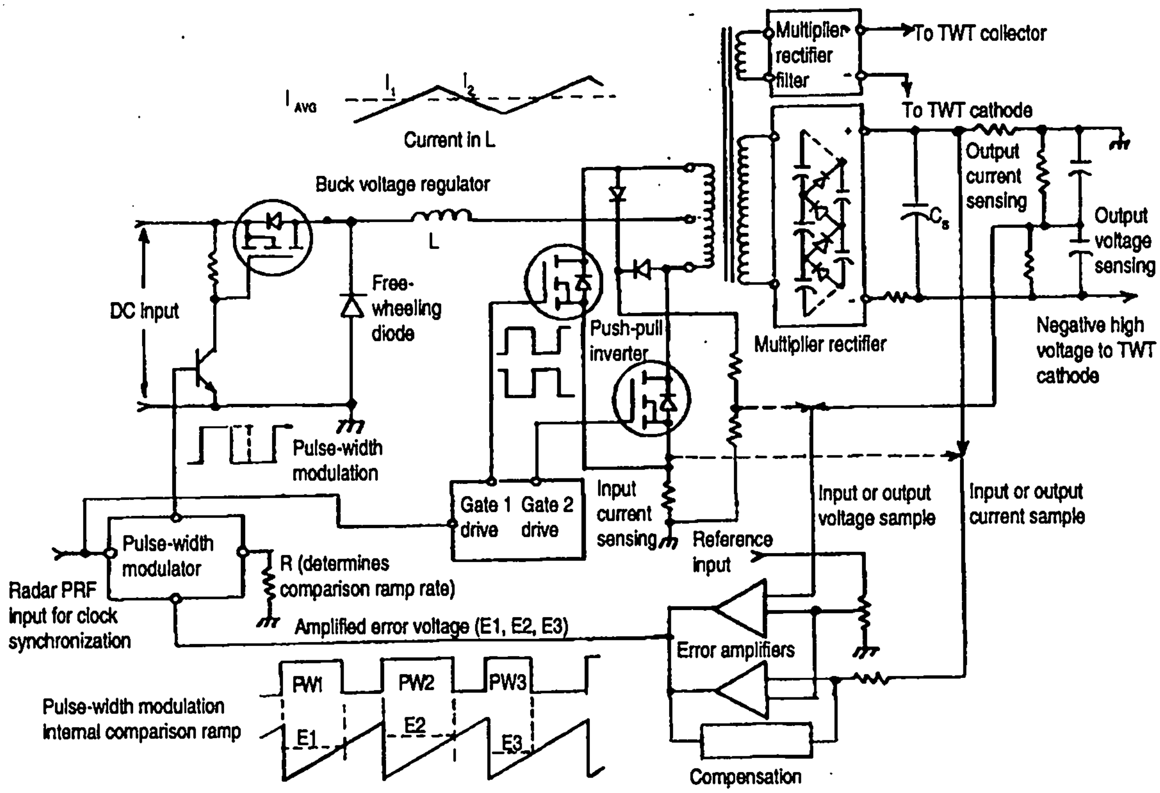


Figure 18-12. Practical TWTA switch-mode power conditioning using a buck regulator and push-pull inverter.

proper value.) A tube intended for continuous duty (CW) may still require some means of inhibiting beam current during turn-on. Often a modulating or isolated anode is provided, as shown. It is held at cathode potential by a series resistor. When cathode voltage reaches the proper value, the anode is connected to ground by means of a high-voltage but low-current relay, which turns on beam current.

A practical implementation of the high-voltage dc portion of an EPC in a TWTA is shown in Fig. 18-12. The power inverter is the push-pull type. Each MOSFET switch operates at 50%-duty factor. Output voltage control is accomplished by means of a buck-type switch-mode regulator that is directly in series with the dc input source. This source is pulse-width modulated to regulate or limit either the input voltage and current to the step-up transformer or the dc output voltage and current, depending upon which sets of sample points are connected to the dual error amplifiers. The error amplifiers compare the sampled signals with a reference voltage, thus developing an amplified error signal at the parallel-connected outputs. The error signal is compared with a sawtooth-shaped voltage in the pulse-width modulator called the comparison ramp. The ramp voltage always starts upwards at the same voltage and resets to that voltage at the end of each clock period. The buck regulator is turned on at the beginning of each clock period, coinciding with the start of the ramp. When the ramp voltage exceeds the amplified error voltage, the drive to the regulator is terminated. For instance, if the error signal is *E1*, the duration of the regulator drive will be *PW1*, if *E2*, it will be *PW2*, and if *E3*, it will be *PW3*, as illustrated. This process is often called delta modulation, which converts voltage change to pulse-duration change.

The particular circuit described was used in a multi-purpose radar/communications transmitter. It utilized a high-performance TWT that was required to operate with low-Doppler sidelobes, which meant very-low-ripple sidebands. One clever trick used to achieve this requirement was to synchronize the inverter clock with a multiple of the radar repetition rate. This effectively "hid" the power-supply-inverter ripple components under the PRF lines of the pulse spectrum. At constant PRF, this technique would work fine. In this particular application, however, PRF was not constant. In fact, it changed frequently under computer control. In a conventional delta modulator, the ramp rate is constant. Therefore, if the clock rate is changed, there will be an immediate increase or decrease in buck-regulator duty factor, depending upon whether the system PRF went up or down. This change in the clock rate causes a corresponding transient increase or decrease in high-voltage dc output. The feedback will eventually correct the output error but not very quickly. The closed-loop bandwidth of such a regulator is only tens of hertz. In actual operation, the regulator was continually hunting for the proper output voltage because PRF changes happened with a frequency not much lower than the closed-loop bandwidth. The solution, which was successfully implemented, was to change the ramp rate as well as the clock rate, thus making the regulator duty factor independent (almost) of the clock rate.

The value of the output inductor, *L*, of the buck regulator must also be carefully chosen. Its purpose is to smooth the discontinuous output of the regulator. However, if it is too large and does the job too well, it severely limits the response time of the regulator loop—which isn't that fast to begin with. The current in the inductor will rise (*I1*) when the regulator is turned on and fall (*I2*) when the regulator is turned off. A peak-to-valley variation of 10% for this type of circuit is considered optimum by many designers.

18.6 An idea too clever not to share (It's not mine, by the way.)

As mentioned before, the leakage inductance of either a step-up or a step-down isolation transformer that is used in a switch-mode power supply imposes a fundamental limitation on the circuit's optimum switching frequency. However, there is a power-supply design covered by US patents that, instead of fighting leakage inductance, actually makes use of it in reaching switching frequencies of up to a megahertz. This design does not appear in high-voltage, or especially high-power, products, but the products it can be found in are nevertheless quite useful in many transmitter applications.

A simplified schematic diagram of the power-transfer portion of a typical circuit, along with its voltage and current waveforms, is shown in Fig. 18-13. The circuit comprises a single-ended MOSFET power switch; a transformer, *T*; a charging diode, *D1*; a storage capacitor, *C*; a free-wheeling diode, *D2*; a smoothing inductor, *L*; and the useful load, shown as a resistor. The leakage inductance of the transformer, $L_{LEAKAGE}$, performs what should be a familiar role. Starting at the beginning of an energy-transfer cycle when the FET switch is first turned on, current will flow in the primary and secondary windings of the transformer and through diode *D1*. The current waveform through *D1* will be in the form of a

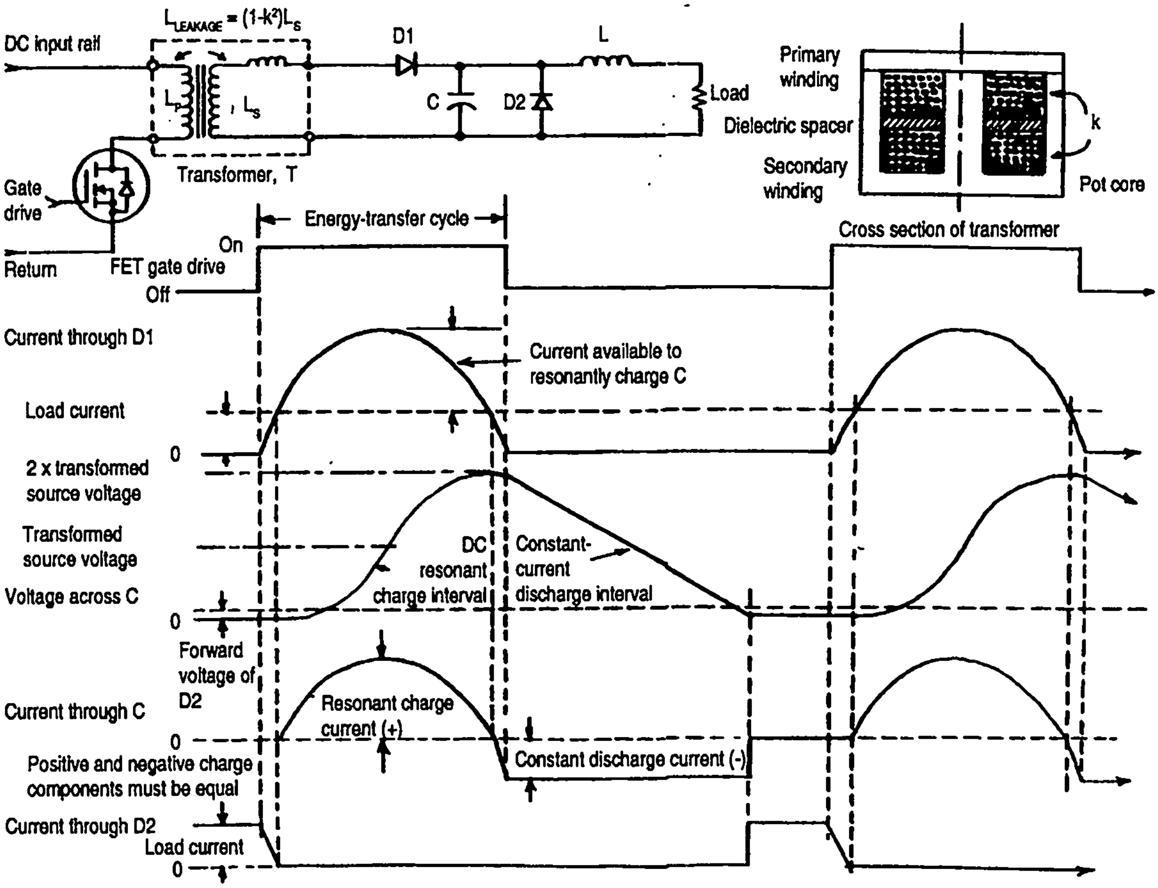


Figure 18-13. Megahertz-rated quasi-resonant (dc-resonant charging) switch-mode power supply.

half sine-wave, the frequency and amplitude of which are determined by $L_{LEAKAGE}$ and C. Assuming that steady-state load-current conditions had already been reached, a steady value of current will flow in *L* and the load. This current is either replaced or maintained by the paths through either *D1*, *C*, or *D2*. As soon as the rising current through *D1* exceeds the constant value of load current, there will be an excess of current available. This current charges capacitor C in a dc-resonant fashion that is almost identical to the dc-resonant charging of the line-type modulator PFNs previously discussed. Except in this case, $L_{LEAKAGE}$ performs the role of the charging inductor in the line-type PFNs. But it should be noted that a conventional transformer may not have sufficient leakage inductance to optimize this circuit. Therefore, the transformer, as shown in the simplified cross-section drawing, must have primary and secondary windings that are deliberately separated from one another on a pot core by a dielectric spacer. This is done to decrease the coefficient of coupling, *k*, and increase the leakage inductance, which is proportional to $(1 - k^2)$. This strategy also enhances the circuit's voltage-hold-off capability and decreases the capacitance between input and output terminals, thus making the circuit more like a true four-terminal network and more useful in many transmitter-circuit applications requiring "floating" power sources.

Once the current through *D1* approaches the end of its half-sine-wave existence, it falls below the value of the load current. This terminates the resonant charging of C and starts its constant-current discharge. Before that begins, however, C has been charged to almost twice the value of the transformed source voltage through the dc-resonant charging action. The constant-discharge current is the load current, which is kept constant by L. The voltage across C will linearly decrease, pass through zero, and reverse polarity until diode $D2$ is forward-biased. Current will continue through $D2$ until the next power-transfer cycle commences. Note that the behavior is quasi-resonant and that a half-control switch could do the job. However, no such switch will operate at a megahertz rate, which this circuit can reach, so a full-control switch is used, thus suffering no first-order switching losses. A fixed amount of energy is transferred to the load circuit during each energy-transfer cycle, so load voltage and current control is accomplished by the modulation of the transfer-cycle recurrence rate, as in other quasi-resonant designs.

This circuit is another example of a pace-setting concept based on a time-honored process.