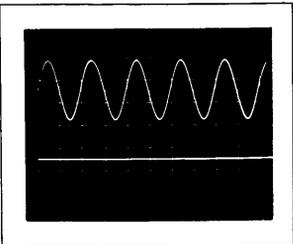
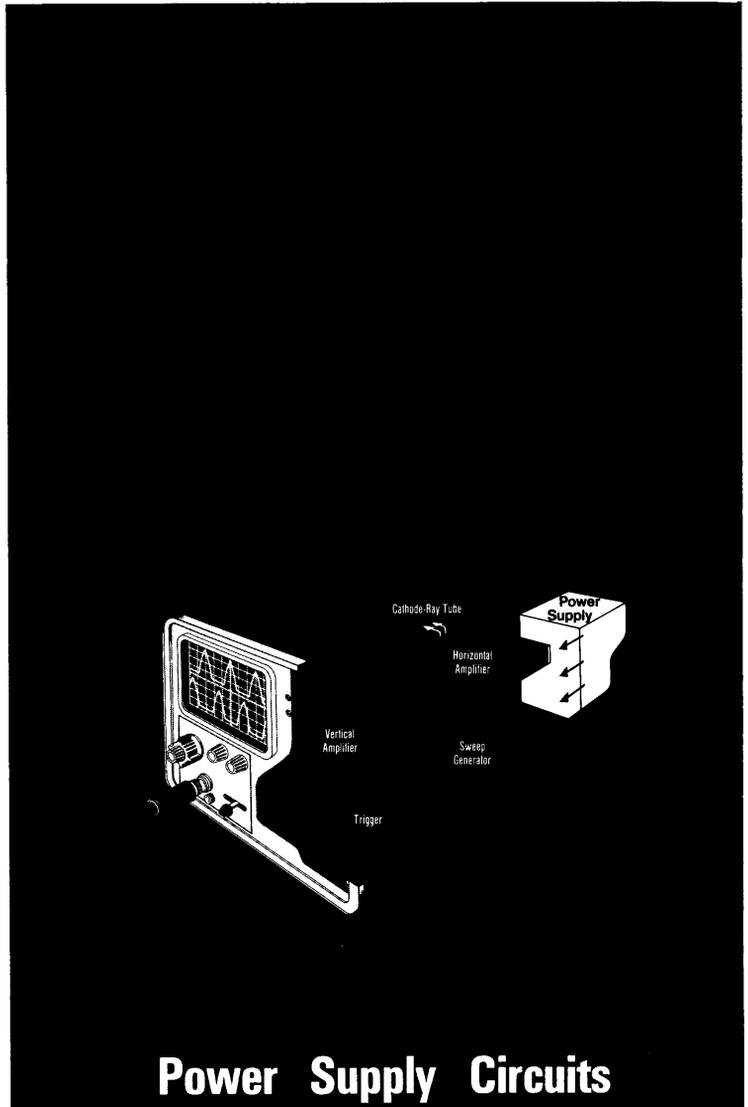




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Circuit Concepts



SECOND EDITION

Power Supply Circuits

POWER SUPPLY CIRCUITS

BY
KENNETH ARTHUR

Significant Contributions
by
JOHN THOMPSON



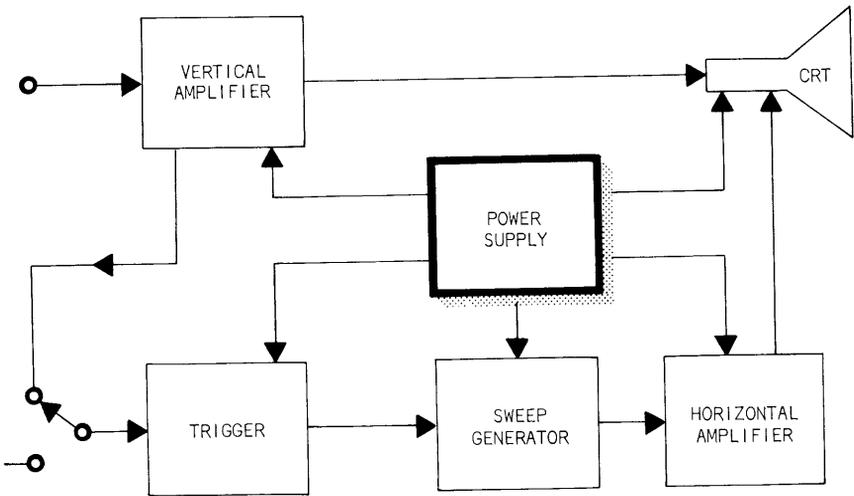
CIRCUIT CONCEPTS

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CONTENTS

	INTRODUCTION	1	
1	RECTIFIERS AND FILTERS	3	
2	VOLTAGE REGULATORS	17	
3	THE PRLV POWER SUPPLY	83	
4	AC-DC POWER SUPPLIES	85	
5	HIGH-VOLTAGE POWER SUPPLIES	109	
6	POWER SUPPLY HARDWARE	127	
	INDEX	147	
	CIRCUIT CONCEPTS BY INSTRUMENT		149



INTRODUCTION

The active circuits of electronic equipment operate on DC potentials, which must usually be derived from conventional AC line voltages. Even when DC line or battery power is available, it must be processed and regulated before being applied to electronic circuits. Almost all electronic measuring instruments, therefore, require a *power supply* to convert locally available power to DC operating potentials. Power supplies are generally included as an integral part of an instrument, but also exist as self-contained units.

Power supply performance has a direct bearing on the caliber of any electronic measuring instrument. Precise and accurate measurements demand *repeatability* of performance; that is, a given event must evoke an identically proportional response in the instrument each time that event is subjected to examination. The circuits responsible for the instrument's response are designed to operate with specific plate (collector), bias, and reference voltages. If these voltages are permitted to fluctuate beyond strictly defined limits, repeatability is lost and accurate measurement becomes difficult or impossible.

Because of its central role in the operation of electronic instruments, the power supply offers a logical point of entry to the study of Tektronix circuit concepts. The objective of this book is to set forth the basic principles and practical considerations governing the operation and design of Tektronix power supplies. The reader is assumed to have only a limited background in electronic theory. Therefore, an operational, rather than mathematical approach to circuit descriptions, is employed. A concept-to-instrument cross-index is included as an aid to the study of specific instruments.

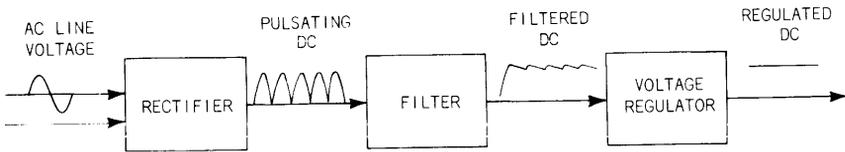


Fig. 1-1. Power-supply block diagram.

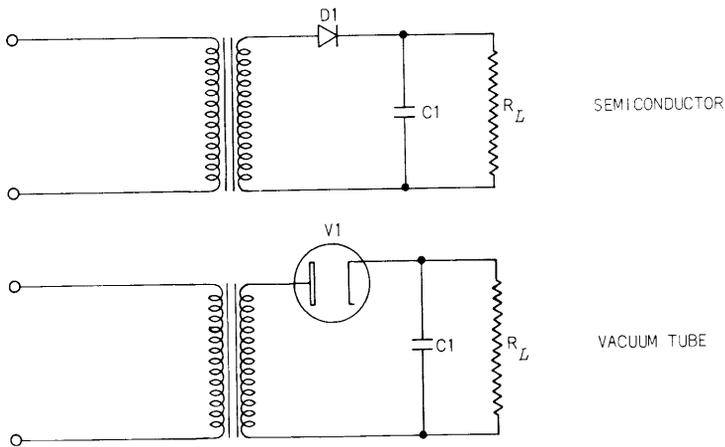


Fig. 1-2. Half-wave rectifiers.

1

RECTIFIERS AND FILTERS

The primary function of any power supply is to convert AC power to useful DC potentials. This process requires at least two, and usually three operations. These are (1) rectification, (2) filtering, and (3) voltage regulation. The circuits which perform these operations are quite understandably called *rectifiers*, *filters*, and *voltage regulators* respectively. Fig. 1-1 is a functional block diagram, showing the relationship between these circuits.

The rectifier first converts the AC source voltage to a "pulsating DC" voltage. This train of pulses can be viewed as a DC voltage, with an AC component. Most of the AC component is removed by the filter circuit. In some cases the filtered output can be utilized without further processing, but usually it is subjected to some type of voltage regulation before being routed to the other circuits of an instrument. The block diagram shows the voltage regulator as a separate circuit following the rectifier. This is the conventional arrangement for the majority of low-voltage, general-purpose power supplies. It should be noted, however, that voltage regulation is not always an isolated function and may be incorporated in other circuits of the power supply.

half-wave
rectifier

The operation of the rectifier and filter circuits is so closely related that the two circuits will be examined as a unit, beginning with a simple half-wave rectifier and a capacitive filter (Fig. 1-2). The rectifier limits the alternating current to one direction of flow through the load resistance. The filter capacitor stores energy and releases it between alternations, thereby maintaining a fairly uniform current through the load resistance. The transformer establishes the nominal level of the rectifier output by "stepping" the AC line voltage up or down to the

desired level. (Some inexpensive radios and TV sets eliminate the transformer entirely, and obtain their operating voltages through the use of voltage multipliers. These circuits are discussed in the chapter on high-voltage power supplies.)

The purpose of the diode is to allow current to flow freely in one direction, while opposing it in the other. The vacuum-tube diode was first employed in this role, and is still best suited for the development of high voltages. However, the parameters of the silicon power rectifier (diode) have been steadily extended in recent years, so that now these small, relatively inexpensive devices have largely supplanted the vacuum tube in modern power-supply circuits. Germanium diodes are also used in some rectifiers where their very low forward impedance assists in achieving good power-supply regulation; but their susceptibility to reverse-voltage breakdown limits their application to low-voltage rectifiers.

In operation, electrons flow through the secondary winding, diode, and load resistance when the secondary voltage has the polarity shown in Fig. 1-3A. As the secondary voltage reverses its polarity, the diode cuts off. In the meantime,

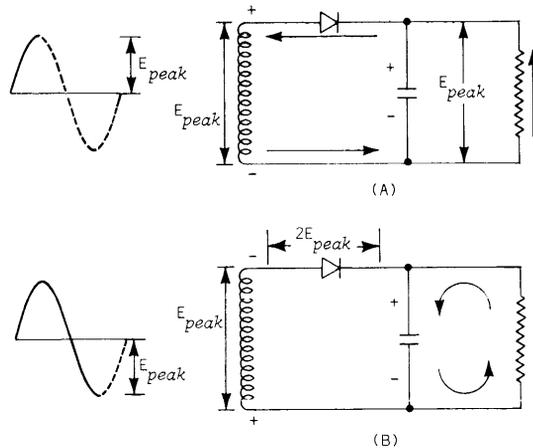


Fig. 1-3. Current and voltage in half-wave rectifier.

the filter capacitor has charged to approximately the peak value of the secondary voltage. As soon as the secondary voltage falls below the peak value, the capacitor begins to discharge through the load resistance. At the peak of the negative alternation, the reverse voltage across the diode is approximately twice the peak input voltage, since the voltage on the secondary and the charge on the capacitor are series aiding (Fig. 1-3B). This value is known as the peak inverse voltage (PIV). Power rectifiers are rated by the maximum PIV they are able to withstand without damage.

When the secondary voltage has completed its negative excursion and rises to a point where it exceeds the charge on the capacitor, electrons again flow in the secondary circuit.

The resultant waveform is shown in Fig. 1-4. The solid line represents the output voltage waveform; the variations in its amplitude are known as "AC ripple." Note that all the current given up by the capacitor between times T_1 and T_2 must be replenished in the interval T_2 - T_3 . Thus the amplitude of ripple current varies inversely with frequency and directly with load current.

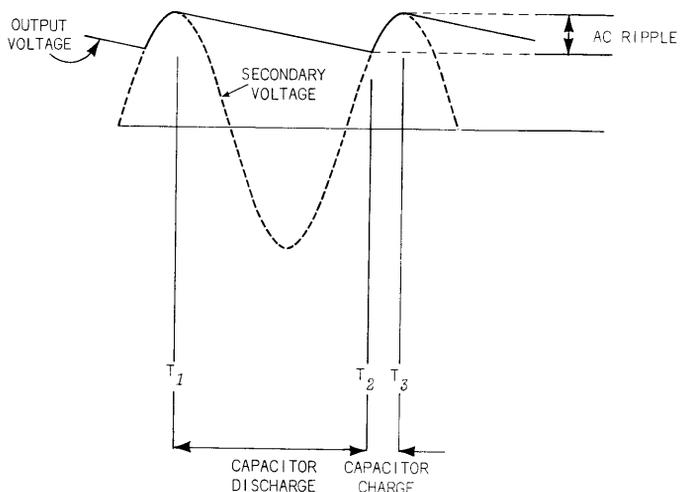


Fig. 1-4. Filtered half-wave rectifier output.

ripple
current

Ripple current in the filter capacitor generates heat, due to the unavoidable resistive component in the capacitor's impedance. This heat must be radiated by the walls of the capacitor. The rate of radiation depends on the area of the radiating surface and the temperature of the surrounding air (ambient temperature). As the temperature of the capacitor rises, its internal resistance rises. Eventually, if a state of equilibrium between the rate of heat generation and radiation is not attained, the capacitor will go into "runaway" and destroy itself. Therefore, most electrolytic filter capacitors carry specifications for maximum ripple current at a given ambient temperature.

diode
ratings

It might appear that increasing capacitance would reduce AC ripple to negligible proportions. However, as capacitance increases so does the amplitude of the charging current. Remember that all the current drawn by the load must be replaced in the charging interval. These large currents may eventually exceed the "recurrent-peak-current" rating of the diode. To handle both these situations, a small amount of resistance can be inserted in the capacitor charging loop to prevent damage to the diode, as illustrated in Fig. 1-5. The resistor may be inserted at point A or B, but, *not* at point C, since it would then also exist in the capacitor's *discharge* loop, and thus tend to increase the AC ripple voltage.

Small, inexpensive carbon-composition resistors are usually used as current limiters. They perform the additional task of protecting the transformer from catastrophic overloads by quickly burning out when subjected to excessive current, much like a conventional fuse.

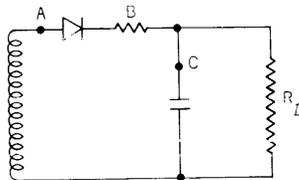


Fig. 1-5. Current-limiting resistor.

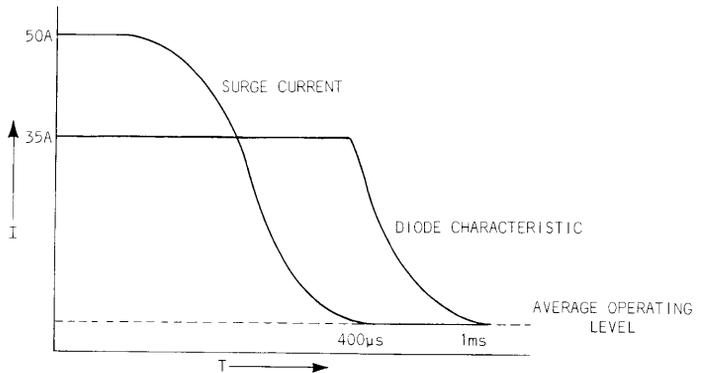


Fig. 1-6. Typical silicon-diode surge-current tolerance.

Also, when power is first applied to the rectifier, the capacitor appears almost as a short circuit, and a high current flows through the diode for a very short period. This "surge" current may exceed the recurrent-peak-current rating of the diode by a wide margin; however, because of its short duration and the fact that it occurs only at the instant of turn-on, it does not necessarily damage the diode. The surge-current tolerance of most power rectifiers, (diodes specially suited for rectifier applications) is stated in maximum current over a specified time interval. The ideal diode would have a surge-current tolerance matching the typical surge-current waveform shown in Fig. 1-6. Unfortunately, the actual tolerance curve is usually lower in amplitude although longer in duration, than the ideal.

The design of a rectifier circuit thus represents a compromise between several conflicting factors. Although AC-ripple amplitude varies inversely with the size of the filter capacitor, the practical limits on capacitance are determined by the characteristics of the diode and the amount of current-limiting resistance required for diode protection.

The half-wave rectifier operates satisfactorily with very small load-current demand; but if a higher output current is required, and passive filters alone are used, the only way to prevent excessive AC ripple is to increase the *frequency* of the input-voltage alternations.

Another disadvantage inherent in the half-wave rectifier configuration is that the load current in the secondary winding always flows in the same direction, even though it rises and falls at the line frequency. It therefore has a DC component which tends to saturate the magnetic core material (see Transformers, page 127). This effect complicates the design of half-wave rectifier transformers to some extent.

The most obvious shortcoming of the half-wave rectifier is its failure to take advantage of both alternations of the AC line voltage. To improve circuit performance a method of charging the filter capacitor twice every cycle is needed. This need led naturally to the development of the *full-wave rectifier*.

full-wave
rectifier

The full-wave rectifier may be regarded as a pair of half-wave rectifiers operating on opposite half-cycles. Such an arrangement is shown in Fig. 1-7. The black dots adjacent to the ends of the transformer windings indicate points of phase identity; that is, at any given instant the voltage at these points has exactly the same phase.

When phased as shown in Fig. 1-7, the two secondary windings can as easily be considered as a single winding with a center-tap. Thus points a and b may be connected and one of the terminals eliminated (Fig. 1-8A). Also, since both capacitors are connected to a common terminal and charge to the same maximum voltage (E), their opposite sides are at all times at the same potential and may also be connected as in Fig. 1-8B. Now the circuit can be redrawn, as in Fig. 1-8C, substituting a single filter capacitor for the two parallel capacitors and a single load resistor for the parallel load resistor. This is the standard configuration of a typical full-wave rectifier. Electrons flow in alternate halves of the secondary winding, charging the filter capacitor on each half-cycle of the primary voltage, thereby achieving the object of the new arrangement.

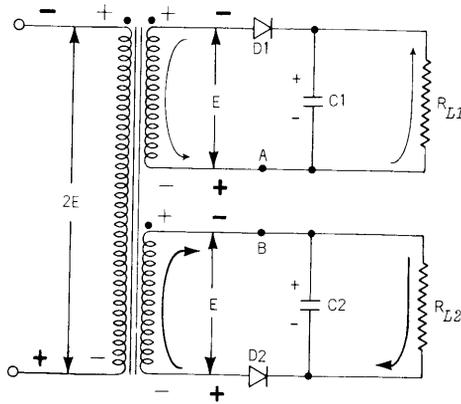


Fig. 1-7. Evolution of full-wave rectifier -- I.

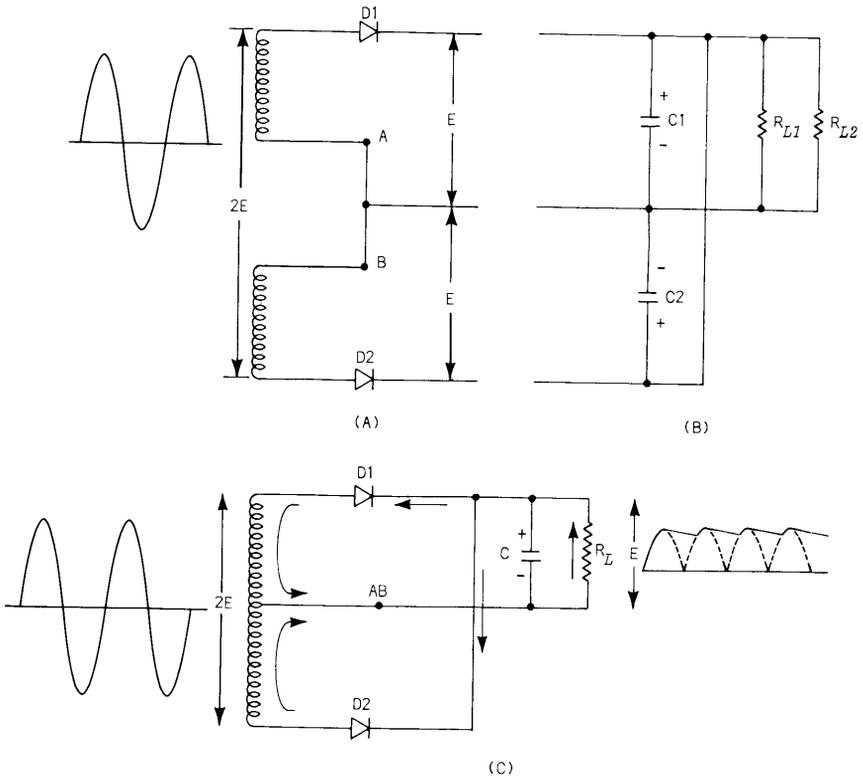


Fig. 1-8. Evolution of full-wave rectifier -- II.

There are some disadvantages inherent in this circuit, in spite of the improvements effected. Instead of an output of E volts for a corresponding E volts on the secondary winding, as was attained by the half-wave rectifier, a secondary voltage of $2E$ volts is now required. This requirement complicates manufacture of the transformer for two reasons. First, for a given total secondary *power* in a transformer, the higher the secondary voltage, the more costly the transformer will be to manufacture, due to insulation requirements. Secondly, connecting the center-tapped secondary winding requires a separate operation in the winding process.

The advantages offered by the full-wave rectifier were most attractive when vacuum tube diodes were the only practical rectifying component available. At this time, the area of primary concern was current, which was limited by the emissive properties of the vacuum tube's hot cathode. PIV was of little concern, since vacuum tube diodes are typically rated in the kilovolt region. Although the PIV on the diodes of the full-wave rectifier equals twice the output voltage, as in the half-wave rectifier, the twofold increase in current achieved by the new circuit gives it a distinct advantage. In addition, because the currents in the two halves of the secondary winding flow in opposite directions, the problem of core saturation is completely eliminated. In positive-output supplies, which represent a large majority of those in common use, still another advantage is gained by using a twin diode, whose cathode requires only one heater supply (Fig. 1-9).

The full-wave rectifier is obviously well suited for vacuum-tube construction and is still quite popular in modern electronic equipment.

bridge
rectifier

The appearance of solid-state diodes posed new problems for the power-supply designer. Their small size and weight, relatively low cost, and lack of heater-voltage requirements made them highly attractive as rectifier components. However, the current and voltage limitations of solid-state diodes are in direct contrast to those of the vacuum tube. High current ratings in semiconductor diodes are rather inexpensively attained, while high PIV ratings are quite difficult to attain and therefore expensive. A silicon diode, for instance, rated at one or two amperes costs very little more than one rated at 500 mA. The same type of diode, rated at one or two kV, however, costs a great deal more than one rated at a few hundred volts. Therefore, the aim in designing a rectifier using semiconductor diodes was to reduce the PIV on the diodes, rather than to increase the rectifier's current output. Also, because semiconductor diodes require no heater voltages and are relatively inexpensive, they could be used more generously than would be practical in a vacuum-tube circuit.

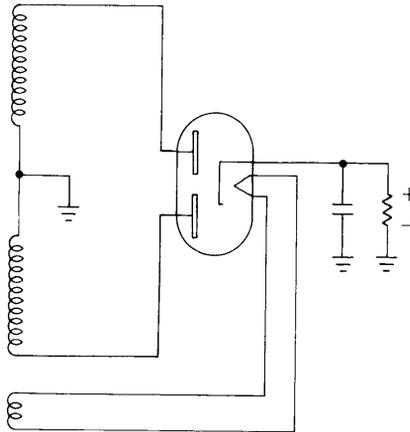
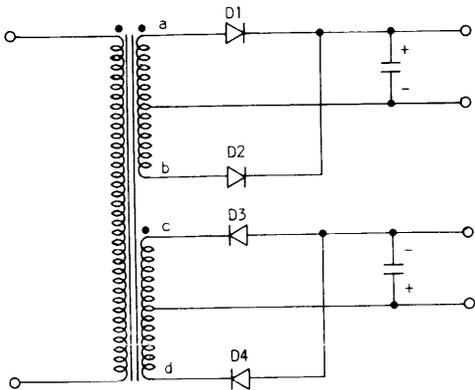
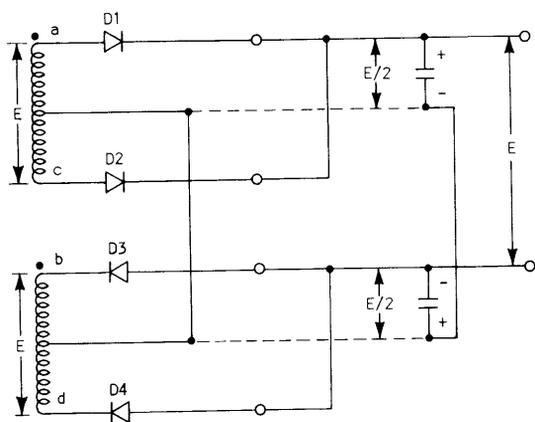


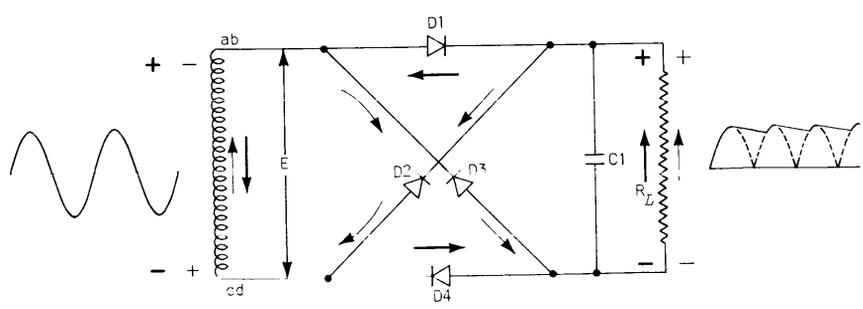
Fig. 1-9. Typical full-wave rectifier.



(A)



(B)



(C)

Fig. 1-10. Evolution of typical bridge rectifier.

These considerations led to the next advance in the history of rectifier development, a circuit combining the advantages of the half-wave and full-wave rectifiers. That is, the high ripple frequency of the full-wave circuit and the relatively low transformer-to-output-voltage ratio of the half-wave circuit were accomplished in the same rectifier. The new design combined two full-wave rectifiers in a single circuit. The evolution of this circuit is shown in Fig. 1-10. In Fig. 1-10A, two full-wave rectifiers, one with a positive output, the other with a negative output are displayed. Both rectifiers "float" free of ground. Since the object of the combination is to double the output voltage, the center taps of the transformers are connected (Fig. 1-10B). The voltage across each of the capacitors ($E/2$) are in series, and add to a combined value of E . When phased as shown, points a and b at any given instant are at the same potential, and may thus be connected without affecting operation of the circuit; the same is true for points c and d. The two windings are therefore in parallel and may be replaced with a single winding. This has been done in Fig. 1-10C, and the entire circuit has been redrawn for convenience of interpretation.

Operation of the bridge rectifier is rather straightforward. On the positive alternation of the secondary voltage, electrons flow through the load, diode D1, the secondary winding, and back to the load through D4. On the negative half-cycle, electrons again flow through the load, through diode D2 and the secondary winding, returning to the load through diode D3.

To find the peak inverse voltage on the nonconducting diodes, it is only necessary to consider the conducting diodes as short circuits. The total voltage across the nonconducting diodes is then the sum of the secondary and filter capacitor voltages, or $2E$. This voltage is divided across the two diodes so that each has an impressed PIV of only $1E$.

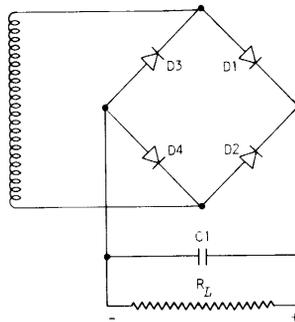


Fig. 1-11. Conventional bridge-rectifier schematic.

The bridge rectifier gets its name from the configuration in which it is usually presented in standard textbooks (Fig. 1-11). A quick examination should suffice to establish the equivalency of this circuit and that shown in Fig. 1-10C, the typical Tektronix presentation.

An evaluation of the bridge rectifier reveals a number of important features. The high-current output of the full-wave rectifier has been retained. The center tap has been eliminated from the secondary winding, and the peak inverse voltage impressed on the nonconducting diodes has been reduced from $2E$ to E volts. The AC-ripple frequency has been maintained at twice the line-voltage frequency. An output voltage equal to the secondary peak voltage, such as was attained with the half-wave rectifier, has also been achieved.

When constructed with semiconductor rectifiers, the bridge configuration has several obvious advantages over the full-wave and half-wave rectifying circuits. However, in vacuum-tube circuits, many of these advantages are lost or offset. Since the PIV rating of vacuum tubes is so high, the bridge rectifier offers no advantage, especially since no current gain is achieved. Also, a vacuum-tube bridge circuit requires at least one more heater supply and occupies twice as much space as the full-wave type. Finally, since the two conducting diodes of the bridge circuit are in series, the voltage drop across each of them subtracts from the total voltage applied to the load. The voltage drop across semiconductor diodes is so small it can usually be disregarded, but the drop

across vacuum-tube diodes is 20 to 30 volts and therefore presents a serious problem, especially in regulated power supplies. For these reasons, the bridge rectifier is seldom used in the vacuum-tube power supplies of electrical instruments.

center-tapped bridge rectifiers

Some full-wave rectifiers act as both bridge and center-tapped circuits (Fig. 1-12).

As the top of the secondary winding becomes positive, diodes D2 and D3 are forward-biased (A). Electrons flow through R_{L1} , D3, the secondary winding, and back through D2 to ground. Diodes D1 and D4 are reverse-biased at this time. One-half the secondary winding acts as a source for R_{L2} (B). On the same alternation as above, electrons flow through R_{L2} , through the secondary winding and back through D2 to ground.

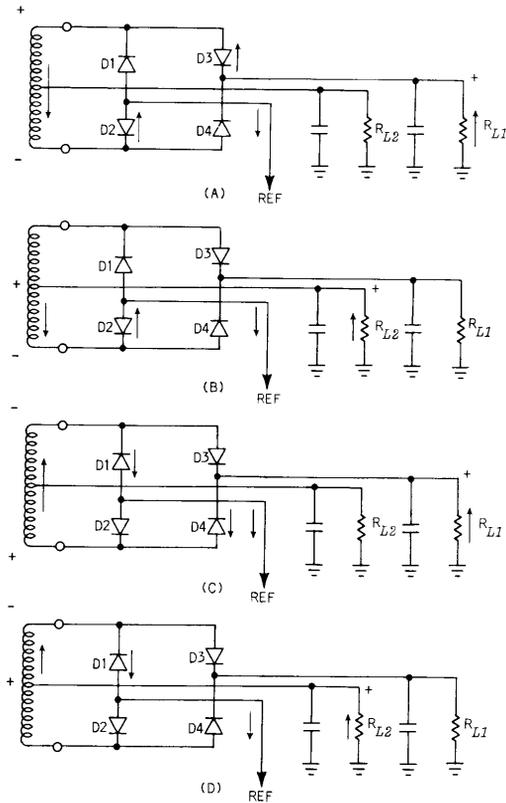


Fig. 1-12. Double-duty rectifier.

On the next alternation (C) diodes D1 and D4 are forward-biased. Electrons flow through the load, through D4 and the secondary winding and back through D1 to ground. D2 and D3 are reverse-biased on this alternation. Electrons also flow through R_{L2} , through the secondary winding, forward-biased diode D1, and back to ground (D). Thus both R_{L1} and R_{L2} receive DC; but the amplitude applied to R_{L1} will be twice that applied to R_{L2} since the entire transformer secondary voltage is applied to R_{L1} , while only half the secondary winding generates a voltage for R_{L2} .

2

VOLTAGE REGULATORS

Without further processing, the DC potentials developed in the power-supply rectifier and filter would lack the stability required by a precision test instrument. In the first place, fluctuations of AC line voltage are quite common; and though they have a negligible effect on the operation of ordinary household appliances, their effect on a rectifier output voltage is sufficient to cause serious degradation of an instrument's performance. Secondly, when an oscilloscope is shifted from one mode of operation to another, the load current may change abruptly. Without regulation, these current changes would result in unacceptable changes in the power supply's output voltage. Finally, the rectifier output, no matter how well filtered, contains an AC ripple component which would render many of the instrument's precision circuits completely useless. In the trigger circuits, for example, appreciable AC ripple would prevent stable triggering; in the sweep circuits it would introduce nonlinearities in the sweep ramp. (It should be noted at this point that although AC ripple is treated in most power-supply literature as a separate phenomenon, it is actually a form of input voltage fluctuation. Both forms are eliminated in the same manner.)

A voltage-regulated power supply must function so that its effective internal impedance over a specified range of load-current and input voltage fluctuations remains extremely small as compared to the load impedance.

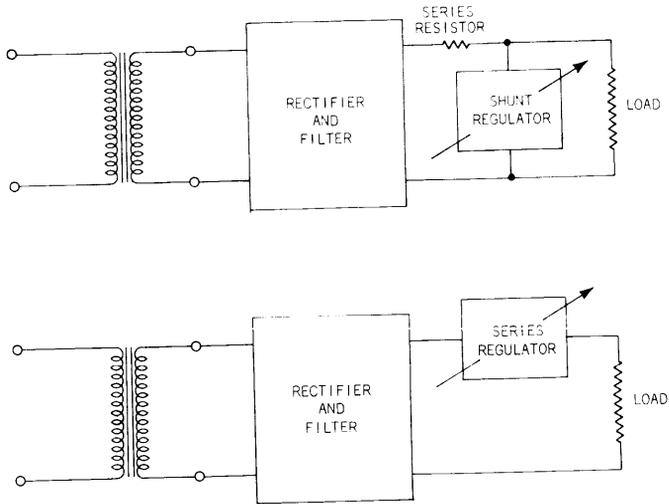


Fig. 2-1. Voltage regulators.

variable
impedances

Two classes of voltage regulators are found in conventional AC power supplies. These are the shunt regulator and the series regulator. Block diagrams of these types are shown in Fig. 2-1. The shunt regulator parallels the load, while the series regulator is in series with the load. Both regulators act as variable impedances, reacting automatically to changes in input voltage or load current demand in such a way that the voltage across the load remains essentially constant. The power source, in most cases is a rectifier similar to one of those discussed in the previous chapter.

shunt
regulator

The operation of the shunt regulator is relatively simple. Since the source voltage is dropped across a series resistor and the parallel combination of load and regulator, the current flowing at any time through the series resistor is the sum of the currents in the load and regulator. It is obvious therefore, that by making appropriate adjustments to the impedance of the regulator, the current through the series resistor can be controlled to maintain a constant voltage across the load impedance. This is true whether the original change in voltage across the load is caused by a change in source voltage or a change in load current demand (change in load resistance).

series
regulator

The operation of the series regulator is even more easily seen, since the source voltage is dropped across the series combination of regulator and load impedances. Therefore, any change in source voltage or load current can be counteracted by appropriate changes in the impedance of the regulating circuit.

monitoring
circuit

From the block diagram it is possible to draw certain conclusions as to the nature of the device which serves as the voltage regulator. Since the shunt regulator is in parallel with the load, any change in load voltage appears across the regulating component; therefore, the regulating component is able to sense or monitor the potential it must control. It is theoretically possible for this to be a single physical component, provided one with the required characteristics can be found. On the other hand, the regulating component of the series regulator cannot monitor the load voltage directly, so some type of monitoring circuit must be included to communicate error information to the regulating device.

Using only the block diagram it is also possible to compare the operating characteristics of the two regulators in abnormal situations. For instance, a short circuit across the load will not damage the regulating device in the shunt regulator, since the voltage across it drops to zero. Current through the series resistor increases, naturally, but not necessarily to the destruction level. An open-load circuit, however, forces the regulating device to carry excessive current, and may damage the device. These same conditions produce the opposite effect on a series regulator. When the load resistance is short-circuited, abnormal current through the regulating device may cause damage, but an open circuit can cause no ill effects.

VR tube

It has already been mentioned that a shunt regulator can theoretically be constructed with a single device possessing the proper characteristics. Two such devices, in fact, are available -- the first of these is the voltage-regulating (VR) tube. This tube, also known as a glow-discharge tube, is a cold-cathode diode filled with one of the inert gases such as neon, argon, etc. When a voltage is impressed on this tube, the gas begins to ionize. As the

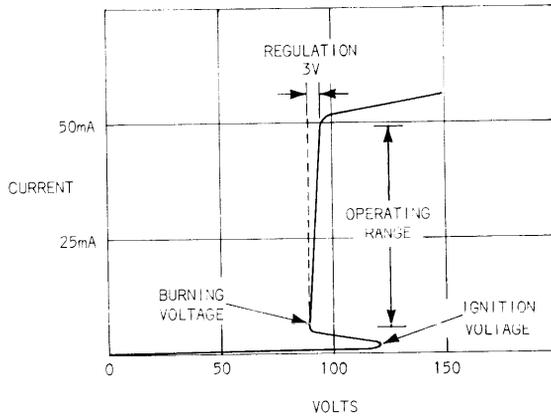


Fig. 2-2. Typical VR-tube characteristic.

ignition
voltage

burning
voltage

voltage rises, it reaches a critical value known as the *ignition voltage*, the point where sufficient ionization has taken place to allow current flow. Once ignition has taken place, the electron stream causes further rapid ionization of the gas, and the internal resistance of the tube drops to an equilibrium point. The voltage drop across the tube will simultaneously fall to a new level, provided of course that some resistance is in series with the tube. The new voltage level is called the *burning voltage*. Further attempts to increase voltage across the tube cause increased ionization of the gaseous conductor and lowered internal resistance of the tube, so that the voltage drop remains essentially constant over a wide range of current levels. Fig. 2-2 is a graph of the voltage-current characteristics of a typical VR tube. Note that the voltage remains essentially constant over the range between 5 and 40-50 mA. When placed in parallel with a load resistance, this device will maintain a stable voltage level by conducting more or less current in response to changes in load current or source voltage.

There are several limitations on the application of a VR tube as a single-device voltage regulator. In the first place, the circuit must be designed so that the voltage across the tube at the time of circuit turn-on equals or exceeds the tube's ignition voltage. Since, prior to ignition, the tube presents an almost infinite resistance, the series resistance

(including the internal resistance in the voltage source) must be limited to a value which will result in a voltage drop across the parallel leg of the circuit (tube and load) equal to or exceeding the tube's ignition voltage.

Another limitation on the VR tube is its heat-dissipation capability. This depends on a number of factors, including the size of the tube, the size of its electrodes, etc. In a typical miniature VR tube the upper current limit is in the 40-50 mA region. Current in excess of tube's maximum-current rating overheats the tube and results in distortion of the E-I curve. Since any changes in load current must be absorbed by corresponding changes in the VR tube current, the range of regulation which can be achieved by such a tube is limited to the difference between the tube's maximum-current and turn-on current characteristics and typically is considerably less.

The negative resistance portion of the curve between ignition voltage and burning voltage (where current increases as voltage drops), also creates difficulties. Gas-filled diodes generate an appreciable amount of *noise* (random current variations); therefore a capacitor is usually placed in parallel with the VR tubes to filter out the high-frequency noise (Fig. 2-3). Under certain conditions, the capacitor will cause the circuit to function as a *relaxation oscillator*. This is apparent when the circuit conditions at turn-on are considered. Initially, the VR tube acts as an open circuit, and the capacitor will charge toward the source voltage. When the charge on the capacitor reaches the ignition voltage of the tube, the tube will start conducting and the capacitor will discharge to the burning voltage. If for any

oscillation

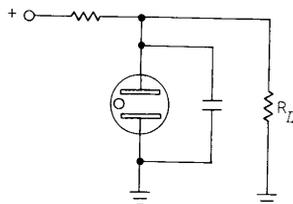


Fig. 2-3. VR-tube noise elimination.

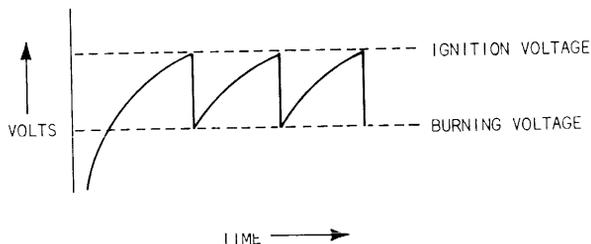


Fig. 2-4. Relaxation oscillation waveform.

reason the load current gets too high, and the series resistance is too high to allow the normal burning voltage to exist across the VR tube, the tube will extinguish itself. This reduces the current through the series resistor, increasing the voltage across the parallel circuit. The capacitor will again charge to the tube's ignition voltage. As this cycle repeats itself, a sawtooth voltage is imposed on the load as shown in Fig. 2-4. These oscillations in the load voltage are, of course, totally unacceptable. Although relaxation oscillation can only occur under abnormal conditions of load current, it does present problems to the designer and limits the regulating range of the VR tube.

Still another problem presented by VR tubes is that the tube's burning voltage, a function of electrode configuration and gas pressure, cannot be established at any desired level in manufacture. Typically, gas-filled diodes are only designable with certain discrete burning voltages, usually about 75 V, 100 V, and 150 V for VR tubes. Burning voltages less than 60 V are not obtainable, and the only way to achieve regulated voltages above 150 V is to connect two or more VR tubes in series. This can be done, providing certain precautions are observed. Fig. 2-5 illustrates the problem and its solution.

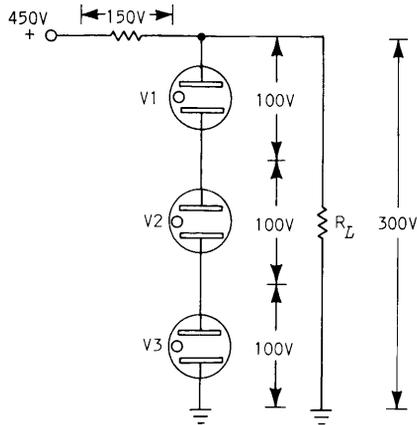


Fig. 2.5. VR tubes in series.

series
VR's

When two or more VR tubes are connected in series, the total voltage required to cause them to ignite (assuming that they have identical characteristics and zero leakage current) is the sum of their ignition voltages. If, for instance, V1, V2, and V3 are 100-V VR tubes with 150-V ignition voltages, the power-supply output voltage must be at least 450 V. Once the tubes have ignited, their burning voltages will add to about 300 volts, so about 150 volts must be dropped across the series resistor. This means that at least one-third (in practice, nearer one-half) the source power is dissipated in the series resistor -- not a very useful arrangement.

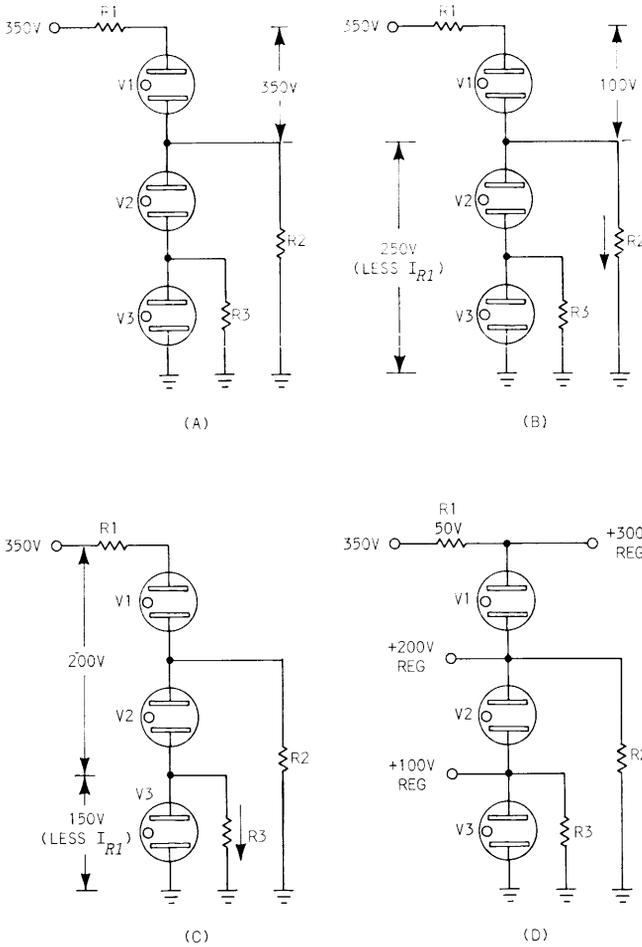


Fig. 2-6. Operation of practical series VR-tube regulator.

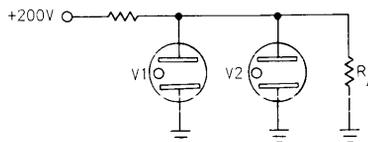


Fig. 2-7. Parallel VR tubes.

A much lower supply voltage can be used, however, if the lower tube in the case of a series pair, or the lower tubes in the case of three or more, are shunted by individual high-resistance bypass resistors (Fig. 2-6). At the instant of turn-on, (A) the entire supply voltage, (in this case 350 V), exists across the unbypassed VR tube, causing it to ignite immediately. The current through the ignited tube now must pass through the next lower resistor, R2, since the other VR tubes are still effectively open circuits (B). At this time the voltage across the second VR tube equals the power supply voltage minus the burning voltage of V1 and a small drop across series resistor R1.

Once V2 has ignited, all the current passing through V1 and V2 must also pass through R3 (C). The voltage across V3 is therefore sufficient to cause ignition. Thus only 50 volts or one-seventh of the power-supply output power is dissipated in the series resistor, and reliable ignition of the VR tubes is assured (D). (A power-supply voltage of about 375 volts would be more typical, to provide a wider operating margin.)

From a series string of VR tubes such as the one just examined, it is possible to take a number of regulated intermediate voltages. That is, in addition to the regulated 300 V drawn from the top of the string, it would be quite valid to use the 200 V and 100 V potentials as regulated supply voltages. In this case it is necessary to evaluate the individual load circuits and determine that under worst-case conditions, no individual VR tube would be "starved" or overloaded. In actual practice it usually proves that only a small amount of current may be drawn from the intermediate supplies without adverse effects.

It is at least logical to inquire why, since the current range of typical VR tubes is so limited, it would not be possible to increase the regulator's effective range by using VR tubes in parallel. The answer is apparent when the circuit in Fig. 2-7 is closely examined. At turn-on, the entire power-supply voltage of 200 volts is impressed across the parallel circuit V1 and V2, a pair of 150-V VR tubes. However, VR tubes with absolutely identical ignition voltages are almost impossible to obtain. Therefore one tube would ignite before the other. At this instant, the voltage drop across the parallel circuit would drop to 100 volts, making it impossible for the other tube to ignite.

parallel
VR's

If carefully selected resistors are placed in series with the individual VR tubes, the circuit can be made to work. However, the added resistance degrades the regulating capability of the circuit. The use of parallel VR tubes is thus considered poor engineering procedure. Clearly, there are rather severe limits to the usefulness of VR tubes as single-device shunt regulators, particularly in semiconductor circuits.

stable
reference
voltage

There is, however, another important application for which the VR tube is particularly well suited. In the series-regulated power supply, which will be discussed shortly, the VR tube is employed to supply a stable reference voltage for the error-sensing circuits. Highly stable, specially designed VR tubes have been developed for this purpose.

For reasons beyond the scope of this discussion, the burning voltage of these tubes falls in the region between 83 and 87 volts with a burning current of 2-6 mA. When provided with the burning current specified by the manufacturer, such tubes exhibit extreme stability (characteristically $\pm 0.01\%$), over as much as ten-thousand hours. Since, in a series regulator, it is the long-term stability, rather than the actual level of the reference voltage that is of major importance, it is obvious that these tubes are almost ideal sources of such reference voltages.

The burning current, however, is rather critical. If, after operating at the proper burning current for a thousand hours or so, the burning current shifted two or three mA for some reason, the tube would behave abnormally. With a series resistance of a few hundred ohms, a shift of two or three mA in burning current would change the reference voltage as much as a volt or so. More important, however, is the effect such a shift would have on the tube's stability. Instead of settling down at the new reference-voltage level, the tube would be relatively unstable, and the reference voltage would drift rather rapidly with respect to time. Under these conditions its stability might be more like $\pm 0.1\%$ over 500 hours, and a relatively long period would elapse before it regained its former stability. Under normal conditions, this situation rarely occurs and is rather easily prevented by careful circuit design. However, it sometimes happens that accidents upset the normal functioning of the circuit, and some kinds of abuse, such as a

reversal of current flow through the tube or current overloads, can render the tube useless as a voltage regulator.

zener diode Another device, the zener diode, may also be employed directly as a shunt regulator and voltage reference. This device resembles the common semiconductor diode, but has been specially doped and otherwise processed so that its reverse-breakdown characteristics are highly predictable. Fig. 2-8 is a voltage-current curve of a typical zener diode, showing the sharply defined reverse-breakdown characteristic. Note that once breakdown has occurred, the reverse voltage remains essentially unchanged over a wide current range. By careful control of the manufacturing process, breakdown voltages ranging from about 2 to 200 volts are obtained.

reverse
breakdown

Two distinct mechanisms are responsible for the characteristics of the zener diode. The first, "zener" breakdown, occurs between about 2 and 7 volts, and resembles the phenomenon known as "quantum tunneling" exhibited by the tunnel diode. The other mechanism, which occurs between about 5 and 200 volts, resembles the phenomenon known as "avalanche" breakdown, which occurs in avalanche transistors. Diodes with reverse breakdown characteristics in this region might more accurately be called avalanche diodes, but for convenience are included in the zener-diode category.

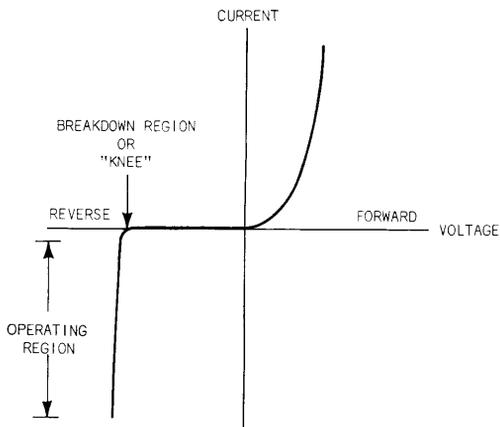


Fig. 2-8. Typical zener-diode characteristic.

temperature
coefficient

The distinction is made here to explain an important characteristic of zener diodes in the 5 to 7-volt range. *True* zeners exhibit a *negative* temperature coefficient of voltage while the avalanche type has a *positive* temperature coefficient. Thus, diodes operated in the overlapping 5 V to 7 V region have a temperature coefficient approaching zero. This means, of course, that these diodes are stable over a wide range of temperatures.

temperature
compensation

Above this narrow range, zener diodes require some form of temperature compensation in order to attain maximum stability. A common technique is shown in Fig. 2-9. An ordinary junction diode and a zener diode are connected in series and encapsulated so that they operate at identical temperatures. The junction diode is carefully chosen so that its temperature coefficient is equal to that of the zener. Since the junction diode has a negative temperature coefficient, a given change in temperature causes an increase in the voltage drop across one of the diodes, but an equal reduction in the voltage drop across the other diode. The voltage drop across the series pair thus remains essentially constant with variations in temperature.

Zener diodes can handle large currents, for practical purposes limited only by their ability to dissipate power in the form of heat. Heat sinks, which conduct much of the heat to the chassis where it is radiated to the surrounding air, increase the power-handling capacity of zener diodes to more than 100 watts. Thus the diode can be used as a shunt regulator for high load currents, especially at lower voltages. Load currents of several amperes are not unusual.

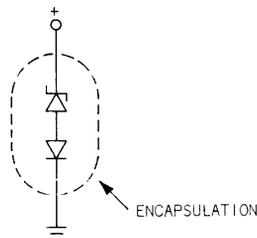


Fig. 2-9. Zener-diode temperature compensation.

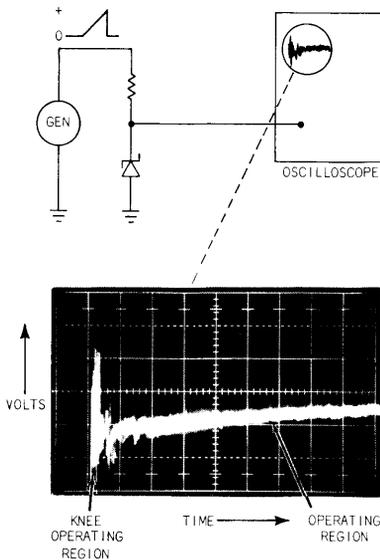


Fig. 2-10. Zener-diode noise.

zener
noise

Like the VR tube, the zener diode generates a certain amount of noise in operation. Most of this noise, however, occurs close to the "knee" of the reverse-transfer curve, Fig. 2-10. Noisy operation is usually avoided by operating the diode at a reverse voltage well in excess of the knee voltage; however, the power dissipated by the diode in these circumstances may be unacceptable for a particular application. In this case the zener may be operated at a reverse voltage close to the knee, and require a filter bypass capacitor as in the case of the VR tube. Since the zener diode curve has no negative-resistance characteristic, no relaxation oscillation problem is encountered. For the same reason, zener diodes may be stacked in series; and, by taking advantage of the wide range of available zener voltage ratings, a combination may be found to meet any practical power-supply voltage.

Zener diodes are especially useful as shunt regulators for high-current, low-voltage power supplies where VR tubes cannot be used. At voltages in the 100 to 200-V range, however, zener diodes become moderately expensive, since their power rating must be increased by such a large factor. In this area the VR tube

may prove more economical. Zener-diode shunt regulators are typically found in power supplies with output voltages of 50 volts or less where VR tubes cannot be used at all.

reference
voltage
source

For the same reasons that applied to the VR tube, the zener diode finds its most important application as a reference-voltage source for series-regulated power supplies, especially low-voltage, transistorized power supplies. Here the relatively high ignition voltage of the VR tube rules it out as a reference voltage source. By choosing a zener diode with an inherently low temperature coefficient (near zero), specifically designed for this type of operation, a reference voltage can be obtained which will remain stable over many thousands of hours of operation. As was true of the VR tube, the zener diode's actual breakdown voltage is relatively unimportant; stability of operation is the important consideration. To attain maximum stability, the zener must be guaranteed a constant current. Once again, the circuit must be designed to "pamper" the diode so that its exact specified operating conditions are met. The extreme stability which can be obtained with zener diodes is illustrated by the fact that some models of a certain high-quality differential voltmeter use a zener diode as a reference voltage source, while other models of the same instrument use a Weston Standard Cell, indicating that the long-term stability and accuracy of the two sources are comparable.

Like the VR tube, the zener diode is seldom found in parallel configurations. Although the zener does not have the negative-resistance characteristic which makes parallel combinations of VR tubes impractical, there is a problem in obtaining exactly matched diode breakdown-voltage and current characteristics. Unless the breakdown voltages are perfectly matched, of course, one diode will do all the work, and the stability of the parallel combination cannot exceed that of the most unstable diode.

Most of the shunt regulators employed in Tektronix instruments are single device types similar to those discussed above. Their application is limited to those infrequent situations where economy or extreme accuracy are the ruling factors.

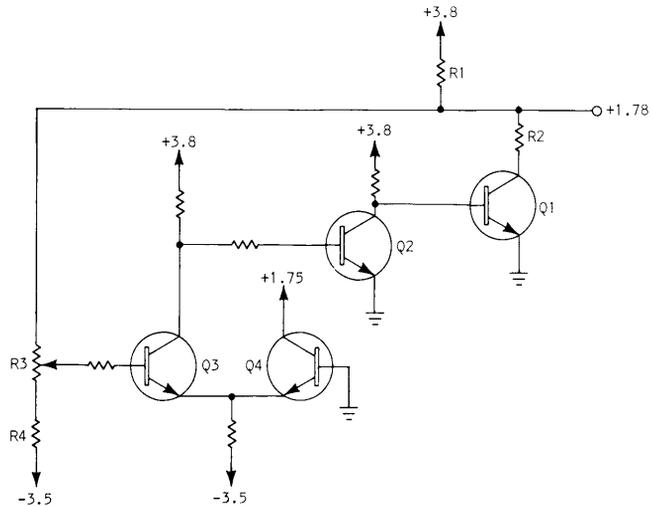


Fig. 2-11. Shunt voltage regulator.

There are a few situations which call for a shunt regulator of more complex design. As will be shown later, a series regulator will not function if load current drops below a certain minimum value. The output voltage in this case tends to wander up toward the value of the unregulated source voltage. A series regulator, therefore, cannot be used in a power supply which provides the operating potential for a circuit which alternates between an "off" and a "full on" condition. The shunt regulator shown in Fig. 2-11 however, operates quite satisfactorily under these conditions.

The source voltage, 3.8 V, is dropped across R1 and the regulator. The output voltage, 1.78 V, is taken from the junction between R1 and R2.

A voltage divider consisting of R3 and potentiometer R4, is connected between the output voltage and a regulated -3.5 volts. The potentiometer is adjusted so that the left-hand base of voltage comparator Q3-Q4 is at zero volts. The right-hand base is connected to ground. If the output voltage varies, an error signal is generated in the comparator amplified in Q2, and applied to the base of Q1 in the proper phase to offset the original change in

output voltage. For instance, if the output voltage tended to move in the negative direction, a small negative signal would be applied to the base of Q4. A positive error signal would appear on the collector of Q3, be amplified and inverted in transistor Q2, and appear as a negative signal at the base of Q1. Current through Q1 would therefore decrease, and the voltage at the junction of R1 and R2 would rise to offset the original change. Should load current go to zero, the regulator would act in the opposite direction. Transistor Q1 would in this event receive a large forward bias signal, decreasing its resistance sufficiently to maintain a constant current through R1 and thus a constant output voltage at the junction of R2 and R3.

In general, shunt regulators have only a limited application in Tektronix instruments. The demands placed upon most power supplies can best be met by employing some type of series regulator.

series-
regulation

As its name implies, the series regulator is simply a regulating device connected in series, between the voltage source and the load, Fig. 2-1. As the load current or source voltage fluctuate, the regulating device must alter its impedance so as to maintain a constant voltage drop across the load.

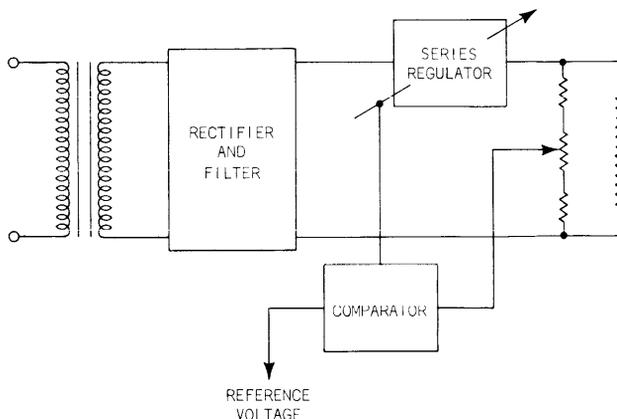


Fig. 2-12. Block diagram of series voltage regulator.

monitor Alone, the regulating device cannot sense changes in load voltage. Therefore, a monitoring circuit, usually some type of voltage divider, is connected in parallel with the load, as in Fig. 2-12. As this arrangement does not permit quantitative measurement of load voltage changes, a stable reference voltage is required, with which a portion of the output voltage can be compared. The reference voltage is usually supplied by a zener diode or VR tube, usually rated at about 85 V for VR tubes and about 5 to 7 V for zener diodes. It is, therefore, usually impractical to compare the output and reference voltages directly. However, when the regulated voltage is at the desired level, a point will exist on the divider, which matches the reference voltage.

comparator Now all that is required to complete the regulator is a circuit or device which will compare the monitor voltage with the reference voltage and generate a corrective signal when a difference exists between them. The corrective signal is applied to the regulating device in such a way as to bring the output voltage back to its former value.

It may be helpful at this point to correct a popular misconception fostered by most textbooks on power supplies. In their descriptions of the series regulator, they state that the power-regulating device acts to maintain a constant output voltage. It actually does nothing of the kind. The regulating device has no way of determining what the output voltage is; it cannot, therefore, act to maintain a constant output voltage, or any particular value of output voltage. What the regulator does "know" about the output voltage is revealed only by the sample taken from the monitoring circuit. The action of the regulator, then, is to maintain equality between this voltage and the reference voltage supplied to the comparator. The regulator *indirectly* controls the output voltage because the sample taken from the monitor is proportional to the output voltage. Once this distinction is firmly established, it is obvious that the output voltage will only be constant if the resistance in the voltage-divider components remains at a constant ratio. A potentiometer in the voltage divider thus permits adjustment of the output voltage to the desired level.

Another misconception encouraged by classical treatments of power-supply operation is that fluctuations in the supply voltage and AC ripple are separate phenomena; the fact is, of course, that they are simply long-and-short-term varieties of the same problem. The same circuit mechanism is called upon to handle both aberrations, and AC ripple is simply a special case of supply-voltage fluctuation.

To simplify the discussion of practical series-regulator circuits, a vacuum-tube regulator will be used as an illustration, since this type is relatively simple as compared to the transistorized variety. (The greater complexity of transistorized regulators arises mainly from the number of protective devices which must be included in the circuit to prevent damage to the semiconductors.) Also for reasons of simplicity, a positive voltage will be assumed.

A schematic diagram of a simple series regulator is shown in Fig. 2-13. Triode V1 serves as the regulating device. Resistors R1 and R3, together with potentiometer R2 form the output-voltage monitoring system. Control and comparison functions are combined in a single tube, V2, a conventional long-tailed differential amplifier. The input to one grid of the comparator is taken from R2. The ratio of the divider is chosen so that under nominal operating conditions, the voltage at the center of the potentiometer is equal to the chosen reference voltage. The other grid receives the reference voltage, in this case a nominal 85 volts, supplied by VR-tube V3. As previously explained, V3 must be supplied with a constant specific current to assure maximum stability; therefore, its current source is the regulated output of the power supply itself. Since the output voltage is held constant, the current to V3, fed from series-resistor R4, is constant.

The grid of the regulator tube, V1, must now be controlled by the output of the comparator. To determine the required polarity of the corrective voltage it is only necessary to assume some change in the load current. For instance, if load current suddenly increases, the output voltage would tend to drop, due to the increased voltage drop across V1. To offset this tendency, V1 must allow more current to flow, which in turn requires a positive signal on its grid.

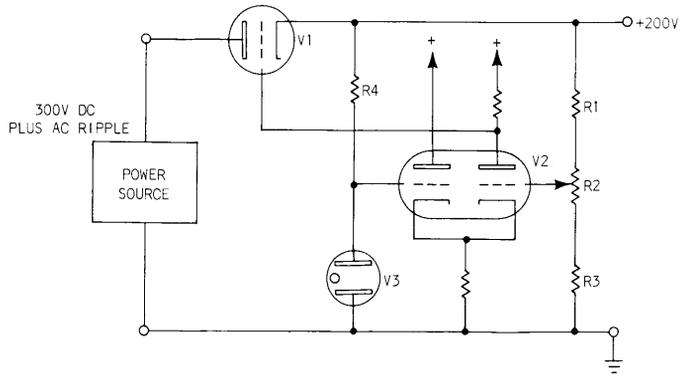


Fig. 2-13. Basic circuit of series regulator.

A drop in output voltage (since this is a positive supply) would be sensed as a negative signal with respect to ground at the right-hand grid of V2. Since the other grid is held fixed by the reference voltage, current would increase in the left side of the comparator and decrease in the right side. The required positive corrective signal must therefore be taken from the right side. A plate-load resistor should thus be placed in the right-hand plate circuit, while the left-hand plate can be led directly to the positive supply. (The regulated voltage of relatively high output supplies can be used as a plate voltage source. In other cases, plate voltages must be furnished by another power supply with a higher output voltage.) The regulator is now complete.

At this point it is advisable to consider how various components affect the performance of the circuit, and some of the techniques available to minimize the circuit's limitations. First, it is necessary to assign some values to the characteristics of the series-regulator tube. This will usually be a triode and, for reasons which will be discussed later, should have a very low amplification factor, or μ . A typical triode for this application might have, for instance, a μ of 2 and a g_m (transconductance) of 5 milliamperes/volt.

It is now possible to consider the performance of this circuit in terms of both load-current changes and supply-voltage changes. It should be stated that the analysis which follows is not intended to

be a precise design study, and that a number of sweeping assumptions will be made whose validity might well be questioned by a designer of electronic circuits. However, this cursory study should emphasize the significance of the various parameters involved, even though from a quantitative point of view, 100% accuracy is not achieved. This approach is justified by the fact that in order to develop completely defensible formulae for the circuit's performance, the discussion would soon become bogged down in a number of side effects and the central point could easily become obscured.

As a point of entry to the discussion, assume that for one reason or another, an increase in load current of 20 mA occurs. It is not necessary to know its former value, but before its effect can be explored it will be necessary to assign some arbitrary values to certain voltage parameters and components of the circuit. Assume for instance, a regulated output voltage of 200 volts and a rectifier output of 300 volts. To simplify the discussion, a reference voltage of 50 volts will be assumed, although, as pointed out in the discussion of the VR tube, it would more typically be around 85 volts.

Other values can now be assigned on the basis of the preceding assumption. The voltage at the pickoff of the potentiometer must be 50 volts, since it must match the reference voltage. The resistance ratio of the monitoring circuit must therefore be four-to-one ($\frac{50 \text{ V}}{200 \text{ V}} = \frac{1}{4}$). It follows therefore that there is a 150-volt drop across the upper portion of the divider and a 50-volt drop across the lower portion. With an output voltage of +200 volts, under reasonable load conditions, V1 will require about -20-volts bias, establishing the grid potential at approximately +180 volts with respect to ground. This indicates about a 20-volt drop across the plate load resistor of the comparator, and about 130 volts on its cathodes.

Returning to the assumed 20-mA change in the load current, if the voltage source is assumed to have zero impedance, the plate of V1 will remain at 300 volts. With the further assumption that no effective change occurs in the regulator's output voltage, it is possible to calculate the voltage change the control circuit must produce to establish equilibrium

in the increased-current condition. Since V1 has a g_m of 5 mA/V, and since the plate-to-cathode voltage has remained constant, the grid voltage must change in the positive direction by 4 volts to produce the required 20-mA change in cathode current. This requires a 4-volt change in the voltage drop across the plate load resistor of the comparator tube.

It is necessary now to make some additional assumptions about the comparator circuit. Under the conditions established by other assumptions, the gain of the comparator between right-hand plate and grid will be about forty. A change of 4 volts on this plate then would require a 100-mV change on the corresponding grid. Since the grid voltage is taken from the 4-to-1 potential divider, a 400-mV change must occur in the regulator's output voltage.

power-
supply
impedance

With this information it is possible to calculate the impedance of the power supply. Since a 20-mA change in current resulted in a 400-mV change in output voltage, the output impedance of the supply is 20 Ω .

line-
voltage
variations

Obviously, considerable improvement is necessary before this regulator can be considered a practical circuit. Nevertheless, it serves to illustrate the basic mechanisms at work in all series regulators. Before going on to the task of improvement, however, it will be advantageous to examine the behavior of this relatively crude regulator as it responds to changes in source voltage.

Assume for the purposes of this examination that the supply voltage rises 20 volts. Since no change in load current demand has taken place, load current itself must remain constant, as must the output voltage. Triode V1, then, has 320 volts on its plate and 200 volts on its cathode. Thus the grid voltage must change by an amount which will maintain the former current under the changed conditions.

With the assumed μ of 2 and a 20-volt increase in plate-to-cathode voltage, the grid-to-cathode voltage must increase by 10 volts; that is, the grid must go 10 volts more negative with respect to the cathode. Following the same procedures as were employed in the previous example, it is apparent that the comparator's plate voltage must also go negative by 10 volts. Assuming again a gain of 40 for the

comparator, an increase of 250 mV must take place on the right-hand grid. Due to the resistance ratio of the monitoring circuit, a 250-mV change at the pickoff point requires a 1-V change in the output voltage. Thus, a 20-volt change in input voltage causes a 1-volt change in output voltage, or a change of 6.6% ($\frac{20 \text{ V}}{300 \text{ V}}$) in the rectifier voltage causes a 0.5% change in output voltage, a ratio of about 13:1.

This ratio introduces a rather important concept concerning the performance of a power supply. Unfortunately, no standard terminology has emerged in the industry to describe it accurately. For the purposes of this discussion, the ratio of *percentage change* in input voltage to the *percentage change* in the output voltage of the regulator will be referred to as the *line-voltage variation rejection ratio*. This awkward term may be improved in the future, but for the present it will serve to describe one of the important capabilities of an instrument's regulated power supply. At first glance the choice of the term "line-voltage variation" may not appear to be justified, since the foregoing analysis compared only the rectifier output to the regulated output. However, since the rectifier output is directly proportional to line voltage, and since changes in line voltage are actually responsible for most significant changes in the rectifier's output, the reference to line-voltage variation is more meaningful.

In the foregoing example, a line-voltage variation rejection ratio of 13:1 is obviously very poor, so the next logical step is to explore the means of improving the regulator's performance in this and other respects.

It can be seen immediately that line-voltage regulation can be improved by increasing the gain of the series regulator, while an increased g_m will improve load-current regulation. Both μ and g_m are related to the plate impedance of the tube; thus, if a high μ and high g_m are desired, the tube must inevitably have a high plate impedance. This means, in simple terms, that in order to extract a certain cathode current from the tube, its plate-to-cathode

voltage must be high. On the other hand, a high plate voltage results in considerable power loss due to dissipation by the tube, and also dictates a higher power-supply voltage. But, excessive power dissipation by the series-regulator tube creates serious problems in eliminating the heat from the instrument and subtracts from the power delivered to the load. Also, the voltage supplied by the rectifier should be kept as low as possible, so as to remain within the voltage limitations of silicon diodes and to avoid the complications introduced by high-voltage transformers. Thus, the requirements of increased plate impedance for better regulation and those of economy and efficiency seem to be incompatible.

It so happens, however, that while load current regulation is irrevocably tied to the g_m of the series-regulator tube, there are ways of improving line-voltage regulation without increasing μ . This subject will be examined shortly. It suffices to say at this point, that because these other methods are available, it is customary to strive for the highest possible g_m and lowest possible plate resistance, accepting whatever value of μ happens to emerge from the eventual configuration. A typical series-regulator triode such as the 6080, for instance, often has a μ as low as 2.

error
amplifier

With so little gain available in the series-regulator tube it is obvious that relatively large changes in grid voltage are required to effectively regulate changes in line voltage. Yet, if regulation is effective, the change in output voltage sensed by the monitoring circuit will necessarily be very small. This leads to the inevitable conclusion that the control circuit must be provided with considerably greater gain than was true in the basic regulator. Because there are limits to the amount of gain available from the comparator configuration, the next obvious step is to add another stage of amplification to the control circuit. This stage, like most amplifiers, will invert the correction signal, so the error signal must be taken from the opposite plate of the comparator. Fig. 2-14 is a schematic diagram of the improved regulator.

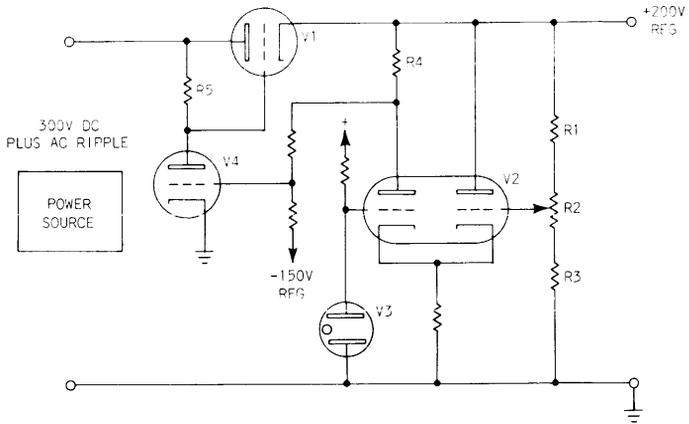


Fig. 2-14. Series regulator with triode error amplifier.

Before proceeding further, various circuit load conditions must be considered. Under the same conditions as those assumed for the basic circuit, the grid of V1 should again be about 180 volts positive, which means that the plate voltage of triode V4 must also be established at 180 volts. In order to get maximum gain from V4, however, its plate resistance should be very high. With a 300-V source supplying the plate, the voltage drop across plate-load resistor R5 will be on the order of 120 volts, a plate-load resistance of one or two megohms is thus established. It is often necessary to use an *external* plate-supply voltage in lower-voltage supplies, in order to assure the desired amount of plate resistance.

The grid voltage of V4 must be of the same order as the cathode voltage, or approximately zero volts. However, it has already been established that the plate of V2 from which the error signal is taken operates at approximately 120 volts. Another potential divider is therefore required between this point and a regulated negative-voltage source to bring the error signal to the required DC level.

Assuming a typical gain of 100 for V4, and thus an additional gain of 100 for the entire control circuit, new figures can be established for the overall behavior of the improved regulator.

Assuming the same 20-mA change in load current as was proposed for the basic regulator, the output voltage now changes only 4 millivolts, and thus exhibits an output impedance of only 0.2 ohms. Similarly, the 20-volt change in supply voltage now causes only a 10-mV change in regulated output voltage. The improved regulator thus exhibits characteristics that approach practical performance levels. Further refinements inevitably lead to additional problems, but the majority of these are simply variations on those discussed in the foregoing presentation.

The first refinement to be considered concerns the error amplifier, V4. Although in the foregoing example a triode was used as the error amplifier, a pentode with a gain in the order of 1000 is almost invariably used in practical circuits (Fig. 2-15). In fact, it is the pentode which actually makes it possible to use a triode as a series regulator.

In order that V4 act as a pentode, its screen grid must be connected to a constant voltage source; that is, the screen grid-to-cathode potential must not be allowed to fluctuate in sympathy with the error signal. The regulated output voltage provides a convenient source, and is brought to the desired level by a voltage divider (R7-R8) connected to signal ground.

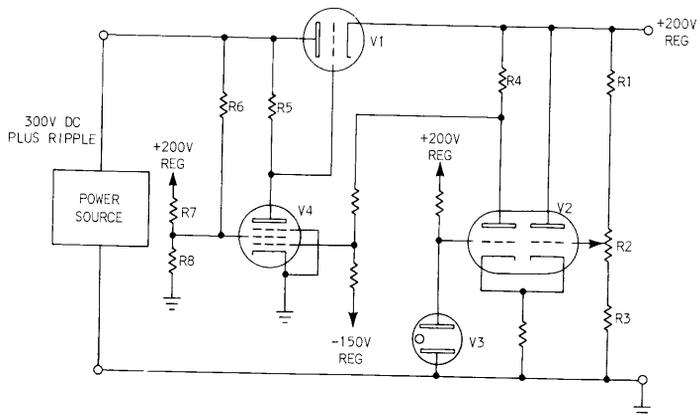


Fig. 2-15. Series regulator with pentode error amplifier.

An important advantage can be gained by connecting the screen grid through another resistor to the unregulated input voltage. In any pentode, the screen grid acts in conjunction with the plate and cathode in much the same way as the control grid of a triode. The pentode therefore has a certain value of μ and ξ_m , as applied to its screen grid. Signals applied to the screen grid are therefore amplified and phase-inverted as they appear on the plate. The parameter of interest here is the μ of the screen grid, sometimes referred to as the "inner μ " of the pentode, which in small pentodes is on the order of ten or twenty. Assuming a relatively high plate resistance and a μ of twenty, the screen grid-to-plate gain may be assumed for the sake of discussion to be the same.

With the parameters of the pentode established, consider the signal applied to the screen grid. In this case it is the unregulated supply voltage, or more accurately, *changes*, in the supply. The screen grid is usually supplied from a three-resistor network. Two of the resistors, R7 and R8 constitute the voltage divider which establishes the DC level of the screen-grid bias voltage. The other resistor, R6, usually has a rather high value and is connected to the unregulated supply, feeding a sample of any change in supply voltage to the screen grid. (Because both R7 and R8, are, in effect, connected to signal ground, only R6 communicates changes in the supply voltage to the screen grid.) The sample, of course, is only a small fraction of the actual supply voltage change, due to the attenuation

$$\frac{R7 \cdot R8}{R7 + R8} \cdot \frac{1}{R6}$$

the amplified signal at the plate of V4 will be $E_{reg}KA$, where A represents the amplification factor.

Now if the supply voltage increases by 20 volts, and the μ of V1 is assumed to be 2, as was assumed in previous examples, the grid of V1 must go negative by 10 volts to effect a full correction. This will occur if the plate of V4 goes negative by the same amount. Thus, if $20 \times K = 10$, the series-regulator tube will absorb 100% of the change in supply voltage and the regulated output voltage will remain unaffected.

In actual practice the screen-grid feedback loop does not function quite this efficiently, due to the tolerance allowed in the voltage-divider resistors, variations in pentode performance, etc. Nevertheless, the small remaining voltage change is easily handled by the control-circuit feedback loop in the same manner described in previous examples. The most important point to remember in this respect is that different mechanisms are involved in the two circuits. The main feedback system of the control circuit is actually an error-nulling system, requiring a certain amount of residual error for its operation. The screen-grid loop, on the other hand, is a straight translation system; and it is at least theoretically possible to obtain from this translation perfect regulation of the supply voltage, including total elimination of AC ripple. The small percentage of the AC ripple found in power supplies whose regulators take this last configuration is usually due to small phase shifts in the feedback loop or ripple in the supply voltage from other power supplies.

It should now be quite clear how the conflicting demands on the series-regulator tube are satisfied. Since high μ , high g_m and low plate impedance cannot be realized simultaneously, a low- μ tube with a high g_m is used, and the necessary error amplification is provided by an additional stage of amplification in the control circuit. An additional benefit is gained when a pentode is employed as the error amplifier, in that its screen grid is utilized to control variations in the supply voltage.

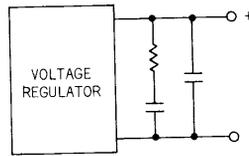
In a later chapter the voltage regulator is shown to be actually a low-gain (less than one) feedback amplifier. It is not surprising, therefore, that among the parameters of interest in power-supply performance is the frequency response, or bandwidth of the regulating circuits.

In the foregoing discussions, the effects of a changed supply voltage or load current have been explored, but no attention has been given to the *speed* with which the regulator adjusts to these changes. There are a number of circuits in an oscilloscope whose normal function causes very abrupt changes in load current. In the sweep generator for instance, there

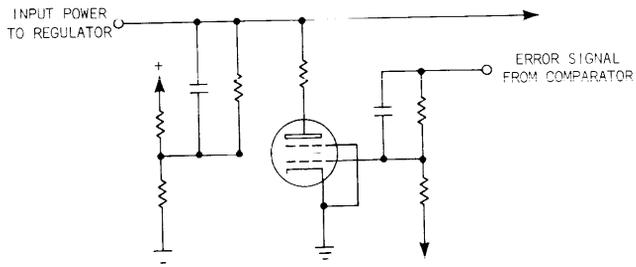
is considerable difference between the load current drawn during the sweep and that drawn during the hold-off period, and the transition from one state to the other is extremely rapid. Unless the power supply can handle these rapid current changes, there will be voltage changes in the output, which will then be coupled to other circuits of the instrument. Therefore, the power supply must be further modified to follow high-frequency changes as well as longer term changes (such as those caused by shifting plug-in units in an instrument). In other words, the regulator must exhibit a wide frequency response.

Very high-frequency changes cannot be handled by any action of the regulator. It has already been shown that the feedback loop of the control circuit must exhibit the maximum possible gain for best regulating action. This requires high plate impedances in the comparator and error amplifier, limiting the frequency response of the circuit to a few tens of kilohertz at the maximum. Frequency response of this order is quite insufficient for the regulating of the fast current changes under consideration. The solution to the problem is to install a large "buffer" capacitor across the output of the regulator (Fig. 2-16). Electrolytic capacitors with values

buffer capacitor



BUFFER CAPACITORS



SPEEDUP CAPACITORS

Fig. 2-16. Improving regulator frequency response.

between 10 and 50 μF are typically used in this application. The value is so chosen that at frequencies where the regulating circuit is unable to respond fast enough, the capacitor acts as a very low impedance; at frequencies where the impedance of the capacitor begins to increase, the regulator takes over. The net result is that the output impedance of the regulator remains quite low over a fairly wide frequency range.

Because electrolytic capacitors exhibit losses at very high frequencies which make them appear inductive, a second paper or ceramic disc capacitor of smaller size is sometimes added to the circuit, in parallel with the electrolytic capacitor. To avoid oscillations due to unavoidable phase shifts introduced by the additional capacitance, a small resistor is sometimes connected in series with the smaller capacitor.

At midrange frequencies (those in the neighborhood of line-frequency ripple) another technique is employed to improve the regulator's performance. The gain of the regulator at these frequencies is approximately equal to the product of the comparator and error-amplifier gains, divided by the attenuation of their associated voltage dividers. The voltage dividers cannot be eliminated, since they are required to establish the DC operating conditions of the circuit, but their attenuating effect on midfrequency signals can be virtually eliminated by connecting capacitors of the proper value in parallel with the upper half of the voltage dividers. These "speed-up" capacitors increase the gain around the circuit by a factor of 3 or 4 at the frequencies under consideration.

speed-up
capacitors

It should be noted that the buffer capacitor is usually located not on the power-supply chassis, but adjacent to that part of its load circuit which is mainly responsible for fast changes in load current. If the buffer capacitor were located on the power-supply chassis, the impedance of interchassis connections could minimize its effectiveness. This example illustrates the importance of controlling chassis currents. Considerable experimentation is usually necessary to find the final configuration in which "ground loops" have been effectively eliminated. The empirical solution of these chassis-current problems often results in puzzling circuit

ground
loops

configurations; for example, in one power-supply circuit, a capacitor is connected between two points, each of which is obviously chassis ground. In another instance, a resistor is found in parallel with a straight wire. Considerable time can be saved in analyzing a schematic diagram if these unusual configurations are quickly identified and their function recognized.

Because gain in the feedback loop is such an important consideration, it is quite natural to suppose that a pentode, used as a series regulator, would further improve the regulator's performance. The essential difference between a pentode and a triode as a series regulator is that the pentode's plate voltage can vary quite widely for a given value of negative bias, without affecting plate current to any significant extent. This means, of course, that a pentode will provide regulation against changes in line voltage much superior to that of the triode. For this reason, pentodes are used as series regulators in two situations:

pentode
series
regulators

(1) When optimum regulation against line voltages is desired and the load current demand is fairly constant, and (2) When the range of line voltages from which the power supply must operate is very wide. This situation occurs primarily in bench-type power supplies, where a specified regulated output current must be maintained over a line-voltage range of several hundred volts. Assuming, for example, an input-voltage range of 400 volts, a triode-series regulator would require approximately a 200 volt change on its grid to keep the output current constant. This is obviously impossible for the triode; however, a pentode would require only a few volts of change in grid voltage to handle the same conditions.

Pentodes have certain disadvantages, when used as series regulators, which outweigh their advantages in all but the special conditions outlined above. In the first place, to operate satisfactorily, the screen-to-cathode voltage must be held constant. This requires a regulated supply voltage from some external source. More important, perhaps, is the additional heat generated by the screen grid. The total dissipation available in a tube is primarily determined by the size of the glass envelope, and is

heat
dissipation

only to a negligible extent, a factor of electrode structure. Above a certain envelope temperature, there is a likelihood that occluded gas will be released from the glass and produce a "gassy" tube. The maximum heat dissipation of the glass envelope at a given ambient temperature therefore approaches a fixed value, say for example, 25 watts. In the triode, the only heat generating elements are the heater and cathode assembly, and the plate. If in this example, the heater and cathode dissipate 5 watts, about 20 watts are available at the plate. In a pentode of the same size 5 watts again will be dissipated by the heater and cathode, leaving 20 watts to be dissipated by the screen grid and plate. Typically, the ratio of dissipation is about 4:1 so about 4 watts will be dissipated by the screen, leaving 16 watts available to the plate. Since it is plate dissipation which determines the amount of load current that can be drawn from the tube, it is clear that the pentode is less efficient than the triode as a series regulator.

plate
dissipation

It is also quite apparent that the output current capability of a vacuum-tube power supply depends primarily on the cathode-current rating and/or the maximum plate dissipation of the series-regulator tube. In a typical oscilloscope the load current is reasonably constant over extended periods, and the power supply is designed chiefly to regulate against changes in line voltage. The series-regulator tube is therefore selected on the basis of its cathode-current rating. It is then possible to determine the plate-dissipation limitations of the tube.

As an example, consider a +200-volt power supply with an output current of 100 mA and a typical input voltage-regulation range of $\pm 10\%$. The lowest voltage which can be permitted across the series tubes to maintain the desired output current will be approximately 50 volts. The rectifier output voltage therefore must go no lower than 250 volts if regulation is to be maintained. This represents the -10% condition. If 250 volts is 90% of the nominal voltage, under high-line conditions this voltage will rise by a factor of $2/9$ to 305 volts. At this time the plate-to-cathode voltage will be 105 volts and the plate dissipation will be 10.5 watts.

The limits thus calculated are quite rigid, and it may be found that the load current and/or maximum plate dissipation under worst-case conditions exceed the capabilities of a single series tube. In this case it is possible to connect two such tubes in parallel. In this fashion, maximum plate dissipation and cathode current capability are doubled, the μ of the tube circuit remains constant and the g_m is doubled. Thus regulation against load-current changes is improved by a factor of two. However, the added tube requires an extra heater supply and doubles the cost of the series regulator; so, while the use of parallel regulating tubes is not unusual, other means of increasing load current capability are usually employed.

shunt
resistor

As long as the load current is reasonably constant, a shunt resistor instead of a parallel tube can be used to increase load-current capability. It is perhaps natural to assume that the added resistor will tend to degrade the voltage regulator's performance under all conditions, but this is not necessarily true. Consider for example, the circuit in Fig. 2-17. This is simply the series-regulator portion of a voltage regulator. Again it is assumed that the regulated output voltage is 200 V, the input voltage from the rectifier is 300 V and the nominal load current is 100 mA. Shunt resistor R_s is chosen to pass approximately one-half the load current under normal operating conditions.

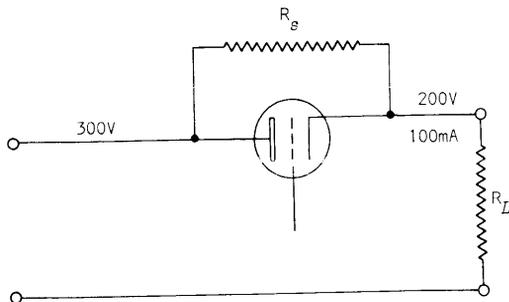


Fig. 2-17. Series tube shunt resistor.

Consider first the effects of load-current changes. It has already been shown that as long as load current remains within the regulating capability of the circuit, the cathode voltage will vary only a few millivolts with load-current changes. The plate voltage will also vary a few volts, due to the finite impedance of the transformer and rectifier; however, as a first approximation it can be said that the voltage across the series tube remains essentially constant. Since this same voltage appears across the shunt resistor, the current through the shunt resistor will also be constant. Therefore, as far as load-current changes are concerned, the shunt resistor plays no part in the circuit; changes in load current are matched by equal changes in the cathode current of the series tube, just as if the shunt resistor were not there. If, as a further simplification, the bias characteristics of the series tube are assumed to be linear, the change in bias voltage will be equal to the change in load current, divided by the g_m of the series tube, or

$$\Delta E_g = \frac{\Delta I_L}{g_m}.$$

Thus with a 20-mA change in load current and a series tube whose g_m is 5 mA/volt, a 4-volt change on the grid will be required. The change in output voltage will therefore be 4 volts divided by the combined gain of the comparator and error amplifier.

The same results will be obtained if the shunt resistor is left completely out of the circuit, so it can be argued that the shunt resistor has no effect whatsoever on the regulator's performance. In actual practice, the nonlinear cathode-current versus grid-voltage characteristic of the series tube introduces certain effects not considered in this first approximation; but it is nevertheless true that for all practical purposes, the addition of a shunt resistor increases the load-current capability of the voltage regulator at very little cost and without impairment of regulator performance.

The shunt resistor does, however, restrict the regulator's capability to a smaller *percentage* of the load current. (The foregoing analysis of the regulator's performance was based on an *absolute* change in load current of 20 mA.) Once a substantial amount of load current is flowing through the shunt resistor, the regulator can no longer handle such a

large percentage of the load current. Consider, for example, the situation in which a nominal 100 mA of load current flows through the series tube and no shunt resistor is used. Typically, the series tube could handle current changes up to $\pm 20\%$, or a range of 80 to 120 mA without loss of regulation. If a shunt resistor is now added to the circuit, and it also passes 100 mA of current, the regulator's nominal load-current capability is increased to 200 mA. In this situation, in order to maintain a $\pm 20\%$ regulating capability, the series tube must handle a range of currents from 60 mA to 140 mA, since the current through the resistor is a constant 100 mA. Thus it is clear that the circuit must be redesigned if the former percentage range of load-current change is to be regulated. In the majority of cases, however, this restriction is not a serious consideration, since absolute changes in load current are the chief consideration.

To examine the effect of a shunt resistor on the voltage regulator's response to line-voltage changes, assume once more a supply-voltage rise of 20 volts and a μ of 2 for the series tube. Without a shunt regulator, the grid voltage must go 10-volts negative to maintain the output current constant. When a shunt resistor is added to this circuit however, the rise in supply voltage tends to increase the current flowing through the resistor. The grid of the series tube must therefore go negative by an additional amount to compensate for the increase in the shunt-resistor current. Thus, the shunt resistor does degrade the voltage regulator's performance in response to line-voltage changes. It has already been shown, however, that control of line-voltage changes is rather easily attained, so this disadvantage is not a serious one.

When the advantages of a shunt resistor are weighed against the disadvantages, it emerges as a very useful and inexpensive choice for increasing the load-current capability of a voltage regulator. For this reason shunt resistors are found in a majority of Tektronix power supplies.

One aspect of voltage-regulator performance that must be considered before proceeding further is the response of such circuits to extreme conditions, or those conditions under which regulating action ceases.

There are two possibilities which will cause failure to regulate: (1) Excessive changes in load current, and (2) Excessive changes in line voltage. The areas of interest are the circuit characteristics which determine the limits of regulating action and what actually occurs in the circuit when these limits are exceeded.

excessive
load-current
changes

For the first situation assume that for some reason, load current decreases, tending toward zero. In an effort to compensate for this change, the grid of the series tube is driven more and more negative. However, the E_g-I_p characteristic of the series tube shows that its grid must be driven negative by perhaps 100 V to cut the tube off entirely, (the condition which must be attained if load current drops to zero). This requires that the plate of the error amplifier also go extremely negative, or in other words, that it draw maximum plate current. Two responses are therefore possible in the error amplifier; either it will begin to draw grid current, or it will saturate. If the plate of the error amplifier drops by 100 volts, its plate-to-cathode voltage will be very low, and the grid-to-cathode bias must approach zero. Under these conditions, the grid will attempt to draw current; but because it is supplied from a high-impedance voltage divider, it cannot do so. The grid voltage is therefore clamped at some negative value and no further plate current can be attained. The feedback loop at this stage is effectively open and regulation ceases.

The other, and more likely possibility is that the error amplifier will simply "bottom out." This phenomenon occurs in a pentode when the control-grid voltage reaches a certain positive value. The cathode and screen-grid voltages are fixed, so that as the plate voltage gets lower and lower, the screen grid begins to draw more current, until eventually it absorbs all further increases in cathode current, and no change in plate voltage occurs. Again, the feedback loop is effectively opened, and regulator action ceases.

With high load current the most likely failure mode is that of the series tube's cathode-current or plate-dissipation rating being exceeded. If these ratings are adequate to handle long-term variations, the only remaining concern is the regulating

capability of the circuit in response to short-term variations. Here the most frequent cause of failure will be grid current in the series tube. As load current increases, the series-tube grid becomes increasingly positive with respect to its cathode until it eventually attempts to draw grid current. Again, as in the case of the error amplifier, the grid is connected to a high-impedance source which cannot pass grid current. Therefore, no further change can take place in grid voltage and regulating action ceases. Another possibility is that the error amplifier will go into cutoff in its attempt to supply the positive grid signal for the series tube. This possibility is rather remote, and in most regulators failure will occur first in the series tube as described above.

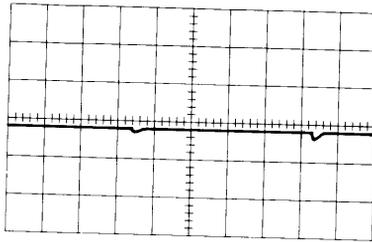
excessive
supply-
voltage
changes

Changes in supply voltage produce the same failure modes, but act in an opposite manner. If the supply voltage increases, the same failure mechanism as that caused by a reduction in load current is brought into play. As the plate voltage on the series tube goes higher and higher, more and more negative bias is required to maintain a constant load current. This creates the possibility of cutting off the series tube, if it is shunted by a resistor. Otherwise, there is the possibility of drawing grid current in the error amplifier or causing it to bottom-out. If the supply voltage drops, the plate-to-cathode voltage on the series tube gets lower and lower until the control grid begins to draw current, and regulation ceases.

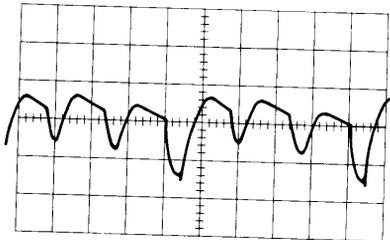
ripple
peaks

No matter what the cause of a particular failure mechanism, the extreme conditions are bound to occur instantaneously at the peaks of the ripple voltage. The negative ripple peaks represent the extreme conditions occurring from a reduction in supply voltage and/or an increase in load current. For instance, as the load current goes up, less and less bias is required at the series-tube grid for a given plate-to-cathode voltage; if the supply voltage drops, the plate-to-cathode voltage goes down and the bias must go less negative to maintain the required current.

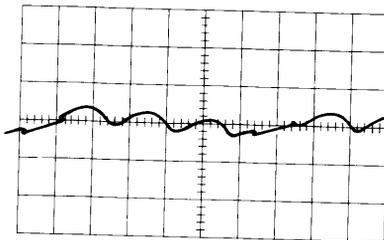
By applying the same reasoning it can be seen that the positive ripple peaks represent the extreme in high-supply-voltage or low-load-current conditions. It follows that regardless of the failure mechanism involved, it will first be apparent in one or the other of the AC ripple peaks. The first symptom of loss of regulation, therefore, is an increased AC ripple in the regulated output, because the ripple peaks are not being removed, see Fig. 2-18.



(A) LINE VOLTAGE 96V AC.
NEGATIVE RIPPLE PEAKS
JUST BEGINNING TO APPEAR.



(B) LINE VOLTAGE 94.5V AC.
NO REGULATION OF NEGATIVE
RIPPLE PEAKS. (VARIATION
IN PEAK AMPLITUDES DUE TO
RAPID FLUCTUATION IN RMS
VALUE OF LINE VOLTAGE.)



(C) LINE VOLTAGE 135V AC.
POSITIVE EXTREMES OF RIPPLE
VOLTAGE NOT REGULATED.

SWEEP SPEED: 5ms/div
VERTICAL LENS: 0.1V/div
NOMINAL LINE VOLTAGE: 120V AC

Fig. 2-18. Effect of regulation loss on CRT display.

negative-
output
voltage
regulators

It has been convenient thus far to confine the discussion of voltage regulators to positive-output vacuum-tube configurations. Although the basic design and operating principles of negative-output regulators are essentially the same as those of the positive type, there are certain differences in configuration which must be understood. This is especially true of vacuum-tube regulators, since the vacuum tube is a unidirectional device. It is therefore not possible to convert a positive-output regulator to a negative supply and have it work backwards.

Fig. 2-19 is a schematic drawing of a typical negative-output voltage regulator. Since a series tube requires a positive plate voltage it cannot be located in the output side of the regulator. It is therefore connected between ground and the positive side of the supply. A portion of the output voltage from the rectifier is dropped across this tube; the remaining portion becomes the regulated output voltage. With E_r volts across the rectifier and E_s volts across the series tube, the output voltage E_o equals $E_r - E_s$. The control circuits manipulate E_s so that E_o is constant, in spite of load-current or supply-voltage changes.

This configuration dictates a number of polarity changes in the error-signal feedback loop. An increase in load current or decrease in supply voltage causes the monitor to detect a less negative, or *positive* signal. The grid of the series-regulator

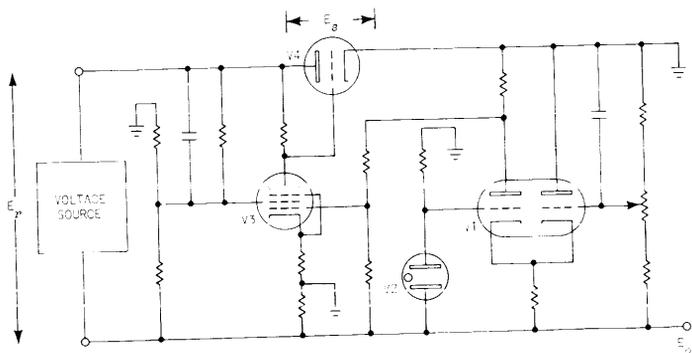


Fig. 2-19. Series-regulator negative output.

tube also requires a positive signal to compensate for the change. Therefore, no phase inversion can take place in the control circuit. Since the error amplifier's grid must receive a negative signal from the comparator, the comparator must invert the positive monitor signal. As explained in previous discussions of comparator circuits, phase inversion can be accomplished by taking the output signal from the right-hand plate of the comparator when the left-hand grid is held at a constant voltage.

However, DC-biasing considerations in a negative-output supply often lead to the configuration shown in the diagram under discussion. Here the VR tube is connected between the negative output and the left-hand grid of the comparator. A resistor between grid and ground forms a voltage divider with the VR tube. Any change in the output of the regulator is felt in its entirety on the left-hand grid of the regulator, since the drop across the VR tube is constant. The right-hand grid, connected to the voltage monitor, receives only an attenuated portion of the same change. The VR tube still provides a reference voltage for the regulator, even though the left-hand grid voltage is no longer constant with respect to ground. Instead, it is now constant with respect to the output voltage. Both grids move in the positive direction; but since the right-hand grid receives a much smaller signal than that on the left-hand grid, the net effect is a negative signal at the left-hand plate. Therefore, the negative signal for the error amplifier must be taken from the *left-hand* plate.

Since the net driving signal for the comparator is the difference between the change felt on the left and right-hand grids, it is advantageous to make the right-hand grid signal as small as possible. This is accomplished by connecting a speed-up capacitor between right-hand grid and ground.

It should be noted that, contrary to the views expressed in many textbooks, no advantage other than convenience in DC-biasing is gained by this configuration. A quick examination of the circuit, using arbitrarily chosen values of voltage and resistance, will prove that the net difference between grids of the comparator is the same whether the VR tube is connected between grid and ground or grid and output.

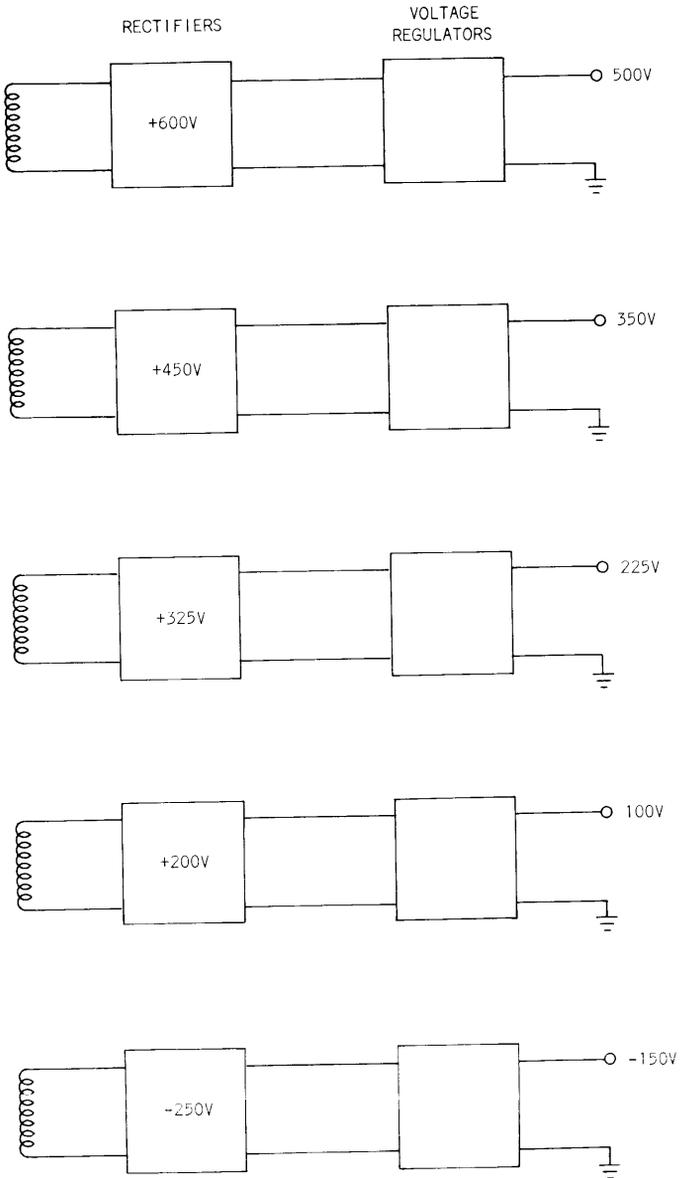


Fig. 2-20. Unstacked power supplies.

power-
supply
stacking

Instruments like an oscilloscope require a number of power supplies to provide the positive and negative operating potentials upon which the various specialized circuits depend. For a number of reasons it is advantageous to "stack" one power supply upon another; that is, to place the individual power supplies in series. The advantages are best illustrated by first considering the features of an unstacked power supply, as shown in Fig. 2-20. The regulated output voltages are typical of those required by a conventional oscilloscope. Assuming for purposes of simplification a 100-volt drop across the series voltage regulators, *rectifier*-output voltages of -250 volts, +200, +325, +450 and +600 volts will be required. It will be shown in a later chapter that high-voltage rectifiers require bulky high-voltage capacitors, and semiconductor diodes with high PIV ratings, both of which are quite expensive. However, when power supplies are placed in series, as shown in Fig. 2-21, the situation is improved considerably. Now the same positive regulated output voltages are attained by stacking two 125-volt and one 150-volt rectifier on the 200-volt unregulated supply. Nothing can be done about the negative power-supply rectifier since its positive side is connected to ground. It therefore remains in isolation. The advantages gained by stacking are not made without some expense, however. Each of the power supplies must be able to handle its own load current, plus all the load current in the power supplies above it in the stack. This requires larger diameter wire in the transformer secondary windings, and high-current capabilities in the rectifier diodes. This latter requirement is met quite easily by semiconductor diodes, however, and in general, the advantages offered by the stacking technique far outweigh the disadvantages.

It might well be asked why the unregulated, rather than the regulated outputs are stacked. The chief disadvantage of the latter system is that all the load current in the more positive supplies must pass through the series regulators of the less positive supplies. Thus all the load current drawn by the instrument must eventually pass through the series regulator of the +100-volt supply. This would require a very expensive series tube in a vacuum-tube regulator; in any case, considerable power would be

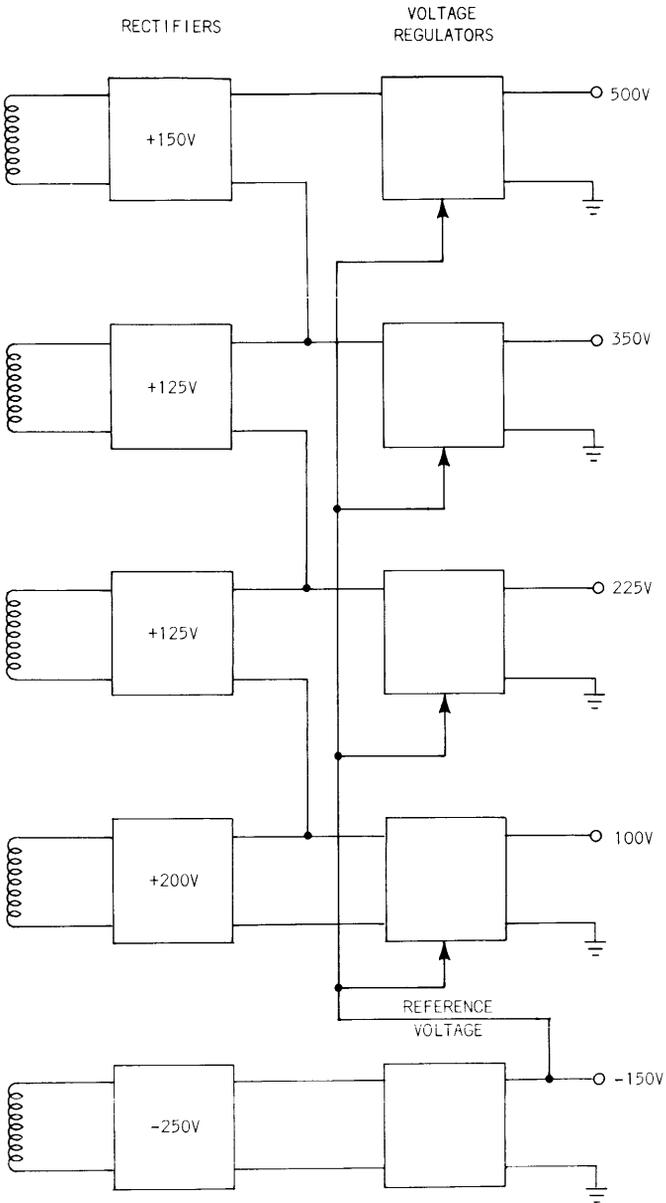


Fig. 2-21. Stacked power supplies.

dissipated in the series regulating device whether vacuum tube or semiconductor. Occasionally, however, it *is* advantageous to stack one power supply on the regulated output of another. First, compare the unregulated output of the isolated 500-volt power supply in Fig. 2-20 with that of the corresponding stacked power supply shown in Fig. 2-21. To simplify the problem, assume that all the unregulated outputs under discussion exhibit 10% ripple. The isolated power supply, with an unregulated output of 600 volts will have 60 volts of ripple. The stacked power supply also has 60 volts of ripple, since all the ripple is in phase and therefore *adds*.

$(\frac{200}{10} + \frac{125}{10} + \frac{125}{10} + \frac{150}{10} = \frac{600}{10} = 60.)$ Obviously, no

reduction in AC ripple is effected by stacking unregulated outputs. However, if, as in Fig. 2-22, a 250-volt unregulated supply is stacked on the 350-volt regulated supply, the unregulated 600-volt output will have only 25 volts of ripple (10% of 250 volts). This technique permits a much simpler design in the series regulator, and can be employed whenever the load current of the higher voltage supply is low enough to prevent overloading the series regulator in the lower supply.

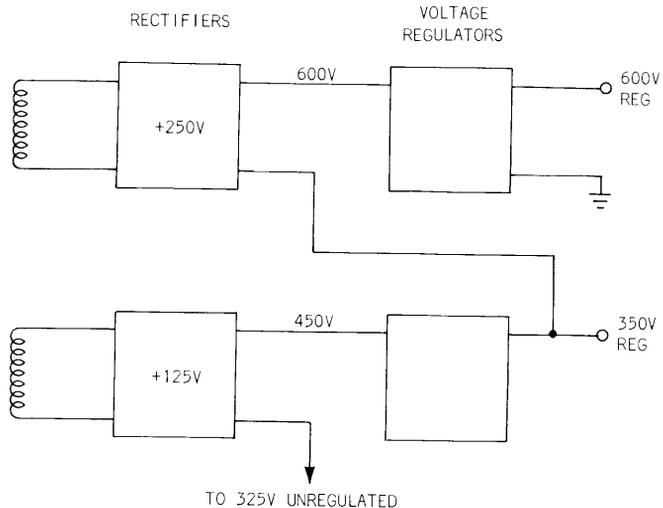


Fig. 2-22. Stacking on regulated output.

Having considered the unregulated side of a stacked power supply, it is next in order to explore the possibilities of simplifying the regulated side.

voltage-
reference
requirements

With a number of power supplies in the same instrument, there is no point in using a VR tube or zener diode as a voltage-reference source in each supply. It is sufficient that one voltage-reference component be used to establish the reference for one supply; the regulated output of this supply can then be used as the voltage reference for all the other supplies (Fig. 2-21). In Tektronix instruments the voltage-reference component is usually found in the most negative supply. The regulated negative voltage then provides the reference for the other supplies. This permits some sweeping simplifications in the circuits of the positive supplies. For example, examine the typical comparator circuit in Fig. 2-23. Note that the VR tube has been removed and the left-hand grid of the comparator has been connected to ground. The monitoring voltage divider, however, is now connected between the regulated positive output and the regulated negative voltage. Obviously there exists some point on the divider at which the potential is zero. When this point is connected to the right-hand grid of the comparator, an error signal will be generated only when the monitor voltage departs from zero. Since one side of the voltage divider is connected to the regulated negative output, only changes in the positive supply's output will produce an error signal.

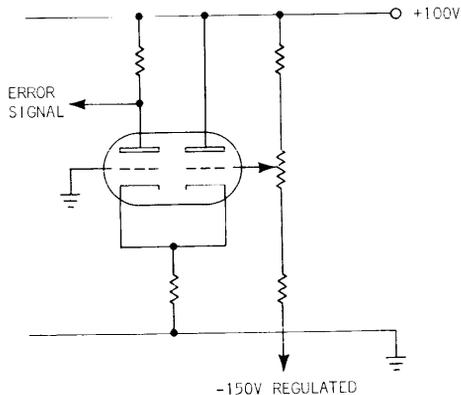


Fig. 2-23. Simplified comparator circuit.

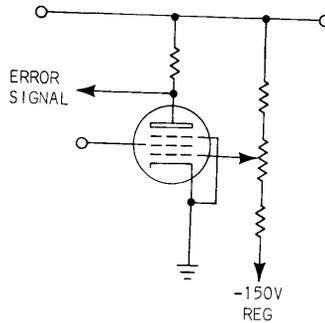


Fig. 2-24. Simplified control circuit.

When voltage-regulation requirements are not too stringent, an additional stage of simplification is still possible. (Fig. 2-24.) Here the comparator stage is eliminated completely and the monitor circuit is connected to the grid of the error amplifier. The pentode thus acts as both comparator and error amplifier. In effect, it compares the voltage on its grid with the voltage on its cathode (ground) and generates and amplifies an error signal when changes in the output voltage occur.

intermediate
voltage
sources

Like many electronic circuits, the voltage regulator requires a number of operating potentials whose levels fall somewhere in between the primary power-supply output levels. These intermediate potentials must be supplied locally, since it is obviously impractical to provide a separate power supply for every one of them.

Obviously any source of power must supply both voltage and current. However, there are many occasions when one of these quantities is of greater concern than the other. For instance, the driving signals for a vacuum tube and transistor may be identical; but where the vacuum tube is concerned, the investigator is primarily interested in the grid-signal voltage, while base-driving current is most often the point of interest in signals applied to a transistor.

Thevenin
equivalent
circuit

According to Thevenin's theorem, any line or network of impedances and generators, when viewed from any two points in the network, can be replaced by an equivalent voltage source and an equivalent impedance in series (Fig. 2-25). As an example, consider the simple circuit shown in Fig. 2-26. R_{int} represents the internal impedance of the source and R_C represents some circuit impedance across the source. The "open-circuit voltage" V_{OC} , (the voltage at points A and B looking back into the network) will obviously be 80 volts, due to the voltage-divider relationship of R_{int} and R_C .

The Thevenin equivalent impedance, R_{th} of the circuit is found by shorting out the generator and calculating the impedance of the circuit at points A and B (Fig. 2-27). This value is $\frac{50 \cdot 200}{50 + 200} = 40$ ohms.

Therefore, the equivalent voltage source can be represented by a circuit consisting of a voltage source of 80 volts in series with an impedance of 40 ohms (Fig. 2-28).

It is now possible to predict the behavior of the original circuit under a variety of conditions. Of particular interest are those situations in which

- (1) R_{th} is much larger than the load resistance,
- (2) R_{th} is much smaller than the load resistance.

"constant-
current"
source

In the first instance, as the value of R_{th} increases with respect to that of the load resistance, small changes in the load resistance have less and less effect on the current flowing out of the circuit. As the value of R_{th} increases, the circuit acts more and more like a "constant-current" source. While a true constant-current source can exist only in theory, the concept is quite useful in analyzing the operation of certain circuits.

"constant-
voltage"
source

In the second case, as the value of R_{th} decreases with respect to that of the load resistance, small variations in the load resistance have a decreasing effect on the output voltage of the circuit. If the value of R_{th} could be reduced to zero, the circuit would be a "constant-voltage" source. Again, although such a circuit can exist only in theory, the concept is useful in explaining the behavior of certain circuits.

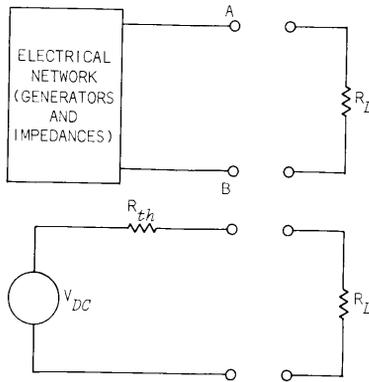


Fig. 2-25. Thevenin equivalent circuit.

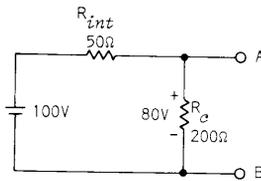


Fig. 2-26. Thevenizing a simple circuit -- step 1.

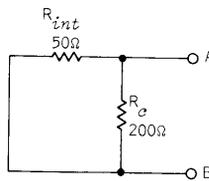


Fig. 2-27. Thevenizing a simple circuit -- step 2.

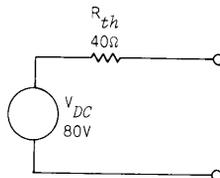


Fig. 2-28. Thevenizing a simple circuit -- step 3.

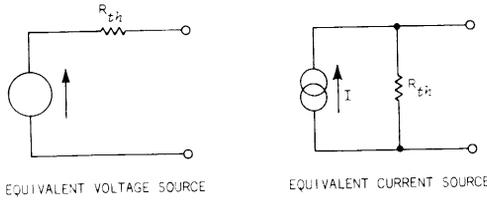


Fig. 2-29. Equivalent sources.

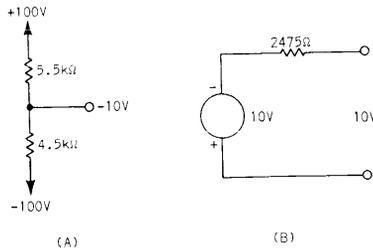


Fig. 2-30. Thevenized voltage divider.

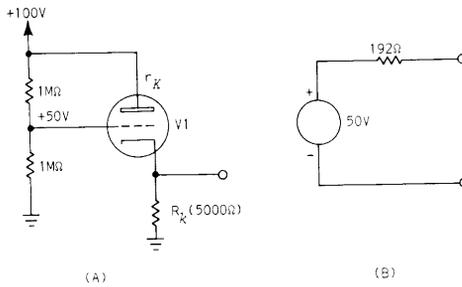


Fig. 2-31. Thevenized vacuum-tube circuit.

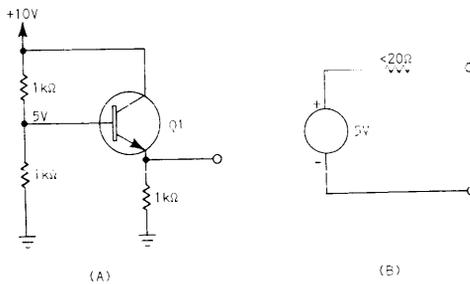


Fig. 2-32. Thevenized transistor circuit.

Fig. 2-29 shows how equivalent voltage and current sources are represented in much of the literature on this subject. Note that R_{th} is shown in *parallel* with a current source and in *series* with a voltage source. The reasons for these conventions are beyond the scope of this discussion. However, it can readily be seen that the current through any load connected to the current source will be determined by the ratio of load to internal impedance.

Fig. 2-30 shows a simple voltage divider connected between a negative 100-V and a positive 100-V potential. When reduced to its equivalent circuit it becomes a 10-V source in series with a 2475-ohm resistor. If the output terminals are shorted, about 3.5 mA will flow in the circuit. This circuit might well be satisfactory as a source of grid bias for a vacuum tube, but the small amount of current available from the circuit may be insufficient to drive the base of a transistor.

In the example above, a lower output impedance can be obtained by lowering the value of the voltage-divider resistors. The disadvantage of this procedure, however, is that large amounts of current would inevitably be wasted in the form of dissipated heat. A more satisfactory technique is illustrated in Fig. 2-31. A very high-impedance voltage divider is used to develop the grid bias for cathode-follower V1. Assuming a typical g_m of 5000 μ mho for V1, the equivalent circuit becomes that shown in Fig. 2-31B. Looking back into the circuit from the output terminals (V1's cathode and ground), the circuit's impedance is equal to the parallel combination of V1's internal cathode impedance r_K and its external cathode impedance R_k . The open-circuit voltage of the equivalent circuit will be about 50 volts. Since

$r_K = \frac{1}{g_m}$, the internal cathode impedance becomes 200 ohms. The impedance of the equivalent circuit is therefore $\frac{200 \cdot 5000}{200 + 5000} = 192$ ohms.

emitter
follower

An emitter follower can be used as a voltage source with even better results (Fig. 2-32). The output voltage is determined by the voltage-divider ratio, in this case 5 volts. The equivalent resistance of the circuit, looking into the transistor is the 1-k Ω emitter resistance in parallel with the transresistance R_t of the emitter follower. Assuming

a β of 50 and with 1-k Ω for each of the voltage-divider resistors, a value of 20 Ω can be calculated for R_t . Thus the equivalent circuit becomes a 50-V source with an internal impedance of 20 Ω .

It is not necessary to solve complex equations to determine the nature of the voltage regulator's equivalent. If typically, a change in load current of 50 mA causes the regulated output to move 1 mV, the output impedance of the circuit is obviously $\frac{\Delta E}{\Delta I} = R = 0.02$ ohms. This is a very low impedance indeed. The circuit clearly suits the definition of a constant voltage source, since its internal impedance is very small compared to any probable value of load resistance which may be connected to it.

operational
amplifier

The series regulator can also be viewed as an operational amplifier, in which voltage is amplified and held at the desired level by maintaining a null at the error amplifier or comparator inputs.

Fig. 2-33 shows a typical positive-output series regulator, together with its operational-amplifier equivalent. Resistor R3 is the feedback resistance, compensated for high frequencies by C1. Grid-limiting resistors R1 and R2 do not enter the input-impedance calculation. Two inputs are shown, since V2 is biased at a nominal -3 volts. Resistors R4, R5 and potentiometer R6 form a summing network at the (-) input.

The output voltage can be calculated by multiplying the separate input voltages by their separate R_f/R_{in} ratios. Since the grid of V1 is at a -3 volts, this amount must be subtracted from the -100 volts before the calculation is made. Thus, $E_{out1} =$

$$-97 \left(\frac{330 \text{ k}}{250 \text{ k}} \right) = -128 \text{ V}$$

Before calculating E_{out2} it is necessary to calculate the Thevenin equivalent of the second input circuit. The result (assuming that the pickoff of potentiometer is set midway between its extremes) is +12.5 V in series with 6.925 megohms of resistance. After once more subtracting the -3 volts at the + input from E_{in2} and multiplying the results by the R_f/R_{in} ratio, an output voltage (E_{out2}) of -0.22 V is found. The

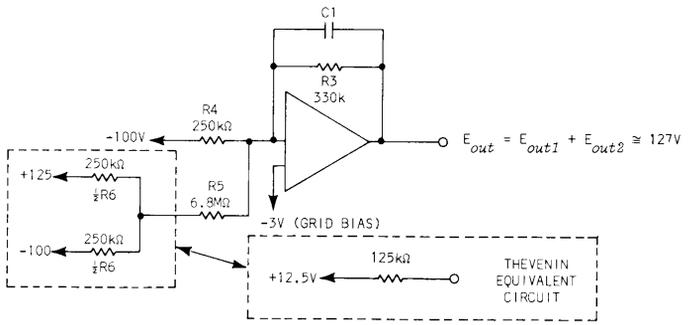
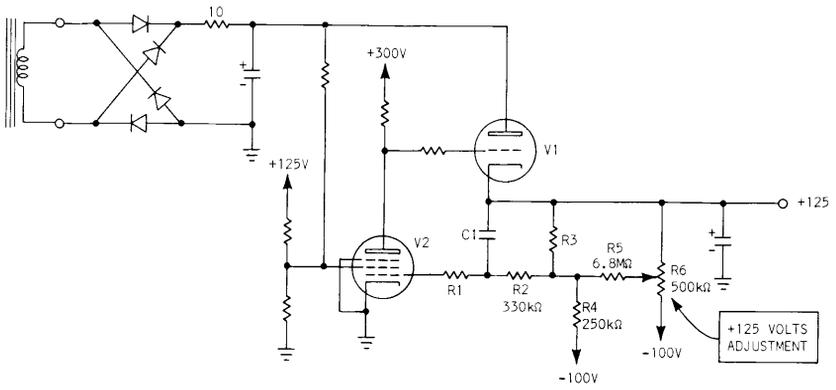


Fig. 2-33. Series regulator -- operational-amplifier configuration.

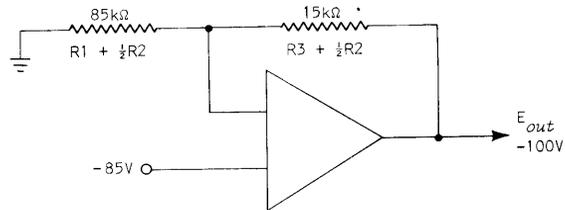
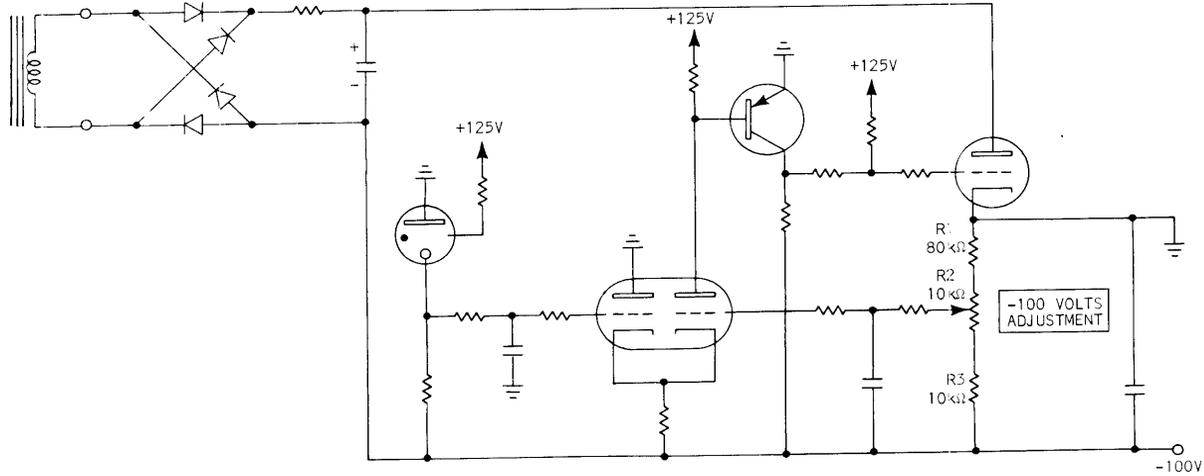


Fig. 2-34. Negative-output operational-amplifier configuration.

sum of E_{out1} and E_{out2} is therefore +127.78 volts. The actual output of this circuit is established at 125 volts through adjustment of R6. A quick calculation shows that from stop to stop, a range of approximately 5 volts above and below the nominal output voltage is provided by this adjustment.

The operational-amplifier configuration for a typical negative-output series regulator is shown in Fig. 2-34. Feedback resistance R_f is the sum of R1 plus 1/2 R2 when potentiometer R2 is centered, while R_{in} is the sum of R3 plus 1/2 R2. The input voltage is -85, developed across the VR tube. The operational amplifier is thus a noninverting type, with a gain

greater than one. The output voltage is $E_{in} \left(\frac{R_{in} + R_f}{R_i} \right)$
 or $-85 \left(\frac{85 \text{ k}\Omega + 15 \text{ k}\Omega}{85} \right) = -100 \text{ V}$.

transistor-
 ized
 voltage
 regulators

Although the operating principles of transistorized voltage regulators are substantially the same as those of the vacuum-tube type, a much wider variety of configurations is possible. This is due principally to the existence of both PNP and NPN transistors. In essence, the transistor becomes a bidirectional device. It is thus possible to construct a negative power supply with the series-regulating device in the output side of the regulator -- a configuration already shown to be impossible with vacuum tubes. Because of this flexibility, an attempt to describe all the possible variations in transistorized-power-supply design would obviously be impractical. However, it is possible to point out the essential differences between vacuum-tube and transistor-circuit designs. These differences are dictated solely by the characteristics of the transistors used in the circuit. The discussion will be largely confined to NPN circuits, mainly to emphasize the analogy between transistors and vacuum tubes.

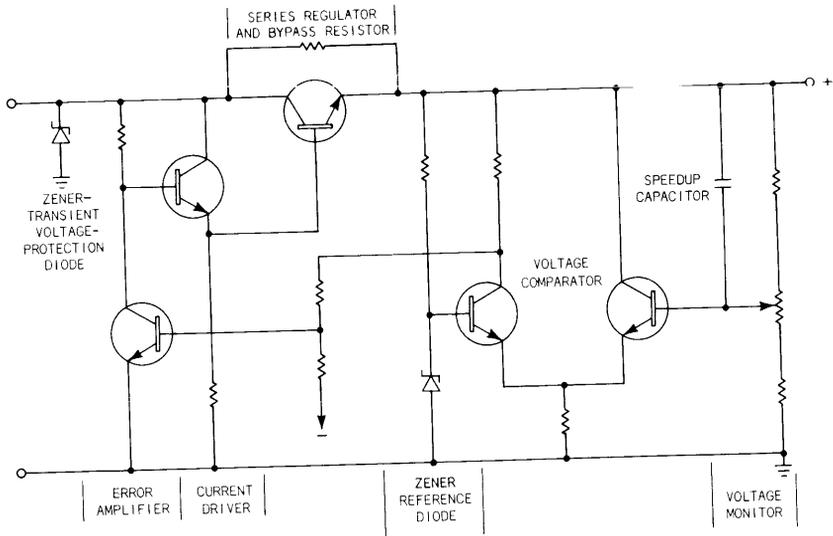


Fig. 2-35. Transistorized series regulator.

series
voltage
regulator

Fig. 2-35 is a schematic diagram of a basic transistorized series voltage regulator. The circuit elements, with one notable exception, bear a close resemblance to their vacuum tube counterparts. However, it is very important to remember that transistors are usually regarded as current-driven devices, whereas vacuum tubes are customarily treated as voltage-driven devices.

In the following discussion the assumed values of voltage and load current as well as those of various circuit components are typical of many transistorized power supplies.

Because the series-regulating transistor is a current-driven device, it requires a substantial base current to control the emitter, or load current. With a characteristic *forward-current gain*, or β of 50, and an emitter current of one ampere, roughly 20 mA of base current must be supplied by the control circuit ($I_b = \frac{I_e}{\beta}$). This volume of current is clearly not available at the collector of the error amplifier, so a low-impedance source must be provided to supply it. For this reason transistor regulators include a circuit element not found in vacuum-tube supplies -- an emitter follower between the error amplifier and series-regulating device.

The same current-driving problem is encountered in the voltage-comparator circuit. In a vacuum-tube regulator, the voltage-divider monitor has a resistance on the order of several hundred kilohms. In a transistorized regulator, a much lower impedance, say 5 to 10 kilohms, must be used so that the base-driving current for the comparator represents only a small fraction of the total current in the voltage divider. The same consideration applies to the value of the speedup capacitance. To be effective at all, the capacitor impedance must be small compared to the divider impedance at frequencies at or above the ripple frequencies. Since the resistance of the voltage divider is 30 to 50 times smaller than those in vacuum-tube regulators, it follows that the values of the speedup capacitors must be 30 to 50 times larger. Thus, where a 0.1- μ F capacitor would be encountered in a typical vacuum-tube regulator, a 3 to 5- μ F capacitor is required in a transistorized supply. Only electrolytic capacitors can provide

these high values without becoming excessively bulky. To minimize the leakage problems associated with the electrolytic type, tantalum capacitors are usually employed in this application. (In trouble-shooting a power supply it is often wise to disconnect the speedup capacitor. If the output voltage changes, the capacitor is leaky and should be replaced.)

The presence of the speedup capacitor introduces another problem not encountered in vacuum-tube regulators. (Fig. 2-36A.) If the output of the regulator is accidentally short circuited, the voltage at the top of capacitor C1 instantaneously drops to ground potential. The voltage across the capacitor cannot change instantaneously, however, so the potential at the base of Q1 (one half of the comparator) is driven negative by an amount approximately equal to the amplitude of the original supply voltage. This usually exceeds the base-to-emitter reverse-breakdown tolerance of any transistor likely to be used in the comparator circuit. The transistor will therefore suffer irreversible damage, unless it is deliberately protected from this eventuality.

protective
devices

The circuit shown in Fig. 2-36B provides such protection. It also serves to introduce a number of protective devices, whose presence in transistorized regulators is dictated by the relative vulnerability of solid-state components to excessive voltage and current conditions.

The protective components in Fig. 2-36B are diode D1 and resistor R1. Under normal conditions a slight "keep-alive" current flows through R1, keeping D1 turned on so that the capacitor can charge and discharge in response to small fluctuations in output voltage. However, if the output is short circuited to ground, D1 cuts off completely, opening the former discharge circuit. Now the principal discharge path for C1 is through R1. The capacitor voltage is thus divided between the resistor and the base-emitter junctions of Q1. By far the largest part of this voltage is dropped across the resistor, so that the reverse bias on Q1 is held to a safe level.

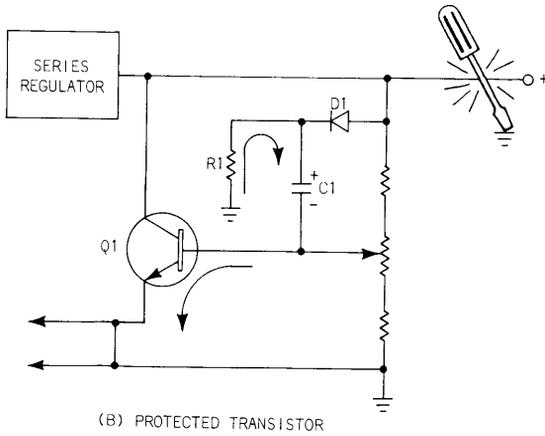
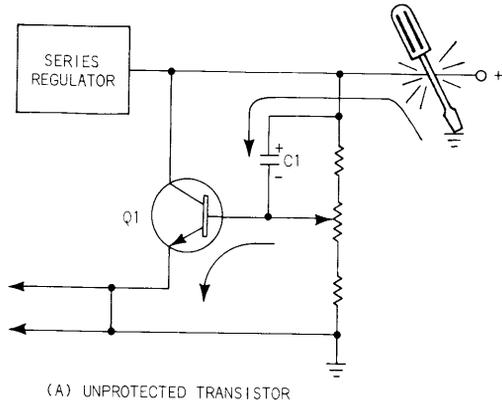


Fig. 2-36. Short-circuit protection.

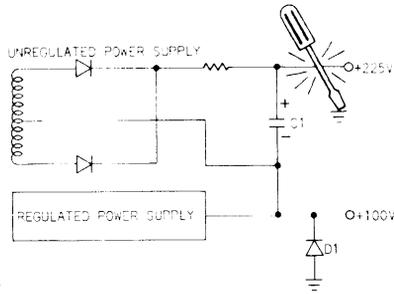


Fig. 2-37. Short-circuit protection.

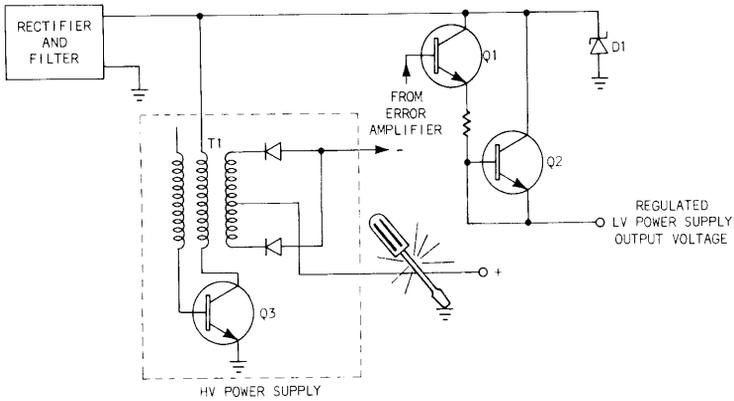


Fig. 2-38. Short-circuit protection.

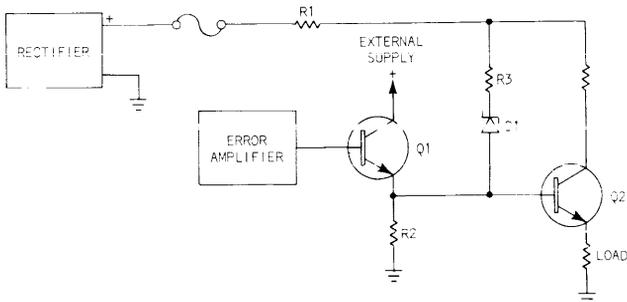


Fig. 2-39. "High-line" protection by zener diode.

In the previous chapter it was shown how certain advantages could be gained by stacking an unregulated supply on a regulated supply. However, this technique creates a potentially dangerous situation if the unregulated supply voltage is more than twice as high as the regulated voltage (Fig. 2-37). If the unregulated output is accidentally short circuited, the 125 V charge on filter-capacitor C1 will momentarily drive the regulated supply 25 volts below ground. Electrolytic capacitors and sensitive transistors in circuits supplied by the regulated output may thus suffer irreversible damage. However, by connecting diode D1 as shown in the diagram, a low impedance path is provided for the capacitor discharge, so that the regulated output is held at or above ground in these circumstances.

Most high-voltage power supplies are driven by an oscillator, for reasons which will be described in a later chapter. The collector voltage for the oscillator transistor(s) is applied through the primary winding of a high-voltage transformer. When this voltage is provided by a transistorized low-voltage supply, as in Fig. 2-38, accidental short-circuit protection is again required. In this case a zener diode (D1) is connected between ground and the collector supply of transistors Q1 and Q2, the current driver and series-regulating transistor. Under normal operating conditions the positive potential on D1's cathode is insufficient to cause zener breakdown, so the diode is essentially an open circuit. If, for some reason, the negative high-voltage output is accidentally shorted to ground, a high-amplitude positive step would be coupled from the secondary to the primary winding of transformer T1. Before the collector voltage of the transistors can rise to a damaging level, however, the zener diode goes into conduction, limiting the collector voltage to the reverse-breakdown voltage of the zener.

When the design of a regulator requires a series-regulating transistor to operate near its base-collector breakdown rating at nominal line voltages, provision must be made to protect it against "high-line" (high AC voltage on the power-transformer primary). In Fig. 2-39 zener-diode D1 is normally turned off, since the combined voltage drop across Q2's base-collector junction and its collector resistor is less than the zener-breakdown voltage.

If the line voltage rises beyond a safe level, however, D1 conducts. The resulting increase in current through R2 produces a positive signal at the base of Q2, allowing the base voltage to "follow" the rise in collector voltage. Although this effect holds the base-collector voltage of Q2 to a safe level, it disables the voltage regulator until such time as the line voltage returns to a lower value.

All power supplies are fitted with fuses which protect the transformer and other circuit components against short circuits in the output. These fuses do not "blow" instantaneously, however, and for a short period of time, very large currents pass through the series-regulator tube or transistor. This is not a serious problem in vacuum-tube regulators, but can cause irreversible damage to a series-regulating transistor. To protect against such overloads, many transistorized regulators include a current overload circuit, such as that shown in Fig. 2-40.

Resistor R1 is a very small resistor, connected in series with the load. Under normal conditions of load current, the voltage drop across R1 is insufficient to bias transistor Q2 into conduction. In this condition, only the collector current of comparator-amplifier Q1 passes through R2 to generate an error signal for the series regulator. If the load current exceeds the design limits of the regulator, however, the voltage drop across R1 biases Q2 into conduction. The collector current of Q2 now adds to that of Q1, producing a negative signal at the control of the series regulator which reduces the current output of the power supply.

temperature compensation Although temperature compensation is not a protective technique in the strictest sense, it does give rise to a number of odd configurations which are found almost exclusively in transistorized power supplies. One example of this technique has already been encountered in connection with temperature-compensated voltage-reference diodes.

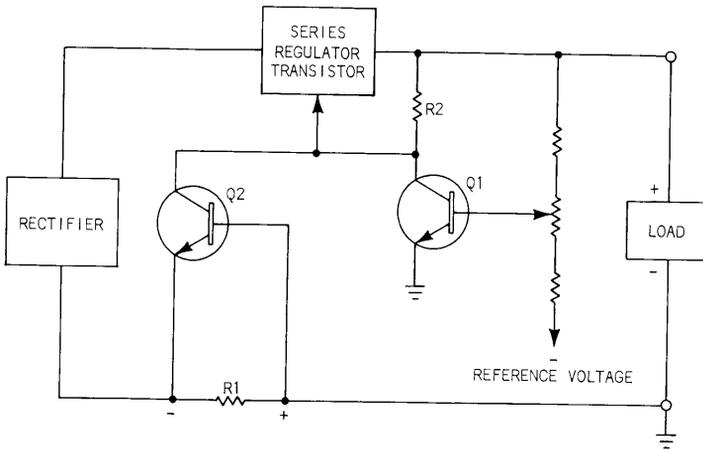


Fig. 2-40. Current-overload protection.

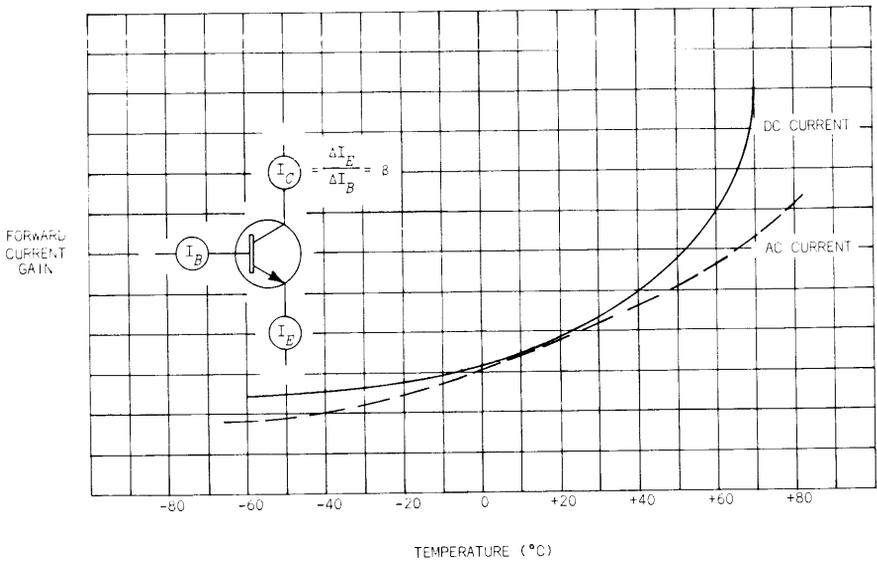


Fig. 2-41. AC-DC current gain versus temperature.

Like the zener diode, the performance of a transistor is significantly affected by its operating temperature. This temperature, in turn, depends largely upon the amount of power it is called upon to dissipate and upon the environmental (ambient) temperature in which it operates. Power-dissipation capability can be enhanced by use of a heat sink. However, although this technique increases the range of current a given transistor can handle, it does not compensate temperature fluctuations in the transistor itself. These fluctuations have a direct effect on the forward current gain of the transistor as illustrated in Fig. 2-41.

As explained previously, the conductance of a common semiconductor diode also varies directly with temperature. A diode can therefore be connected in a transistor circuit to compensate for the effects of temperature changes. This technique is illustrated in Fig. 2-42.

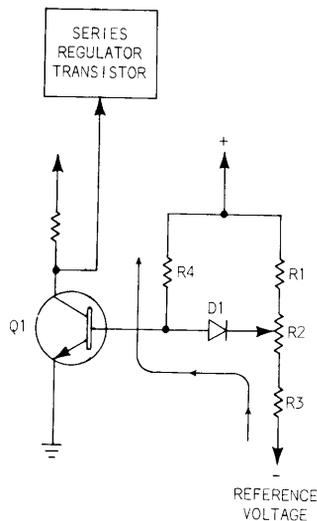


Fig. 2-42. Temperature-compensated transistor.

Transistor Q1 is a comparator-amplifier whose collector voltage should always be proportional to the regulated output voltage. Without compensation, however, a rise in Q1's operating temperature would result in increased collector current and a corresponding decrease in the forward bias applied to the base of the series-regulator transistor. Thus, without a change in line voltage or load-current demand, the current output, and therefore the output voltage, of the regulator would fall below its former value.

Diode D1, acting in conjunction with resistor R4 and the bottom half of the monitoring voltage divider, offset the effects of temperature changes in the following manner: As the operating temperature of the transistor rises, so does that of the diode, since it is situated as close as possible to the transistor and shares a common environment. As D1's temperature rises, its resistance decreases, causing an increased current through R4, R3, and the bottom portion of potentiometer R2. The increased voltage drop across R4 reduces the forward bias on Q1, so that its current at the increased temperature remains unchanged. Since the collector voltage also remains unchanged, the output of the regulator remains constant.

Clearly, for maximum effectiveness, the temperature coefficients of the transistor and diode should be equal, but of opposite sign. Perfect matching, however, is seldom if ever achieved, so that there are always definite upper and lower limits to the temperature range over which compensation is effective.

In Chapter 4 a new type of power converter called a DC-DC regulator is described. Included in its circuits are many transistors which operate as switches and drive inductive loads. This arrangement generates a need for additional protective circuits, some of which are quite complex. A better understanding of these circuits can be obtained when the overall operation of the power supply is thoroughly understood. Discussion of these protective circuits will therefore not be included in this chapter, but will be incorporated in an explanation of the DC-DC regulator's operation.

NOTES

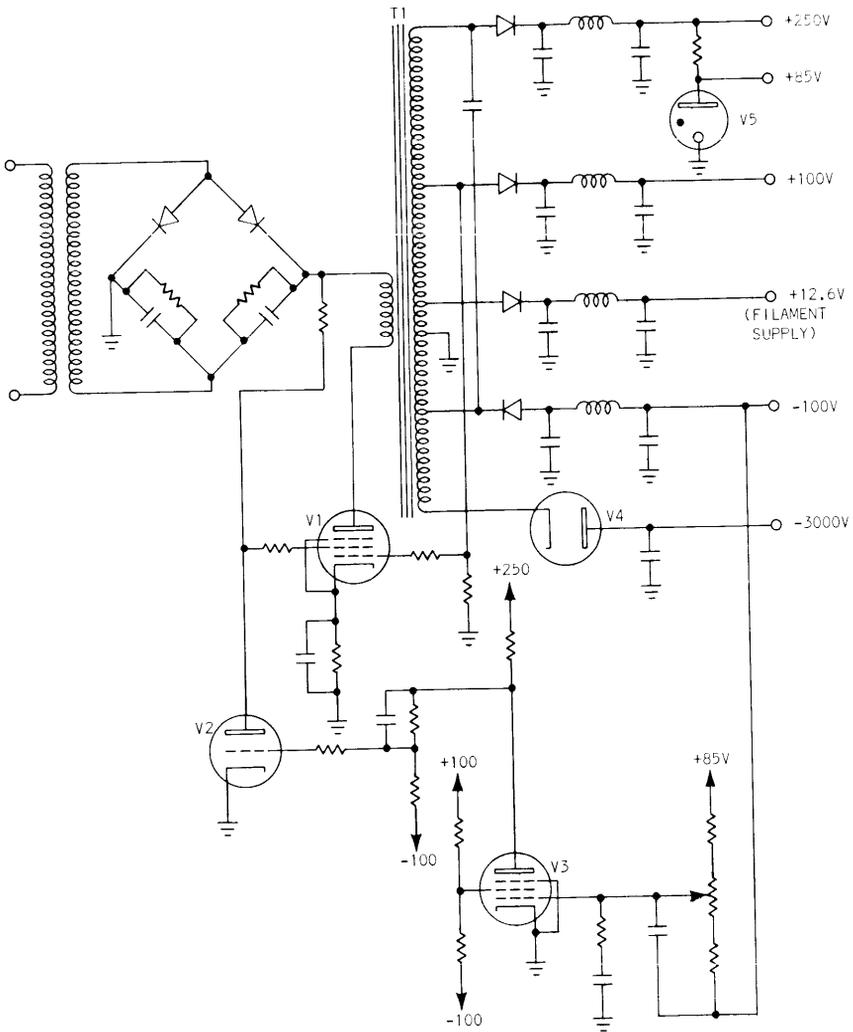


Fig. 3-1. PRLV power supply.

3

THE PRLV POWER SUPPLY

The primary-regulated low-voltage (PRLV) power supply represents one of the evolutionary steps in the trend toward more versatile, lightweight, high-efficiency power supplies. In most respects it resembles a typical high-voltage power supply of a conventional instrument, such as the one discussed in the next chapter. Because the operation of a high-voltage power supply is so closely related to the cathode-ray-tube (CRT) circuits, a detailed discussion of circuit operation is mandatory in that chapter. Therefore, only certain distinguishing features of the PRLV power supply will be pointed out here.

primary
power

Primary power for the PRLV power supply is generated by high-frequency oscillator V1, Fig. 3-1, whose plate supply (approximately 500 V) is furnished by a full-wave voltage doubler. The oscillator's output is stepped up in HV transformer T1, whose secondary winding supplies a number of half-wave rectifiers. Only the +85 V output is directly regulated through shunt-regulating VR-tube V5. One output at -3000 V is routed to the instrument's CRT circuits. The others provide the necessary low-voltage operating potentials for the rest of the circuits.

regulation

The principle regulating mechanism is provided by amplifiers V2 and V3 and their associated circuitry. Feedback from the secondary side of the transformer is taken from the -100-V output and applied to a voltage divider whose other side is connected to the regulated 85 V. Therefore, any change in voltage level of the -100-V supply is sensed at the grid of V3. This change is compared with the cathode voltage (ground) and an error signal is generated. After amplification in V2 the error signal is applied to the screen grid of V1 to increase or reduce the amplitude of V1's oscillations, depending on the phase of the error signal.

A second effect assists in the regulation process. Since the individual rectifiers are supplied by a single winding, a change in any output voltage changes the volts-per-turn ratio across the entire secondary and causes a corresponding change in the other output voltages as well. The feedback loop from the -100-volt supply senses this change, generates an error signal, and causes the oscillator output voltage to increase or decrease to offset the change. This regulating action, through the effect of "reflected impedance," is not 100% effective in actual practice. To find the actual quantitative response of the circuit to various conditions of loading would require a rigorous mathematical treatment outside the scope of this discussion. Generally, however, it can be stated that the corrective action produced in any one of the outputs will be proportional to the impedance it has in common with the -100-volt supply. Thus, if the +100-volt supply should drop 1 volt, the -100-volt supply would also sense a 1 volt change. Feedback action through the oscillator would thus restore both supplies to almost their original value. The same change in the +250-volt supply, however, would be only partially corrected.

Since no regulating devices (other than the one shunt regulator) are present to assist in AC-ripple removal, the filtering process must be as effective as possible. For this reason, π LC filters are used, which at the frequencies involved, reduce AC ripple to an acceptable percentage.

Considerable savings in space and weight are effected by using a HV transformer as the main power-supply transformer, and by the elimination of the large number of components required by separate voltage regulating circuits. On the other hand, the instrument containing this type of supply must be carefully designed to present a relatively constant load to the individual output potentials. Other limitations are imposed on the CRT circuits, as will be explained in the following chapter.

4

AC-DC POWER SUPPLIES

portable-
oscilloscope
character-
istics

The history of the oscilloscope is one of continuous expansion into new areas of application. This process has spurred development of a wide variety of instruments to fill specific needs. Among these was an instrument which could be used in the field under a wide range of operating conditions. This instrument, ideally, should (1) be relatively small and light, so that it could be easily carried; (2) operate on a number of different AC voltages and frequencies; (3) operate on a wide range of DC voltages, such as those generated in aircraft, diesel locomotives, boats or motorized vehicles; (4) operate on its own batteries when external power was unavailable; (5) offer performance comparable to that of laboratory instruments.

Until recently it was impossible to satisfy all these requirements simultaneously in one instrument. Savings in size and weight could only be made at a sacrifice in performance. Also, the high operating voltages and power consumption of vacuum-tube circuits made battery-powered instruments impractical.

power-
supply
efficiency

With the advent of solid-state technology came the first real opportunity to create a truly versatile, high-performance, portable oscilloscope. Before such an instrument could be realized, however, a new type of power supply had to be developed; for while conventional power supplies could be reduced in size and weight to a striking degree they had one disadvantage in common: Poor *efficiency*. That is, much of the power delivered to the instrument was wasted in the voltage-regulating circuits themselves. Although not an important consideration in laboratory instruments, the *efficiency* of a power supply (its ratio of input power to delivered power) determines how long a battery-powered instrument containing that supply can operate between recharging periods. Thus the design of battery-powered instruments called for an entirely new approach.

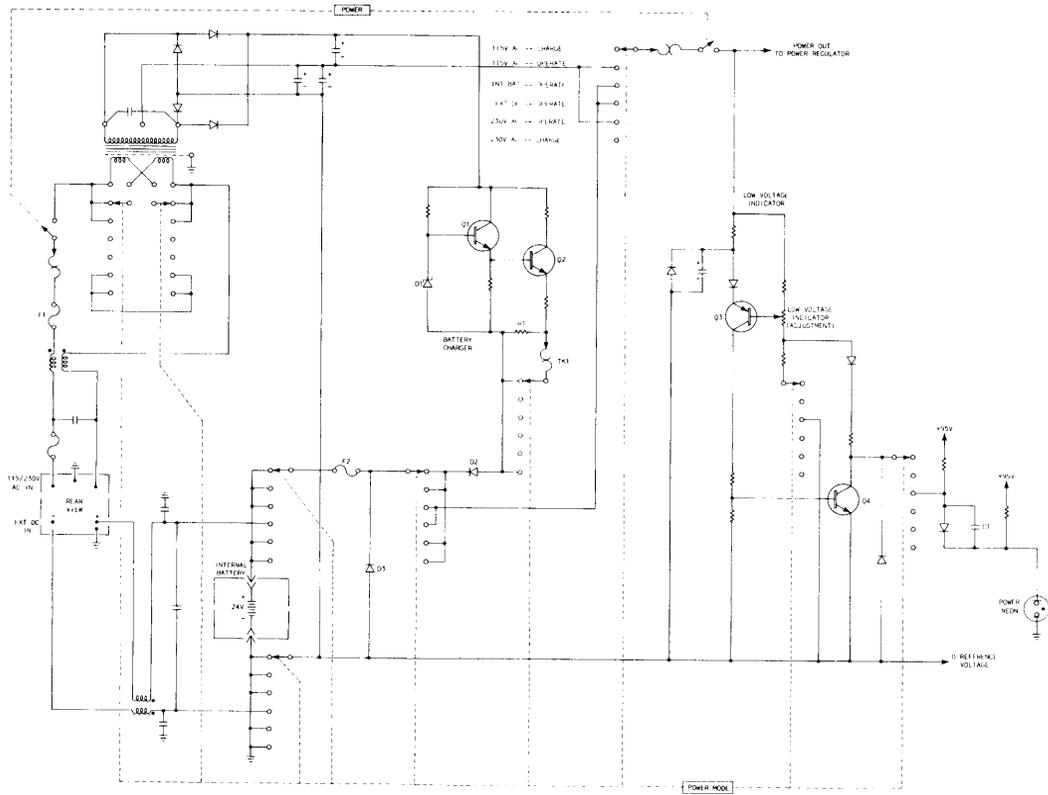


Fig. 4-1. AC-DC selector.

Tektronix has recently developed a number of new-design, high-efficiency power supplies. A typical example is provided by the AC-DC power supply of the Tektronix Type 422 oscilloscope, the first portable, battery-powered oscilloscope developed in the industry whose performance is comparable to that of laboratory-type models. Its power supply has an efficiency of approximately 80-90% (as compared to about 40% for a conventional power supply). It is also light, compact, and adaptable to a wide variety of AC and DC sources, including internal battery power.

The AC-DC power supply can be divided into two sections for purposes of discussion. The input section, or AC-DC selector, includes (1) the necessary input circuitry for AC and DC line voltages, (2) conventional rectifiers for converting AC line voltages to low-DC voltage, (3) a "battery-pack" consisting of a bank of rechargeable nickel-cadmium batteries or conventional dry cells, and (4) battery-charging circuits.

AC-DC
selector

The AC-DC selector is shown in Fig. 4-1. When operating on AC power, the line voltage is converted to 24-volts DC (to match the output of the power pack) by conventional rectification. A second rectifier supplies charging current to the power pack when the POWER MODE switch is placed in the 115-V AC-CHARGE or 230-V AC-CHARGE position. (Note that in either case the voltage on the transformer primary is 115 volts.) Transistors Q1 and Q2 and their associated components constitute a current-regulating circuit. Zener-diode D1 establishes a constant forward bias for emitter-follower Q1, which in turn supplies current drive for Q2. When the power mode switch is in either charging position, Q2 delivers about 400 mA of charging current to the batteries. When in the AC-OPERATE positions, current-limiting resistor R1 is switched into Q2's emitter circuit, thereby reducing the charging current to approximately 40 mA. Thermal relay TK1 senses the battery temperature and also switches R1 into the circuit when the temperature of the batteries rises above 105°F during charging. Diode D2 disconnects the charging circuit from the batteries when AC power is turned off, so that when battery or other DC power is turned on, the brief but high currents drawn by the rectifier filter capacitors do not damage Q1 (forward bias for Q1 would be developed across D1).

When operating on DC power the AC circuitry is completely bypassed. If the DC power connections are accidentally reversed diode D3 conducts and causes fuse F2 to blow. Note that none of the circuits are grounded to the chassis. This "floating" design allows the power supply to operate on either positive or negative DC power and/or sources that may be referenced to a high positive or negative potential.

Transistors Q3 and Q4 and their associated components form a low-voltage warning circuit, which causes the POWER-ON neon to blink when the battery voltage falls below the point where further operation would cause damage to the batteries. The circuit functions only when the POWER MODE switch is in the BAT-OPERATE position, since in other positions, Q3 is cut off by reverse bias. In the operate position, however, Q3 conducts whenever the battery voltage falls below the critical level, causing Q4 to be forward biased. This places Q4's emitter at approximate ground. Capacitor C1 and the POWER-ON neon then function as a relaxation oscillator, causing the POWER-ON neon to turn on and off at the oscillation frequency.

Whether the primary power be supplied by an AC line voltage, an external DC source, or the instrument's own power pack, the output of the AC-DC selector is a DC voltage. If the input to the selector is taken from the power pack or an AC line, the output voltage will be a nominal 24 volts; if taken from an external DC source, the output can range from 11.5 to 35 volts.

DC-DC converter

The second section of the AC-DC power supply is called a DC-DC converter. Its function is to accept the unregulated DC output of the AC-DC selector at any level between 11.5 and 35 volts, and convert it to a number of stable DC operating potentials for the instrument.

The DC-DC converter operates on what is termed the "energy-storage" principle. When voltage is applied to a purely inductive circuit, current in that circuit will eventually rise to infinity regardless of the voltage applied, since by definition the circuit contains no resistance to limit current. The *rate* of rise is governed, however, by the level of the applied voltage and the amount of inductance in the circuit, or for a given inductance, on the voltage alone. As current rises, a magnetic field builds up around the inductor which, at any given

energy
storage

instant (before the inductor's core becomes saturated) is proportional to the instantaneous value of current in the inductor. This magnetic field represents "stored energy," since if the applied voltage is removed at this instant, the collapsing field induces in the inductor itself a current in the same direction as the original charging current.

If the inductor happens to be the primary winding of a transformer, the stored energy in the primary winding will be discharged into the secondary winding every time the applied voltage is removed. Therefore, to produce a pulse of energy of a given amplitude in the secondary winding it is only necessary to remove the primary voltage at the proper time.

This principle is illustrated in Fig. 4-2. A DC potential E_1 is applied to the primary winding. At time T_1 , current in the primary winding has risen to level $Y-Y'$. At this time E_1 is cut off and the stored energy in the winding represented by the shaded area is discharged into the secondary of the transformer. Now a lower DC voltage is applied to the primary. In order that current rise to the same level as before, this second voltage must be maintained until time T_2 . However, when it too is cut off, the same stored energy is discharged into the secondary winding. If this process is repeated in rapid succession, a train of pulses will be generated in the secondary winding whose amplitude depends only on the duration of the applied voltage periods. When rectified and filtered, these pulses can be converted to a DC voltage in the same fashion as an AC sinewave.

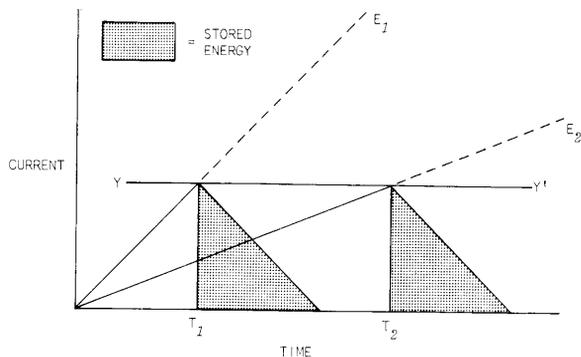


Fig. 4-2. Energy-storage principle.

A crude circuit for generating a given DC potential from a number of different DC input voltages is shown in Fig. 4-3. SW1 is opened and closed at a rate which allows the primary current to build up to a specified level. During the closed-switch period (A), no current flows in the secondary circuit due to the reverse polarity on diode D1. When the switch is opened (B) however, the collapse of the field around the primary winding causes its polarity to reverse, and induces a secondary current in the opposite direction. Thus, as long as the "switch-closed" period is made to compensate for different voltage levels, the current pulses in the secondary circuit will have the same amplitude, and the voltage drop across load resistor R_L will remain constant. The voltage waveform at the primary winding will, under these conditions, be seen as a rectangular pulse whose duration varies inversely with its amplitude (C). The primary current waveform will be a constant-amplitude triangular pulse whose ascending slope varies but whose descending slope is constant.

At this point the basic requirements of a practical DC-DC regulator should be fairly clear. It will consist essentially of an electronic switch, so controlled that it repeatedly turns the input voltage on and off, varying the "on" period in inverse proportion to the amplitude of the input voltage so as to maintain a constant current in the secondary circuit.

In actual practice, a number of rather complex circuits are required to achieve the desired result. Nevertheless, once the basic principle of operation is understood, the purpose of the individual circuits is quite clear and the explanation of actual circuit details that much easier to assimilate.

Fig. 4-4 is a block diagram of the complete DC-DC regulator. Before examining the details of the individual circuits it will be advantageous to establish the function of each circuit as it relates to the overall operation of the regulator.

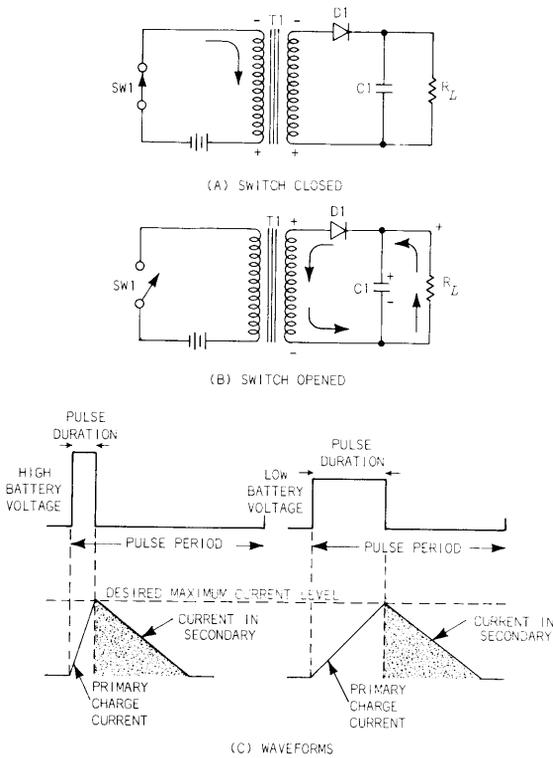


Fig. 4-3. Energy-storage switch -- principles of operation.

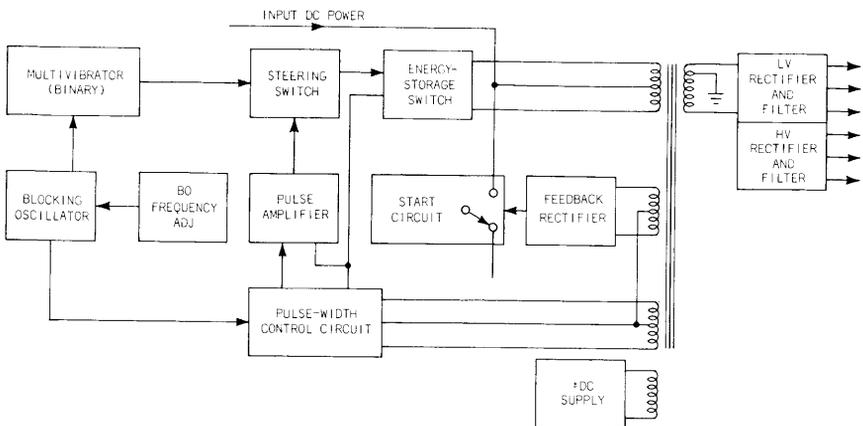


Fig. 4-4. DC-DC regulator -- block diagram.

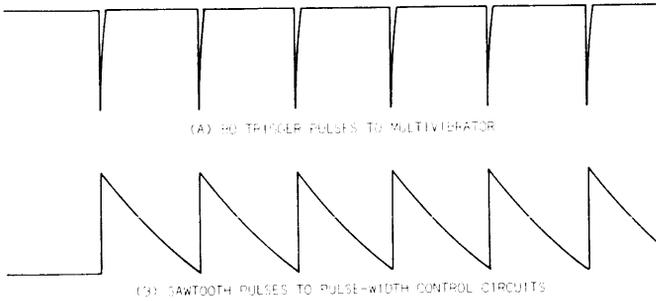


Fig. 4-5. Blocking-oscillator output waveforms.

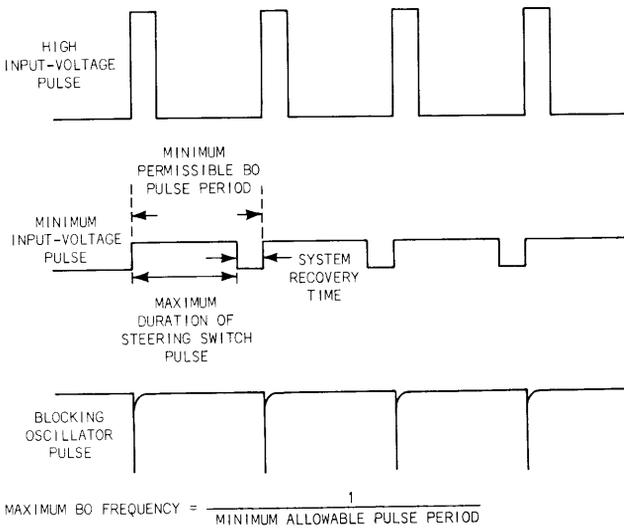


Fig. 4-6. Determination of maximum BO frequency.

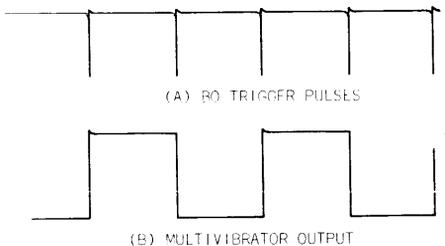


Fig. 4-7. Waveform time-relationship comparison.

blocking
oscillator

The purpose of the blocking oscillator (BO) is to establish the basic pulse (switching) frequency of the energy-storage switch. It does so indirectly, operating through the multivibrator and steering switch. It also "starts the clock" for pulse-width-control circuits which determine how long the energy-storage switch remains closed (Fig. 4-5). The BO frequency is adjustable over a narrow range. This control has little if any effect on the output voltage, but is included to assure a pulse period equal to or slightly longer than the pulse duration required by the lowest DC input voltage in the power supply's specified range (Fig. 4-6).

multi-
vibrator
(binary)

The multivibrator is triggered by the BO and produces a squarewave output voltage which enables the steering switch to guide the energy-storage switch. Fig. 4-7 shows the time relationship between the BO triggers and the multivibrator output.

steering
switch

Inputs from the multivibrator and pulse-width control circuits cause the steering switch to develop a pulse of the proper duration for the energy-storage switch. The time relationship of this circuit is compared to the BO and multi outputs in Fig. 4-8.

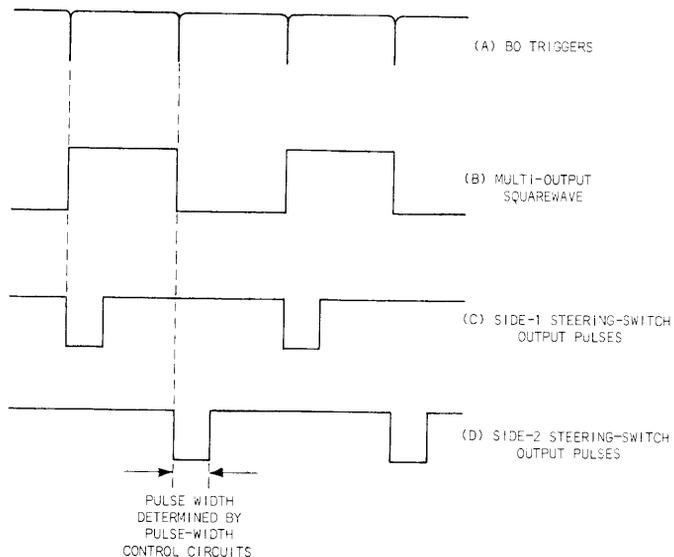


Fig. 4-8. Waveform time-relationship comparison.

energy-storage switch

The energy-storage switch electronically closes the circuit between alternate ends of the primary of transformer T1 and the DC power input at the center tap. The resulting primary current produces a series of positive and negative pulses in the secondary, which are conventionally rectified and filtered to produce stable DC potentials. The energy-storage switch is guided by the steering switch to deliver the same stored energy to the secondary of T1, regardless of the voltage applied to the center tap of the primary.

pulse-width control circuit

At the instant of firing the B0 generates a sawtooth-shaped current pulse in its emitter circuit, which is transmitted to the pulse-width (PW) control circuits (Fig. 4-9). This signal, when combined with a sample of the average current in the primary of T1, determines the point in time at which a tunnel diode switches from its high to low state. The tunnel diode in turn controls the action of the pulse amplifier, which opens and closes the current path between the center-tap of the primary in the steering switch and the DC power input. By this means, the steering-switch-pulse duration is controlled by the average energy level in the primary winding of transformer T1.

plus-and-minus-DC supply

The plus-and-minus-DC supply is a small power supply which provides an isolated source of positive and negative potentials for the sole use of the power regulator. It rectifies both the positive and negative pulses coupled to its secondary from the primary of the transformer T1.

Since the regulator operating potentials are generated in the regulator itself, an external DC supply source must be employed to initiate the regulator's

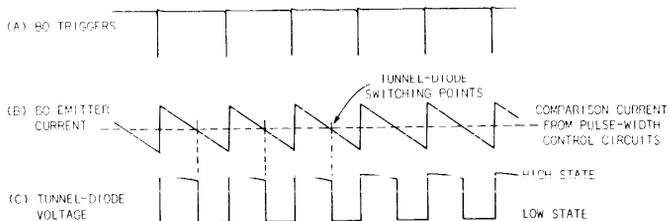


Fig. 4-9. Waveform time-relationship comparison.

start
circuit

action. The start circuit automatically connects the regulator to a portion of the DC power input until operation is initiated, then disconnects to allow the ± 12 -V power supply to take over.

Now that the various circuit functions have been established, it is possible to proceed with an examination of the individual circuit details. This discussion will begin again with the blocking oscillator, Fig. 4-10.

power-
regulator
blocking
oscillator

When power is applied to this circuit Q1 conducts through winding 1 of transformer T3, coupling a positive pulse through winding 3 to the base of Q1. This regenerative feedback action causes a very rapid rise in Q1 collector current which is inductively coupled to winding 2 and applied to the multivibrator as a negative pulse. When Q1 reaches saturation, the magnetic field in winding 1 collapses. A negative pulse is thus induced in winding 3, reverse-biasing Q1 and cutting off its conduction. The

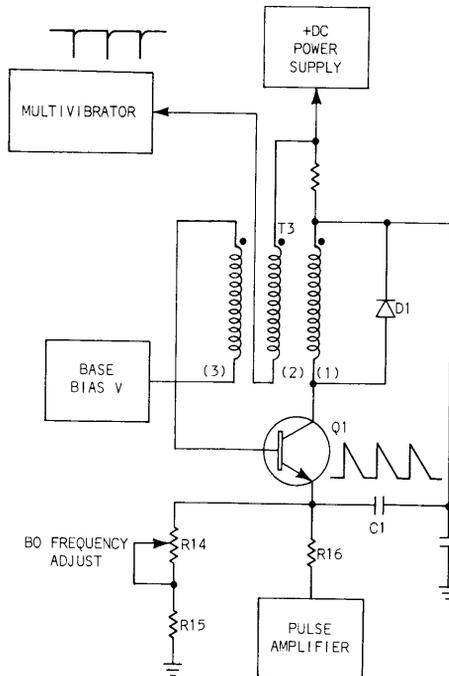


Fig. 4-10. Blocking-oscillator circuit details.

pulse is of very short (fractional microsecond) duration. The current generated in winding 1 by the collapsing field is allowed to discharge through D1, thus preventing Q1's collector from rising above its normal voltage.

Q1's emitter voltage follows the sharp positive rise on its base and C1 quickly discharges through Q1 to the small difference between emitter and collector voltages. When Q1 cuts off, however, the charge path for C1 is through R14 and R15, R16, and the impedance of the pulse amplifier. As C1 charges, Q1's emitter voltage falls. When it reaches a value less positive than the fixed base voltage, Q1 again goes into conduction and the cycle is repeated. Potentiometer R14, by varying the resistance in C1's charging path, allows adjustment of the B0 frequency over a limited range.

multi-
vibrator
(binary)

The negative output pulse acts as a trigger for the "flipflop" multivibrator (Fig. 4-11). In the quiescent state, one transistor conducts and the other is cut off. Commutating capacitor C1 charges to the difference in emitter potentials. The B0 trigger is applied at the junction of steering diodes D2 and D3. If, at this time, Q2 is cut off and Q3 conducting, D2 will be back-biased by the negative-to-positive voltage drop across collector resistor R4, while D3 will be forward biased by the +DC supply voltage. The negative spike from the B0 is therefore felt at the base of Q3 and the collector of Q2, holding both transistors in cutoff for the short duration of the B0 trigger pulse. When the trigger pulse ends, both transistors tend to go into conduction, but the charge on C1 causes Q2's emitter to be more negative than that of Q3. Q2 therefore goes more rapidly into conduction, and the negative-going voltage on its collector, coupled to the base of Q3, cuts off Q3 until the arrival of the next B0 pulse.

The output of the multivibrator consists of two trains of squarewave pulses (in push-pull), whose frequency is one-half that of the blocking oscillator. These pulses are applied as inputs to the steering switch (Fig. 4-12).

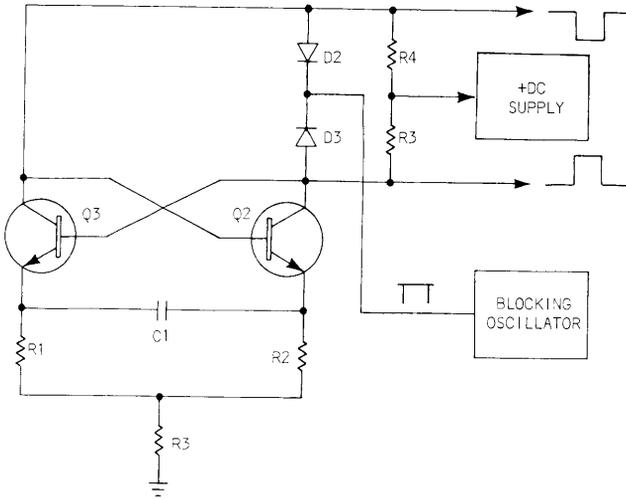


Fig. 4-11. Multivibrator (binary).

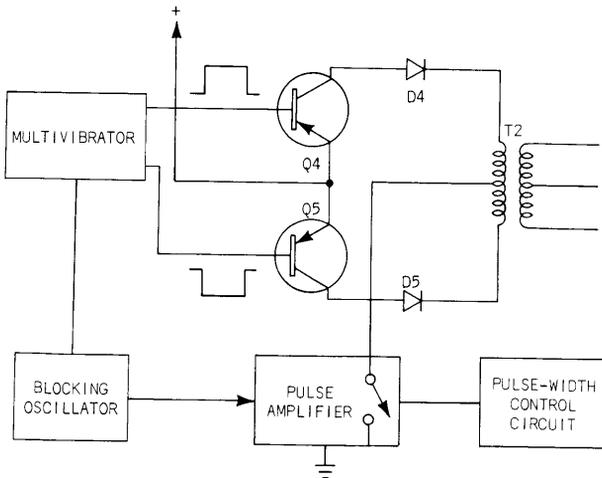


Fig. 4-12. Steering switch.

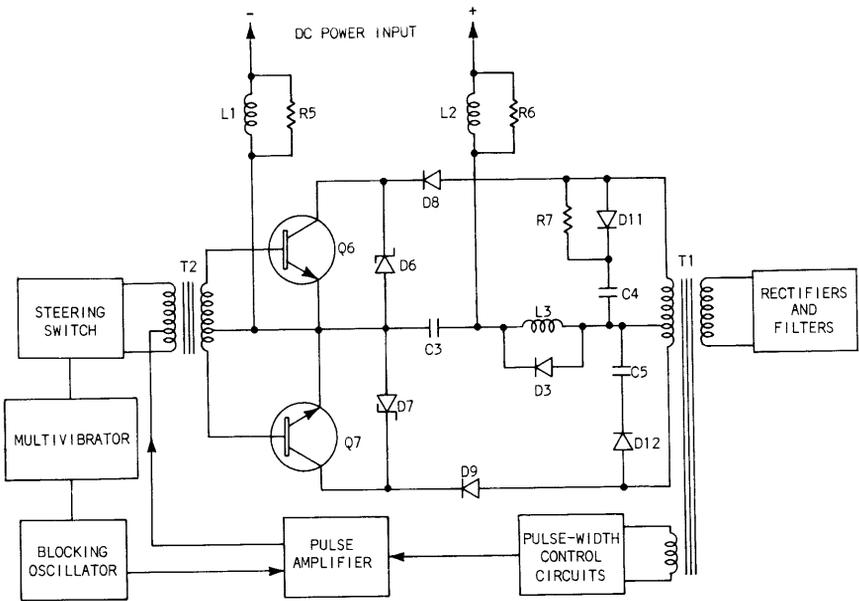


Fig. 4-13. Energy-storage switch.

steering
switch

The multivibrator pulses turn on either transistor Q4 or Q5 and drive the other into cut-off. At the same time the fast-rise portion of the sawtooth voltage generated in the B0's emitter circuit causes the pulse amplifier to connect the center tap in T2's primary winding to ground. (This action will be explained later.)

Current flows through one-half of T2's primary winding, inducing a voltage across the secondary winding. At a time determined (in the pulse-width control circuits) by the *average* primary current in transformer T1, the pulse amplifier opens the connection between T2's primary winding and circuit ground. Current in T2's primary winding therefore ceases. As the input squarewave changes polarity, the other transistor conducts, inducing a voltage of opposite polarity across T2's secondary winding.

When current ceases in T2's primary, due to the action of the pulse amplifier, the collapsing primary field momentarily generates a high positive transient in the collector circuit of the conducting transistor. However, diodes D4 and D5 are reverse-biased by these transients and disconnect the transistors from the primary windings. This action isolates the transistors and avoids a damaging base-collector forward bias caused by the positive transient.

energy-
storage
switch

Fig. 4-13 is a schematic diagram of the electronic energy-storage switch. Its function as previously explained, is to periodically apply DC input power to the primary winding of transformer T1, giving it sufficient time to accumulate a predetermined amount of energy. It then cuts off the power, allowing the stored energy in the primary winding to be discharged into the secondary winding.

Width-controlled pulses from the steering switch are coupled to the secondary winding of T2, causing transistors Q6 and Q7 to conduct on alternate pulses. As one transistor conducts it connects one end of transformer T1's primary winding to circuit ground. The center tap of this winding is connected through L3-D10 and L2-R6 to the DC power input. Therefore, for the duration of the width-controlled pulse, current builds up to a predetermined peak level and is discharged into T2's secondary when the conducting transistor is cut off. The peak current of each pulse in the primary winding is the same for all applied input voltages. Hence, the average rectified DC output voltage is held constant.

As transistors Q6 and Q7 are switched, sharp negative and positive spikes are generated in their collector circuits by (1) inductive coupling from the conducting to the nonconducting half of T1's primary winding and (2) the sudden collapse of magnetic fields, at turn-off time. Without protective measures these transients would apply excessive forward and reverse bias to the base-collector junctions of the transistors. Diodes D8 and D9 therefore are added to cut off and isolate the collectors from negative transients, while zener diodes D6 and D7 conduct to absorb the positive transients.

Capacitor C3 supplies added current to Q6 and Q7 when they first turn on, and inductors L1 and L2 isolate these current surges from the DC supply. Resistors R5 and R6 "damp out" oscillations in the inductors.

Diodes D11 and D12 provide a low-impedance charging path for C4 and C5. When the transistors again go into conduction D11 cuts off, allowing Q6's collector voltage to drop immediately without having to first discharge C4. R7 provides the necessary RC time for C4's discharge. However, the small stored charge in D11 still provides C4 with a discharge path for a very brief period. Inductor L3 absorbs this fast current change. When L3 saturates, its collapsing field (which would add to the inductive transient from the transformer primary winding) is discharged through diode D10.

Capacitors C4 and C5 protect the transistors at turn-off time. The collapsing field in the upper half of T1's primary winding tends to maintain current through Q6 at the time its collector voltage is rising. This combination of high voltage and current would cause large amounts of power to be dissipated by the transistor. However, the rise in collector voltage is delayed by the charging of capacitor C4 until Q6's collector current has fallen to a relatively low value.

pulse-
width
control

Fig. 4-14 is a schematic of the pulse-width-control circuit. In configuration it resembles a conventional regulated power supply. However, its *input* rather than *output* voltage is sampled by the error amplifier. Its output *current* (proportional to the average current in T1's primary winding) is applied to the pulse amplifier to assist in controlling pulse width.

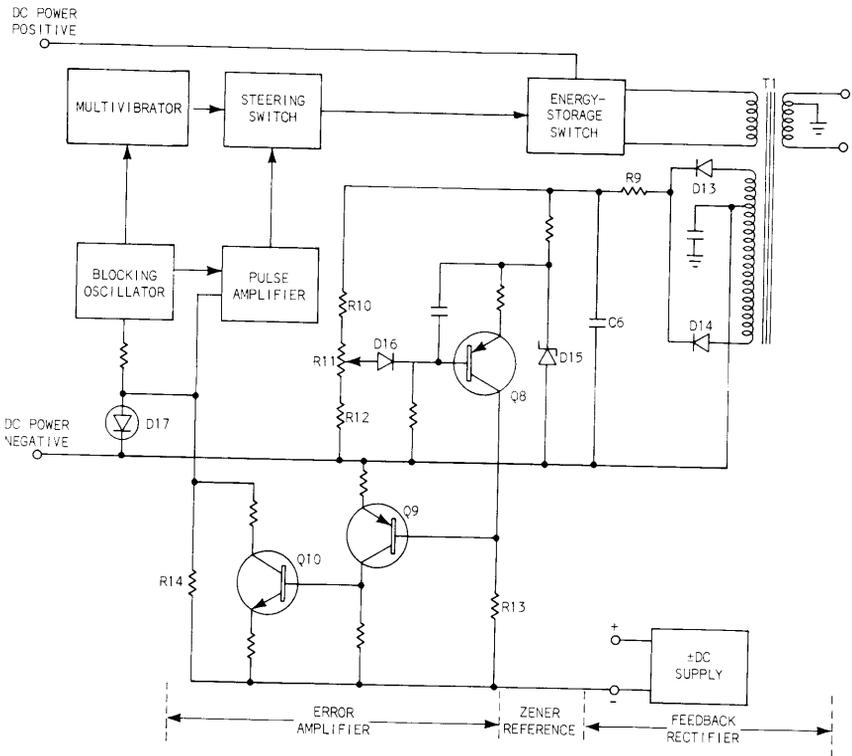


Fig. 4-14. Pulse-width control circuit.

The pulse-width-control circuit includes three subdivisions: (1) a feedback rectifier, (2) a zener reference, and (3) a three-stage error amplifier.

feedback
rectifier

The feedback rectifier consists of a center-tapped secondary (part of transformer T1) and diodes D13 and D14. R9 is a current-limiting resistor and forms an RC filter with capacitor C6.

The output voltage of the feedback rectifier is proportional to the peak current level of the pulse train generated in T1's primary winding by the energy-storage switch at any given time. This peak current level is, in turn, proportional to the width of the pulses generated by the steering switch. Therefore, the *output voltage* of the feedback rectifier is directly proportional to the *width* of the pulses generated in the steering switch.

zener
reference

Zener diode D15, connected between circuit ground and the emitter of transistor Q8, establishes a reference voltage with which the output voltage of the feedback rectifier is compared.

error
amplifier

Transistors Q8, Q9, and Q10 form a three-stage error amplifier. The negative output voltage from the \pm DC supply is passed through series-regulating transistor Q10 and its current-bypass resistor R14 to the pulse-amplifier circuit.

Assume that the positive output voltage from the feedback rectifier increases due to an increase in the regulator's input power. This increase will be sampled by potentiometer R11 and applied through temperature-compensating diode D16 to the base of Q8 as a decrease in forward bias. Q8's collector will become more negative, increasing the forward bias on Q9. Q9's collector will become more positive, applying forward bias to Q10 and an *increased* negative current output from the error amplifier.

Note that an *increase* in the output current of the energy-storage switch produces an *increase* in the output current of the pulse-width-control circuits. This increased current must act in the pulse-amplifier circuits to *decrease* the width of the steering-switch pulses and thereby *reduce* the "on-time" of the energy-storage switch. In this way the amount of stored energy discharged into the rectifying and filtering circuits will be lowered and the output voltage of the regulator returned to its specified level.

pulse
amplifier

The effect on the tunnel-diode switch of the output current from the error amplifier is best understood in conjunction with the action of the B0 and pulse-amplifier circuits. Fig. 4-15 is a schematic of the pulse amplifier, the final link in the pulse-width control regulating loop. The blocking oscillator is included in this figure to clarify the action of the pulse amplifier and tunnel diode.

As the blocking oscillator fires, the sharp rise in its emitter current switches tunnel diode D17 to its high state (Fig. 4-16A). The positive-voltage step on D17's anode is transmitted to the base of Q11, turning it on. Current through Q11 causes a voltage drop across collector resistor R14 and a negative signal is applied to the base of Q12. Q12 conducts, completing the circuit between the center tap of the steering switch and ground. Simultaneously a pulse from the multivibrator turns on one of the steering-switch transistors and current flows through one side of transformer T2, turning on one side of the energy-storage switch.

As the B0 trigger voltage drops to zero, Q1 cuts off and C1 starts to charge through three parallel paths: (1) potentiometer R16 and resistor R17, (2) emitter resistor R15 and tunnel diode D17, and (3) the impedance of the error amplifier.

If a large amount of current is available from the error amplifier (due to a high output from the energy-storage switch), capacitor C1 will charge rapidly, current through the tunnel diode will quickly fall to its valley-current rating causing the diode to switch to its low state (Fig. 4-16B). If current from the error amplifier is low (due to a low output from the energy-storage switch), D17 will remain in its high state for a longer period.

When D17 switches to its low state, the negative step transmitted to the base of Q11 cuts the transistor off. Its collector voltage rises, cutting off Q12 and opening the circuit between the primary of T2 and ground. The abrupt change in current through T2 turns off the energy-storage switch and the energy in T1's collapsing primary field is coupled to the secondary winding.

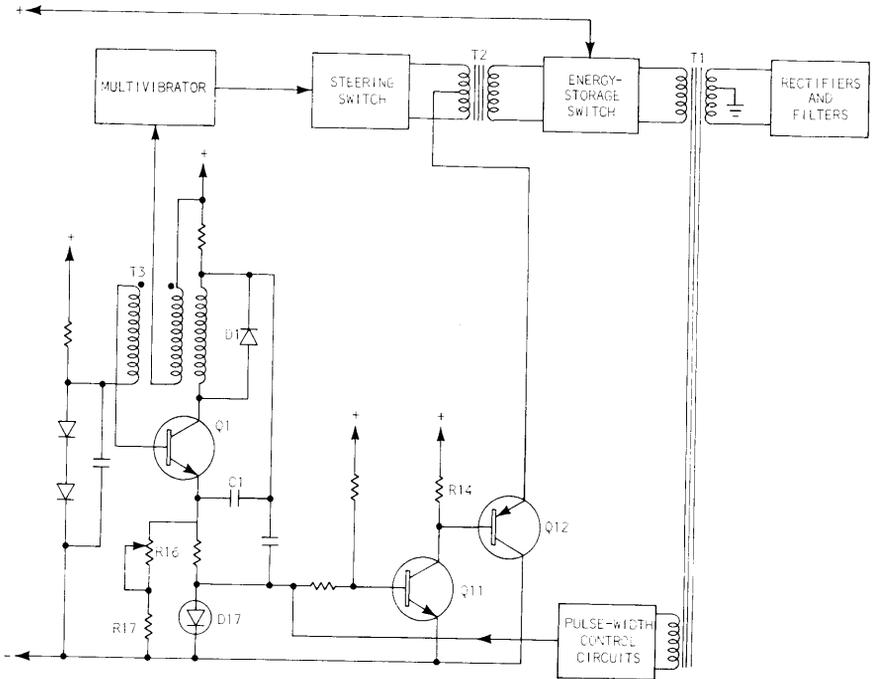


Fig. 4-15. Blocking oscillator and pulse amplifier.

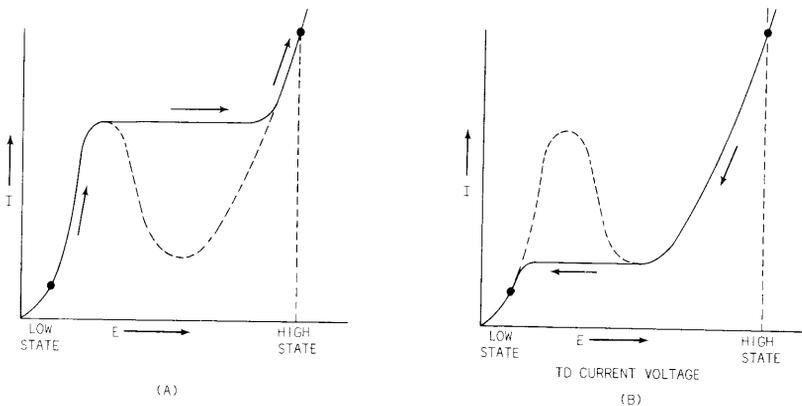


Fig. 4-16. Tunnel-diode characteristic.

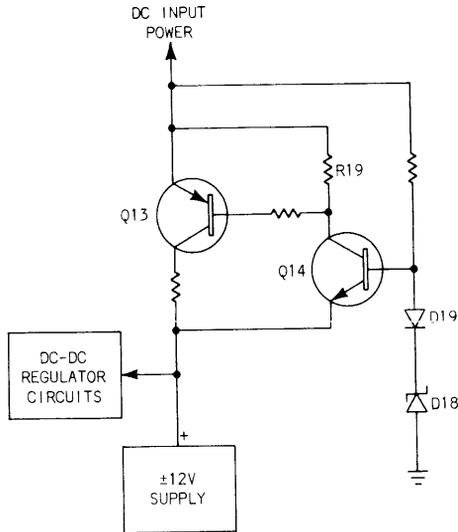


Fig. 4-17. Start circuit.

start
circuit

Fig. 4-17 is a schematic of the start circuit. When the instrument is first turned on there is no output from the isolated ± 12 -V supply to energize the B0. At this time, external DC power is applied to the emitter of Q13 and the collector of Q14. Forward bias for Q14 is established by the constant voltage drop across zener diode D18. As Q14 conducts, current through resistor R19 causes a negative signal to be applied as forward bias to the base of Q13 which is in series with the 12-V load. Q13 and Q14 now operate as a voltage regulator, applying about 7.5 V to the 12-V bus. This voltage is sufficient to energize the blocking oscillator, multi, steering switch, and energy-storage switch. The positive output from the ± 12 -V supply now rises to 12 V and is applied to the emitter of Q14 as reverse bias, cutting it off and causing its collector to rise to the level of the DC input. This positive voltage reverse-biases Q13 which also goes into cutoff. The ± 12 -V supply is now isolated from the DC input and is able to supply the operating voltages for the DC-DC regulator. Diode D19 compensates Q14 for temperature changes, thereby assisting in stabilizing the voltage at 7.5 V.

Briefly summarized, the DC-DC regulator operates as follows: DC power at some level between 11.5 and 35 volts is applied to the energy-storage switch. This switch is activated by pulses from the steering switch. The energy level in the primary winding is sampled by the pulse-width-control circuits which determine the duration of the steering-switch pulses. Since the duration of the steering-switch pulses determine the level of stored energy attained by each pulse of the energy-storage switch, a feedback loop is established which maintains a constant energy transfer to the secondary windings. Since the instrument for which the regulator is designed presents a constant load to the secondary circuits (rectifiers), the voltage across the load remains constant. Because power is consumed only when the energy-storage switch is closed, a very high efficiency is attained by this type of regulator.

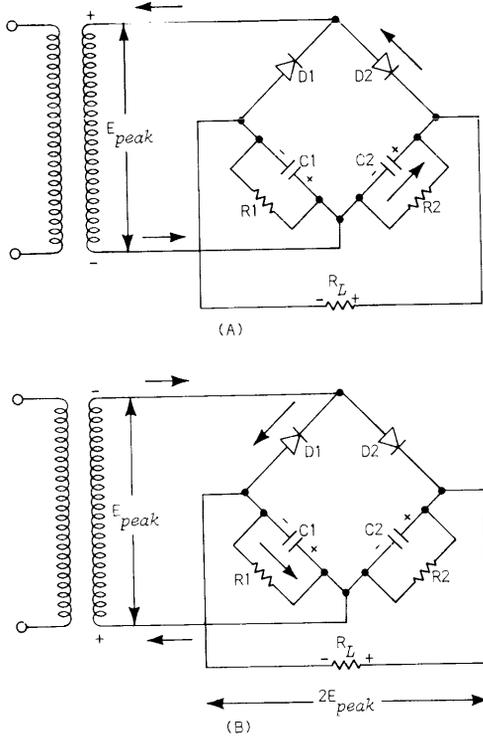


Fig. 5-1. Full-wave voltage doubler.

5

HIGH-VOLTAGE
POWER SUPPLIES

Any test instrument which uses a cathode-ray tube, such as the oscilloscope, requires DC operating voltages much higher than those provided by the power supplies discussed in previous chapters. Such voltages cannot be efficiently and economically attained by simply increasing the number of turns on the rectifier secondary winding. Such an expedient would call for large, expensive vacuum tubes, bulky and dangerous capacitors, heavily insulated transformer windings, and other objectionable features. The design of high-voltage power supplies is therefore considerably different than that of low-voltage supplies.

voltage
multiplier

The very low current required by CRT circuits makes it possible to use a circuit called a *voltage multiplier* to generate very high voltages at a considerable savings in cost and space. For example, the full-wave, or bridge doubler shown in Fig. 5-1 provides an output voltage twice as high as the peak AC voltage on the transformer secondary winding.

On the first half-cycle, as the top of the secondary becomes positive (Fig. 5-1A) diode D2 is forward biased and capacitor C2 charges to E_{peak} through D2. Diode D1 is reverse-biased. On the next alternation diode D1 is forward-biased and C1 charges through D1, while D2 is reverse-biased (Fig. 5-1B).

The charge across capacitors C1 and C2 is series-aiding, and their discharge through the load produces a voltage equal to twice the peak voltage of the secondary winding. Succeeding alternations replenish the charge on C1 and C2, lost in the form of load current. (R1 and R2 are "bleeder" resistors which discharge the high voltage on the capacitors when the instrument is turned off.)

The chief advantage of the full-wave doubler is that it utilizes both alternations of the secondary voltage and can thus supply more load current than the half-wave doubler discussed below. Unfortunately, the bridge configuration cannot be extended to produce higher multiplication factors. Their application in Tektronix instruments is confined to generating the plate voltage for certain high-voltage power-supply oscillators. Higher multiplication factors, however, can be achieved by using the "cascade" configuration.

cascade multiplier

The cascade multiplier is constructed by arranging components so that the transformer secondary and the rectifier output have a common reference. It is then possible to stack or "cascade" one stage of multiplication on another. The first step in this process is illustrated by the cascade doubler shown in Fig. 5-2A.

As the top of the transformer secondary winding becomes negative, diode D1 is forward-biased and capacitor C1 charges through diode D1 to (almost) peak voltage. On the next alternation the top of the transformer secondary winding becomes positive (Fig. 5-2B). Diode D1 is reverse-biased while diode D2 is forward-biased. The charge on C1 and the voltage on the secondary are series-aiding and charge capacitor C2, through D2, to the sum of the peak voltage across C1 and the transformer secondary winding.

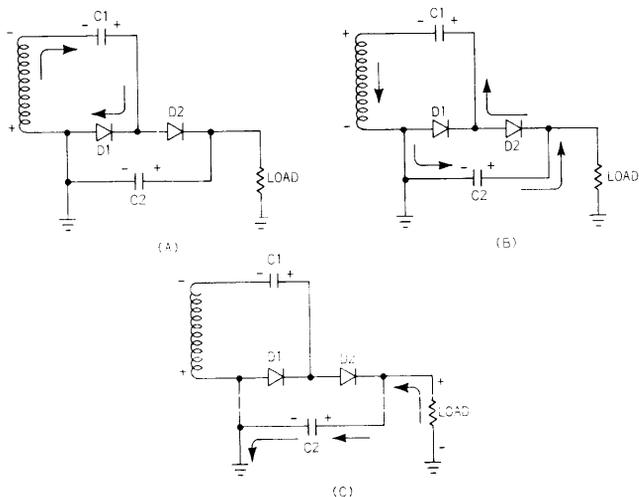


Fig. 5-2. Cascade doubler.

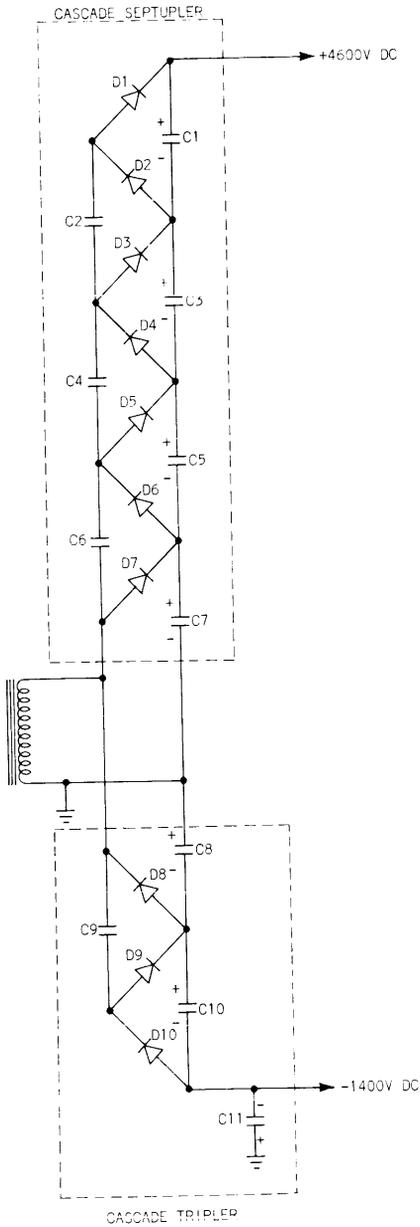


Fig. 5-4. High-voltage power supply.

high-voltage supply

Fig. 5-4 is a schematic diagram of an actual high-voltage power supply, consisting of a cascade septupler and a cascade tripler. The layout of components, typical in Tektronix circuit diagrams, simplifies analysis of the circuit's operation. Capacitors C1, C3, C5 and C7 charge to a total voltage of +4600 V. The output of the tripler is a negative 1400 V.

The behavior of voltage multipliers depends to a large degree upon the amount of load current they are required to supply. The foregoing explanations of voltage-multiplier operation are actually highly oversimplified and do not take the effects of load current into consideration. In the first place, the individual capacitors do not attain their full charge on single alternations of the AC voltage, but build up slowly over many successive cycles. Also, when load current is drawn from the circuit, the capacitors cannot attain their full charge at all. For instance, in Fig. 5-4, C7 should charge to E volts under no-load conditions; however, when current is drawn from the septupler, C7 might well charge to only 0.7E. C6 would then charge to only 1.4E, rather than 1.7E on the second half-cycle. The total voltage across C5 on the third half-cycle is therefore E + 1.4E or 2.4E, minus 0.7E across E6 for a total of only 1.4E. C5 in turn would charge to only about 1.1E due to load-current drain, and so on down the chain, losing a little of the ideal capacitor charge each time. The end result is that the multiplier's output voltage drops off quite rapidly if a significant load current is drawn for a given value of capacitance. This characteristic does not constitute a significant drawback for CRT applications, however, particularly in the post-deflection-acceleration (PDA) voltage supply, where a current drain on the order of only a few hundred microamps is typical.

voltage regulation

More serious is the problem of filtering out AC ripple in the multiplier output. As demonstrated in the discussion of rectifiers (Chapter 1), several μF of capacitance are required to hold AC ripple down to a percent or two at conventional line frequencies. In a 7 kV septupler each capacitor would charge to about 1 kV. Needless to say, such a situation would be highly dangerous. What is more, a two μF capacitor capable of withstanding 1 kV across itself would be very large and expensive. The only

solution to this problem is to increase the supply frequency so that much smaller capacitors can be used. For a given load current and percentage of ripple, the required capacitance is inversely proportional to the supply frequency; thus with a typical supply frequency of 60 kHz, the required capacitor values are reduced by a factor of one thousand. In a typical 7-to-10-kV multiplier, with a reasonable number of stages, the capacitors will range from 0.02 to 0.03 μF and carry voltages of 2 to 3 kV. Very small and inexpensive disc capacitors can be used in this situation, and their charge is so small that it is relatively harmless.

Line frequencies in the 60-kHz range are obviously unavailable. Therefore, a high-voltage power supply must include some type of high-frequency generator (oscillator) to supply the primary of the high-voltage transformer.

Regulation of high-voltage power supply outputs must be examined in the light of CRT characteristics. Consider first the "gun voltage" or anode-to-cathode voltage of the CRT. This voltage has a direct effect on deflection sensitivity. A change of 10% in the gun voltage will result in a 10% change in sensitivity, and thus upset the calibration of the oscilloscope. Therefore it is mandatory that the anode-to-cathode voltage be regulated.

The post-deflection-acceleration (PDA) voltage is not quite so critical. Ordinary CRT's (those without a mesh) exhibit compression, which is a function of the PDA voltage. Fluctuations in PDA voltage will, therefore, have some effect on deflection sensitivity, but it will be a second-order effect. In CRT's with a mesh or frame-grid system where the geometry is adjusted to give a neutral accelerating field, the deflection sensitivity is almost completely independent of the PDA voltage. In some cases it is possible to remove the PDA voltage entirely without affecting the deflection sensitivity to an appreciable extent. Therefore, where the high-voltage power supply is concerned, it is most important to regulate the cathode-to-anode supply, while regulation of the PDA voltage is considerably less important.

The power supply for a typical grid-blanked CRT usually consists of one positive and two negative supplies. The cathode and floating-grid voltages are both negative and are usually derived from ordinary half-wave rectifiers. The accelerating-voltage is positive, and is derived from a voltage doubler, tripler or higher-order multiplier. The cathode voltage must be directly regulated. The grid voltage must also remain constant, (except for unblanking pulses and changes in the intensity control) with respect to the cathode voltage. Regulation of the PDA voltage, as shown above, is of only minor importance.

Theoretically, there is no reason why the tried-and-trusted series-regulator techniques could not be applied to high-voltage power supplies. Practically however, such regulation would be wasteful and expensive in this application. A 1.5-kV power supply, to regulate over a source-voltage range of $\pm 10\%$ would require several hundred volts to be dropped across the series regulator. Considerable power would thus be dissipated in the series regulator tube -- power generated at considerable expense in an oscillator. Large tubes or expensive transistors would be required.

Satisfactory regulation of high-voltage power supplies is achieved in most Tektronix instruments by controlling the amplitude of the high-frequency oscillator output. The output DC level of one of the rectifiers on the secondary side of the transformer is compared with a voltage reference. In this manner an error signal is generated when the rectifier output changes, and is used to control the amplitude of oscillation in the high-frequency oscillator. This arrangement constitutes a feedback regulation system by which the power supply can be quite tightly regulated. The disadvantage of the system, obviously, is that only one of the secondary supplies can be directly regulated, since any change in the oscillator output amplitude will affect all the other rectifiers as well. The output of these other rectifiers will only remain constant as long as their load currents are constant. It is important to remember this point as the discussion proceeds.

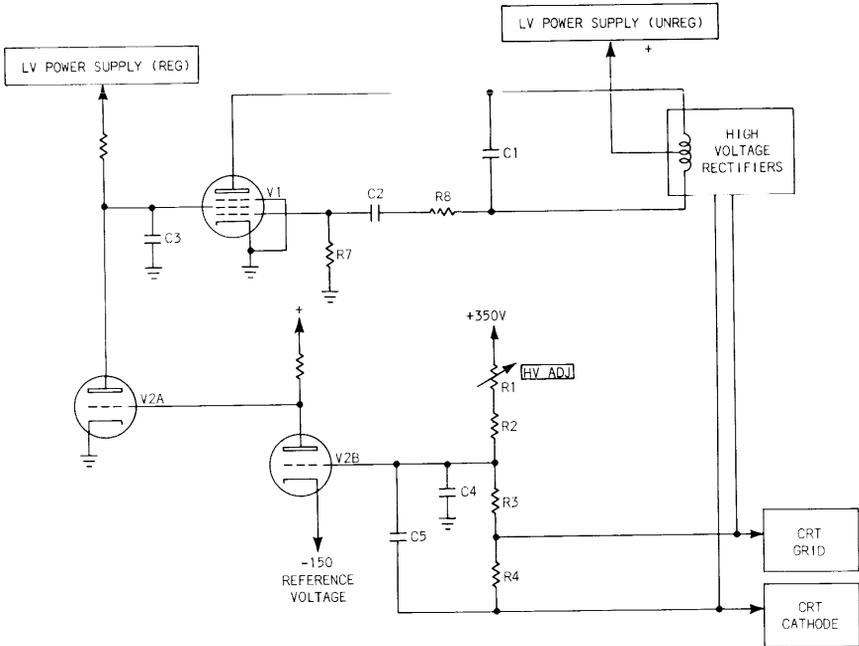


Fig. 5-5 High-voltage oscillator circuit.

The primary side of a typical high-voltage power supply is shown in Fig. 5-5. A long voltage divider consisting of HV-ADJ potentiometer R1, and resistors R2, R3, and R4 is connected between the cathode-supply output and the most positive regulated voltage available -- in this case +350 volts. The CRT-cathode supply is chosen as the source of the feedback signal because, in a grid-blanked CRT, cathode current varies from zero to maximum as the unblanking pulse turns on and off. The output of the cathode supply must therefore be regulated to assure a constant cathode voltage.

The cathode of triode V2B is connected to a regulated minus 150 volts, and its grid is connected to a point on the divider which is also minus 150 volts. Because the positive voltage at one end of the divider is regulated, only changes in the cathode-supply output cause a change in V2B's grid voltage. This voltage is compared with the regulated minus 150 volts at V2B's cathode, generating an error signal on the plate. The plate-load resistor and positive plate supply are so chosen that the quiescent plate voltage is approximately -1 volt.

The plate of V2B is applied to the grid of V2A, the other half of the twin triode, whose cathode is tied to ground. Plate current from V2A flows through the screen-supply resistor of pentode V1, the oscillator tube. Since the amplitude of V1's oscillation depends in part on the potential on its screen grid, a means of amplitude control is established by this arrangement.

high-
voltage
oscillator

Operation of the circuit is as follows: The oscillator, a modified Hartley configuration, utilizes the transformer primary winding and capacitor C1 (together with the stray capacitance in the transformer) to form an LC (tank) circuit between the plate and grid of pentode V1. As plate voltage is applied, oscillations appear in the tank circuit and are coupled to the grid in the proper phase to maintain oscillation. The waveform at the plate is a fairly "clean" sinewave with a frequency of approximately 25-50 kHz. No tuning adjustment is built into the oscillator, since small frequency variations have little or no effect on its output power.

Capacitor C3 filters out the ripple voltage occurring at the oscillator frequency (which would cancel oscillations if allowed to reach the screen grid as a degenerative feedback). Capacitor C2 and resistor R7 provide RC coupling from the tank circuit to the grid of V1.

R8 limits grid current during the positive excursions of the tank circuit's oscillations. Current-limiting prevents the buildup of an excessive negative potential at the grid of V1 due to the RC time constant of R7 and C2 and thus prevents intermittent cutoff of the oscillator.

Assume a decrease in the negative output voltage of the cathode power supply due to component deterioration, line-voltage fluctuations, or load current increase. The grid of V2B, tied to a junction in the voltage divider, will move slightly positive and V2B will increase conduction. The negative signal on its plate will be coupled to the grid of V2A, amplified again and applied to the screen grid of V1 as a positive signal. Conduction in V1 will increase. The oscillator's amplitude will therefore increase, and the rectified output of the cathode power supply will increase to its original value.

The plate supply of V1 does not have to be a regulated voltage, even though changes in this voltage would ordinarily tend to change the amplitude of the oscillator output. These changes would be reflected in the cathode-supply DC level, however, and would set the feedback action in motion to correct the change. The oscillator tube must generate upwards of 5 watts of power, so the tube must be moderately large with a cathode current of 50 to 60 mA and a plate-to-cathode voltage of 450 to 500 volts.

The complete high-voltage power supply is shown in Fig. 5-6. One problem inherent in this configuration is immediately obvious. Diode V1, the cathode-supply half-wave rectifier, provides all the cathode current for the CRT -- about 2 mA. On the other hand, the current that the grid-supply rectifier must provide is that which flows in the voltage divider R7, R9, R10 and R5. This amounts to only a few hundred microamps. The tripler also supplies only a few hundred microamps to the PDA terminal; so, of the five vacuum-tube diodes in the rectifier circuits,

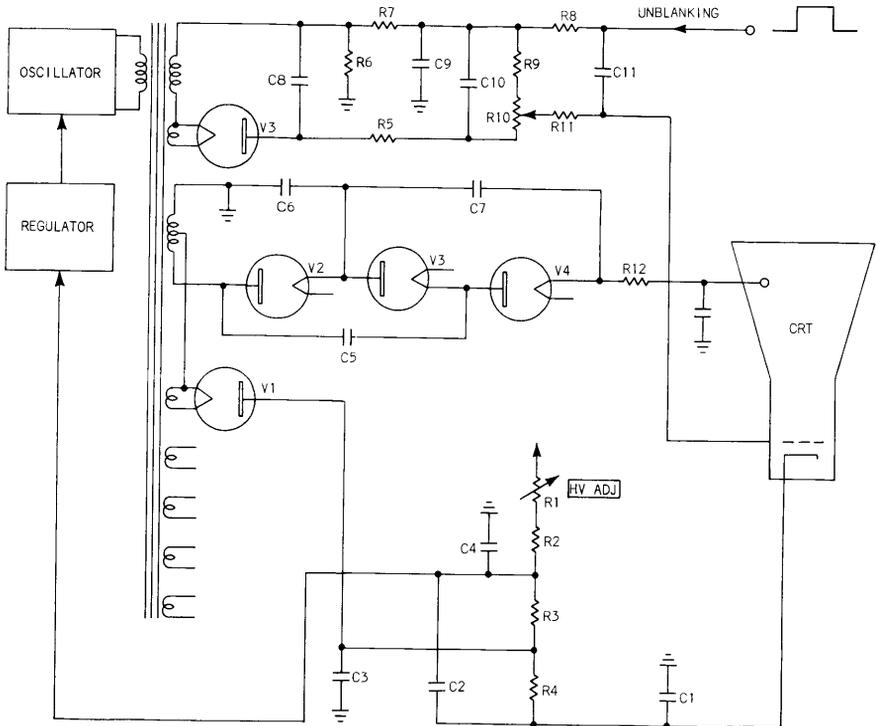


Fig. 5-6. High-voltage rectifiers.

(assuming that they are similar), V1 will lose its emission first. When this happens, the DC output of the rectifier falls off and the regulating circuits cause the oscillator to increase the amplitude of its oscillations. This corrects the cathode-supply voltage, but it also increases the voltage on the secondary windings supplying the other rectifiers. Since the other diodes do not age as quickly as V1, the output of the other rectifiers increases steadily as V1 ages. This causes the grid of the CRT to be driven more and more negative with respect to the cathode. Eventually a point is reached where the intensity control can be driven into the stops in the clockwise (positive) direction without turning on the CRT beam. This symptom is often mistakenly ascribed to failure of the CRT itself. The true situation can be quickly determined however, by replacing V1 with a new tube or switching it with one of the diodes in the other rectifiers.

Another problem is also inherent in this type of high-voltage supply. Although it rides at a negative-kilovolt level, the CRT grid must respond to the *positive* unblanking pulse from the pulse amplifier. Furthermore, this pulse must be direct-coupled to the grid to preserve the DC component of the pulse. Since it is obviously impractical to reference the sweep generator's unblanking circuit to a negative kilovolt level, the entire CRT-grid power-supply must float free of ground and follow the unblanking pulse. Unfortunately, the bifilar high-voltage transformer has a large stray capacitance which tends to degrade the unblanking-pulse risetime. This would delay the appearance of the trace on the CRT at the beginning of the horizontal sweep, especially at higher sweep speeds. A pulse correction circuit is therefore included in the grid-supply rectifier. This circuit is shown in its equivalent form in Fig. 5-7. Capacitor C1 represents the transformer stray capacitance. Resistors R1 and R2 isolate the input and grid from the loading effect of C1, while capacitor C2 couples the fast-risetime leading edge of the unblanking pulse directly to the grid.

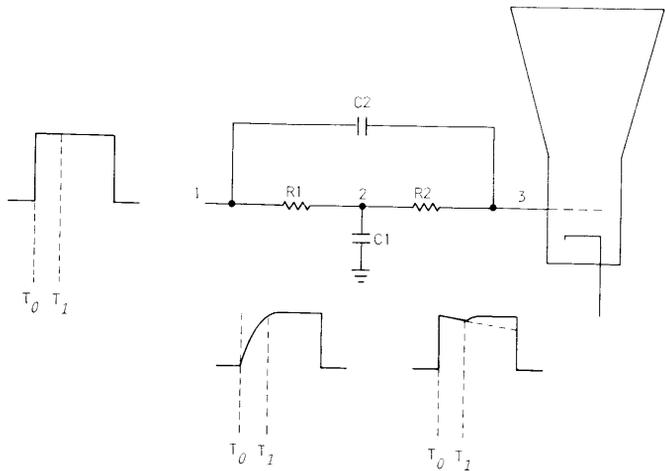


Fig. 5-7. CRT-grid-supply unblanking-pulse correction circuit.

unblanking
pulse

The unblanking pulse is shown as it appears at the input (Point 1). At time T_0 capacitor C1 starts to charge to the pulse voltage (Point 2). C2 meanwhile couples the fast rise to the grid of the CRT and begins to discharge slowly toward the voltage at Point 2. At time T_1 the potential at Points 2 and 3 are equal and the grid rises as C2 reaches full charge. The resulting waveform (Point 3) has a slight dip after the start of the pulse but is essentially the same as the input pulse.

In the actual circuit (Fig. 5-6), C11 acts as the coupling capacitor while resistors R7, R8, and R11 provide the necessary isolation between transformer capacitance, CRT grid, and the unblanking-pulse input. C9 and R6 provide RF filtering for the HV-oscillator frequency, preventing the RF from being coupled into the horizontal sweep generator through the unblanking circuit. C8, C10, and R5 constitute an RC filter for the rectifier output. R2, R3, R4, and HV-adjust potentiometer R1 form a voltage divider across the cathode-supply rectifier output. The adjustment of R10 determines beam intensity for the CRT.

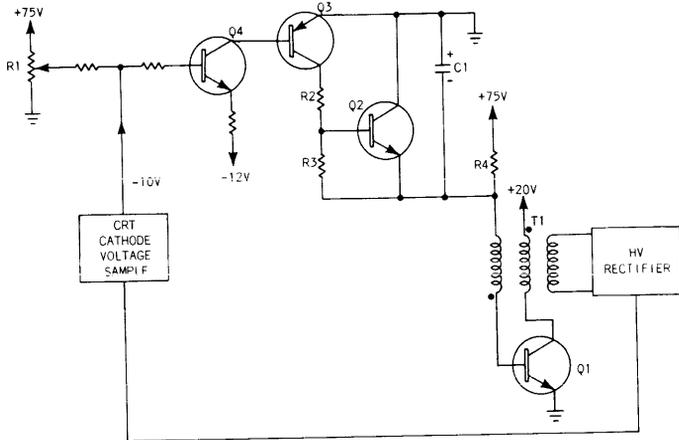


Fig. 5-8. Transistorized HV oscillator.

transistor-
ized
high-voltage
supplies

Transistorized high-voltage supplies for grid-blanked CRT's operate on principles identical to those of the vacuum-tube variety. Because the oscillator is a transistor, however, it has no screen grid through which the amplitude of the oscillator output may be controlled. Instead, the error signal is applied to a circuit which acts to increase or decrease the standing bias on the oscillator transistor. This circuit is shown in simplified form in Fig. 5-8.

Q1 is the oscillator transistor. Its collector load is the primary winding of high-voltage transformer T1. When the instrument is turned on, Q1 goes rapidly into conduction, due to the high forward bias on its base. As current builds up in the primary winding it induces a voltage in the feedback winding. The windings are so phased that the feedback voltage at the base of Q1 is positive, satisfying the requirements for amplifier oscillation. Meanwhile, the other side of the feedback winding has gone negative with respect to ground, charging capacitor C1.

As the collector voltage of Q1 approaches maximum, the field in the feedback winding collapses, driving the base negative and cutting off the transistor.

Capacitor C1 then begins to discharge through R4, Q2, and the series leg R2, R3, and Q3. Q2 is the principal discharge path. Its impedance depends on the forward bias developed by discharge current through R2 and R3. This in turn depends on the current drive applied to Q3.

When the voltage on C1 rises above ground, Q1 again goes into conduction and the cycle repeats itself. A sample of the CRT-cathode voltage is applied to the base of error amplifier Q4, which provides the current drive for Q3.

If the cathode voltage tends to decrease, the base of Q4 will receive a positive signal from the voltage divider in the cathode circuit. This increases the forward bias on Q4 and increases the base current drive of Q3. C1's discharge current through R2 and R3 therefore also increases, resulting in an increased forward bias on Q2. The decreased resistance of Q2 allows C1 to discharge more quickly, thereby reducing the average negative bias applied to oscillator Q1. The increased current through the primary winding of T1 induces a higher voltage in the secondary winding. The output of the cathode supply thus increases (goes more negative), cancelling out the original change. Adjustments to potentiometer R1 have the same effect on the circuit as an error signal and thus establish the output voltage at the desired level.

battery-
powered
high-voltage

Battery-powered instruments present special problems in their high-voltage power supplies. These instruments develop their low-voltage potentials in some types of oscillator-driven power supply (see DC-DC regulator). If their high-voltage potentials were developed in conventional fashion, an additional oscillator would be required. It is clear that a considerable savings in space and power consumption can be effected by using the output of the first oscillator to drive the high-voltage as well as the low-voltage transformer. However, this arrangement presents a problem in regulation. If a sample of the cathode voltage were used to develop a feedback error signal for the oscillator, all the low-voltage supplies would be affected by changes in cathode current. This is obviously an impossible arrangement.

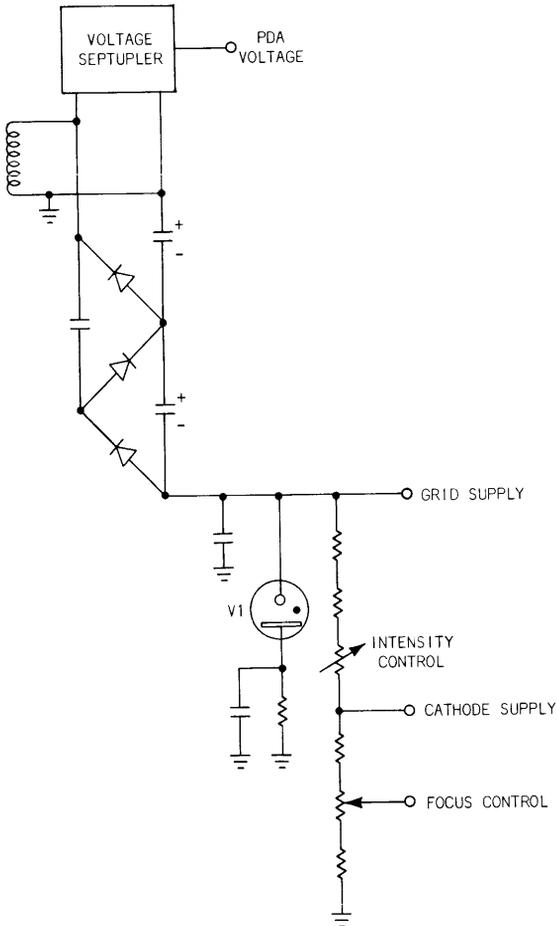


Fig. 5-9. HV supply for deflection-blanked CRT.

The only practical solution is to regulate the secondary voltage independently, Fig. 5-9. As has already been shown, series regulators consume considerable space and power, so a simple shunt regulator is used in the negative high-voltage supply. As a further economy measure, both grid and cathode potentials are taken from the same multiplier circuit -- in this case a tripler. This arrangement dictates a deflection-blanked CRT, since the cathode voltage, which must be slightly positive with respect to the grid, is taken from a point on a long resistive voltage divider. If a grid-blanked tube were used in this application, the changing cathode current, which is generally considerably larger than the divider current, would cause changes in the grid potential as well as in the levels established by the focus and intensity controls. Because deflection-blanked CRT's maintain a constant cathode current, these problems are avoided. The cost of this tradeoff, however, is in the shortened life of the CRT itself.

shock
regulator

corona
stabilizer

The shunt regulator, V1, is not a conventional gas diode, but a "shock regulator," or corona stabilizer. It behaves in the same way as a gas diode, in that it maintains a constant voltage across its terminals over a specified current range. Its operation, however, depends on a corona discharge across two electrodes. The actual voltage rating of the tube depends upon electrode spacing, gas pressure, and other factors beyond the scope of this discussion. In the example shown in the diagram, the tube is rated at a nominal 1400 volts over a current range of approximately 100 to 1000 microamperes.

The PDA voltage is not regulated so the CRT must incorporate mesh or frame-grid construction with a mutual post-acceleration field.

6

POWER SUPPLY HARDWARE

A practical understanding of Tektronix power supplies requires some knowledge of the actual hardware used in their construction. Much of this knowledge can be obtained from a study of standard works on vacuum tubes, solid-state devices, etc. However, certain devices, for one reason or another, merit special attention in a publication of this kind. Foremost among these is the power-supply transformer.

transformers In order to satisfy the demanding criteria established by its designers and engineers, Tektronix has found it necessary to manufacture its own transformers, including those used in power supplies. Although the basic principles of transformer operation are probably familiar to most readers, a brief review may serve to focus attention on some of the practical considerations which govern transformer design.

The power transformer is probably the most efficient electrical machine yet devised. It consists essentially of a magnetic core structure wound with wire or some other conductor, electrical insulation, and suitable mechanical mounting. The alternating magnetic field set up in the primary windings, concentrated by the magnetic core material, induces current flow in the secondary windings which supply the load. In an ideal transformer, all the power supplied to the primary winding is coupled to the secondary winding(s). The current and voltage relationships depend upon the number of turns in the respective windings. Using the subscripts "p" and "s" to indicate "primary" and "secondary," together with the conventional symbols for voltage and current, the following proportionalities can be stated for the ideal transformer:

$$\frac{E_p}{E_s} \approx \frac{I_s}{I_p} \approx \frac{N_p}{N_s}$$

where N = number of turns in windings

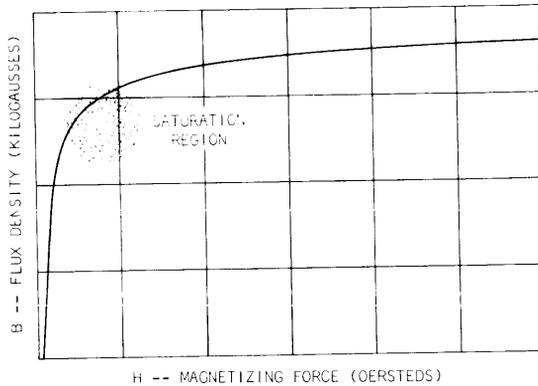


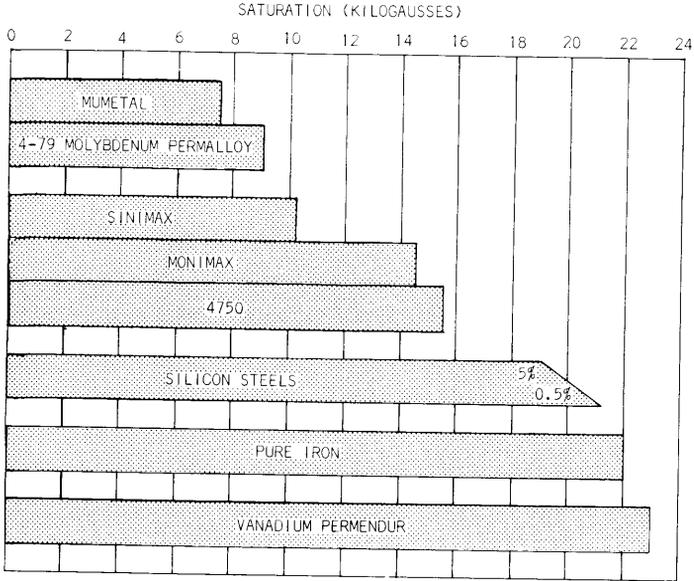
Fig. 6-1. Typical magnetization curve.

flux density
magnetic field
permeability
saturation

The magnetic field set up by a current-carrying coil is measured in terms of imaginary "lines of force" per unit area. The unit of *flux density* is the gauss, and is equal to 6.45 lines of force per square inch. Flux density B , in a given system, depends on the number of turns in the primary coil and the amplitude of current through the coil. The *force* H which creates the magnetic field is measured in oersteds. One oersted is equal to 2.015 ampere-turns. The ease with which a given material concentrates magnetic lines of force (increases flux density) is called *permeability*, represented by the symbol μ . This quantity is found by dividing flux density (B) by the magnetizing force (H) which produces it. A graph of B versus H for a typical core material is shown in Fig. 6-1. Note the sharp "knee" of the curve. Beyond this region, further increases in current produce no *useful* increase in flux density. The material has thus reached *saturation*. Every magnetic material has its own characteristic saturation point as shown in Fig. 6-2.

maximum permeability

When permeability is plotted against flux density, another important characteristic, maximum permeability, is displayed (Fig. 6-3). This characteristic can be controlled to a certain extent through heat treatment and other metallurgical processes. It is obviously desirable that the core material exhibit maximum permeability at the transformer's operating flux density.



THE PROPRIETARY MATERIALS SHOWN ARE MANUFACTURED BY ALLEGHENY-LUDLUM STEEL CORPORATION.

Fig. 6-2. Saturation points for magnetic-core materials.

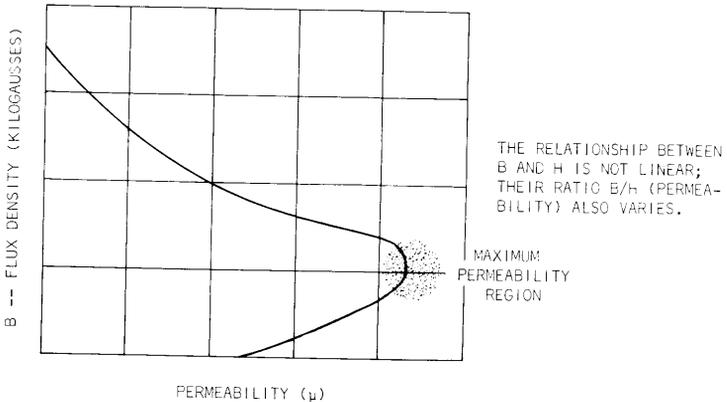


Fig. 6-3. Typical permeability curve.

- coupling losses
- The power losses which occur in a transformer are due to a number of unavoidable, but reducible, effects. The first of these arises from the fact that part of the magnetic flux generated in the primary coil is lost in the coupling process. Although this type of loss can be reduced to a minimum by the proper choice of core material, transformer geometry, etc., some of the primary flux inevitably remains uncoupled and therefore constitutes a small power loss.
- copper losses
- skin effect
- proximity effect
- More important are the losses in the form of heat generated in the transformer. The resistance of the windings gives rise to what are known as "copper losses." These losses increase with the frequency of the applied voltage, due to "skin effect." This is the tendency of high-frequency currents to travel at or near the surface of a conductor, thereby reducing the effective cross-sectional area of the conductor. A similar phenomenon, known as "proximity effect," is caused by distortion of the current-carrying area in one winding by the magnetic field of the windings closely adjacent. This effect also increases the resistance of the windings to a small degree.
- eddy currents
- A third type of heat loss is caused by small currents called "eddy currents" which are induced in the core material itself by the alternating magnetic field (Fig. 6-4). The amplitude of these currents depends on the frequency and flux density of the field as well as upon the specific resistance and thickness of the core material. The resistance offered to eddy currents is highly increased by constructing the core from thin insulated laminations. This technique has no adverse effect on the core's ability to concentrate the flux field. The specific resistance of the core material is controlled to a certain extent by various metallurgical processes.
- hysteresis
- Finally, there is the loss due to a phenomenon called "hysteresis." As the magnetic field shifts back and forth in polarity, the molecules of core material align themselves first in one direction and then in the opposite. The resistance of the individual molecules to these constant changes in alignment varies with the type of core material and determines the degree of hysteresis loss.

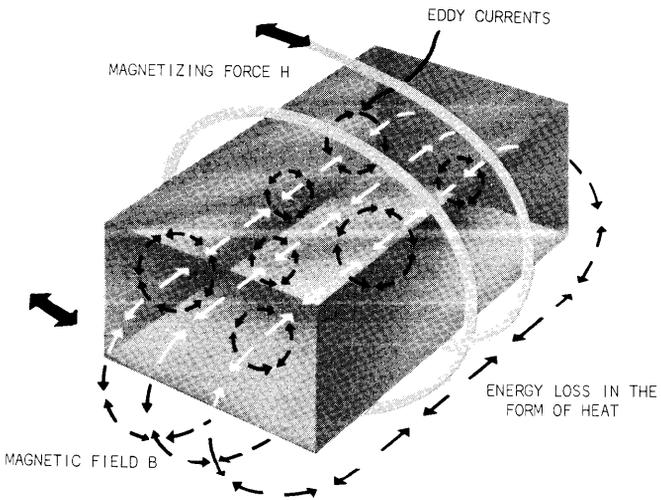


Fig. 6-4. Eddy currents.

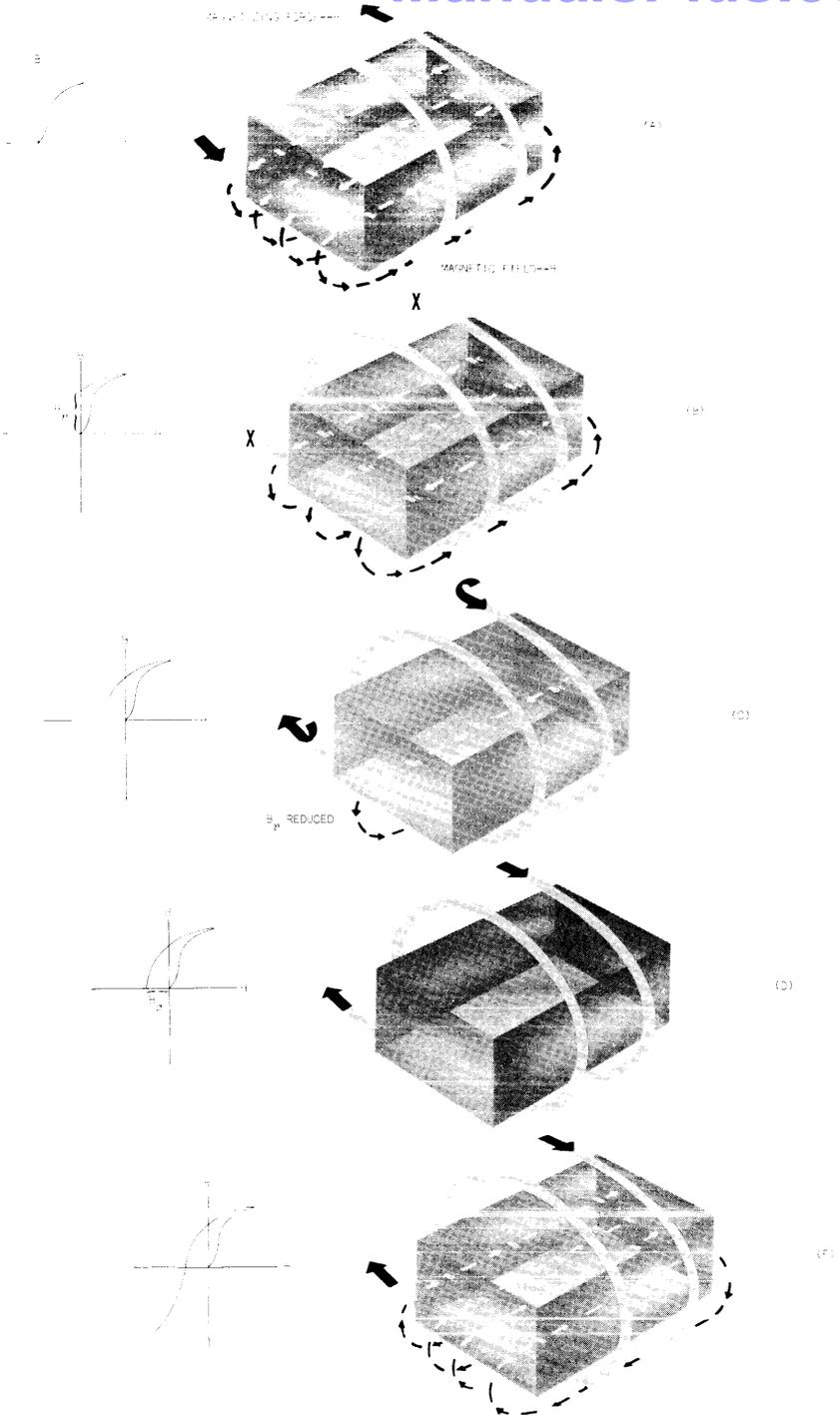


Fig. 6-5. The hysteresis cycle.

The *hysteresis cycle* begins when the core is first magnetized by the current flow in one direction (Fig. 6-5A). When current, and thus, the magnetizing force H , is reduced to zero, some of the molecules in the core remain in or near their aligned positions. The flux density therefore does not return all the way to zero (Fig. 6-5B). The residual flux is designated B_r .

As the magnetizing force is applied in the opposite direction by a reversal of the current (Fig. 6-5C), a "demagnetizing" force starts to overcome the residual magnetism. Eventually a point is reached where the flux field is reduced to zero. The total force required to reach this point is called the coercive force H_c (Fig. 6-5D). As the magnetizing force is further increased to maximum value, the material again attains its maximum flux density, but in the opposite direction (Fig. 6-5E).

hysteresis
loop

Completing the hysteresis cycle produces a closed "hysteresis loop" (Fig. 6-6). The enclosed area is a measure of the energy lost in the core material during that cycle. In alternating-current applications the total hysteresis power loss will depend on the frequency. A typical hysteresis loop for a good magnetic core material will exhibit a large B_r -to- H_c ratio; that is, only a small coercive force is required to eliminate a large residual magnetism.

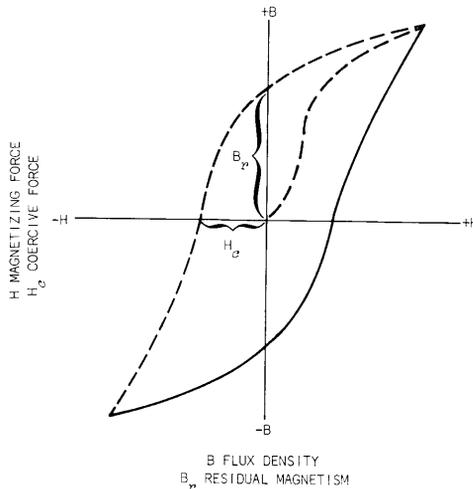


Fig. 6-6. Typical hysteresis loop.

Radical improvements have been made in core efficiencies in the more than 80 years of design and manufacture of electrical equipment. In the early days "hoop" iron of uncertain chemistry and haphazard rolling techniques was used for core structures. The industry has progressed to steels with carefully controlled analyses; much attention is paid to the improved electrical characteristics which can be obtained by proper rolling procedures.

core
materials

The silicon steels are the most widely used core materials. Produced by several steel manufacturers under a variety of trade names, these range from about 0.5% to 5.0% silicon. Other materials for special applications at Tektronix are high-purity irons, iron/nickel alloys, iron/nickel alloys with silicon, molybdenum, copper and chrome, iron/cobalt alloys (with or without vanadium) and "ferrites," or ceramic-processed-alloy cores.

The addition of silicon to steel increases its electrical resistance, thus reducing eddy-current losses. It also minimizes the effects of "aging," a phenomenon resulting in increased energy-loss after the steel has been in service for some time. But silicon steel must be properly heat-treated to achieve the metallurgical stability necessary to prevent aging. On the debit side, silicon lowers magnetic saturation, and in the higher percentages, it causes embrittlement.

The arrangement of the crystalline structure, or grain orientation, is a significant factor affecting the magnetic characteristics of steels (Fig. 6-7). When measured at right angles to the rolling direction, many of the desirable properties are from 10 to 15 per cent poorer compared to the same properties measured in the direction of rolling. Oriented-grain core materials are punched or sheared into required shapes along the direction of rolling so that the longest portion of the magnetic path is in the best magnetic direction.

Fig. 6-8 shows the wide differences in permeability of a single crystal of iron when measured along different axes of the crystal.

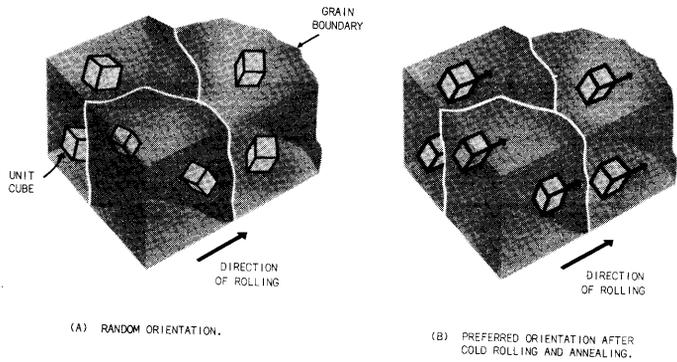


Fig. 6-7. Orientation in grains of iron.

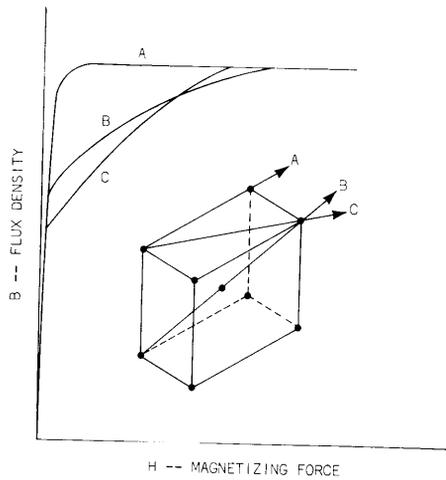


Fig. 6-8. Permeability along different directions of a single crystal of iron.

At normal flux densities, core losses in *oriented* silicon steel are very low. The limitation on the core, therefore, is its flux-handling capacity (regular and external field), rather than loss; the overall permeability of core structure is the important consideration.

transformer
geometry
consider-
ations

The physical size and shape, or geometry, of a transformer is governed by factors too numerous to consider fully in this brief discussion. A few of the more important considerations, however, are listed below:

Volt-Ampere Load: This factor largely determines the overall size of the transformer. To double the volt-ampere load it is necessary to double the wire size and the size of the core.

Type of Load Resistance: Inductive loads versus capacitive loads.

Type of Rectifier: Full-wave, bridge, or center-tap rectifiers.

Line Frequency: Although in general higher line frequencies permit a reduction in transformer size, a number of active variables must be considered before a particular geometry is settled upon. As frequency increases, flux density decreases; so losses due to hysteresis and eddy currents are reduced. However, copper losses, due to skin effect and proximity effect, increase with frequency. Leakage inductance (an effect produced by coupling losses) is a relative constant in a given transformer and increasingly opposes current at higher frequencies.

Allowable Operating Temperature: In general, heat radiation must be kept low to protect components in the instrument. Heat radiation is improved by increasing the size of a transformer. The operating temperature of a transformer equals its "temperature rise" (internally generated heat) plus the environmental (ambient) temperatures.

Flux Density: Higher flux densities require more rigid control of lamination manufacture and assembly.

Copper-Space Factor (Current Density): Greater flux density can be achieved by crowding as many windings as possible into the "window" of the transformer. However, this demands smaller diameter wire or lighter insulation and increases copper losses or reduces the voltage-breakdown rating.

Voltage-Breakdown Rating: Transformers subject to high-voltage transients require heavier insulation. This reduces the number of windings or increases size of transformer.

Regulation: Ratio of the output voltages under load/no-load conditions.

Efficiency: Ratio of output power to input power. Especially important in battery-powered instruments.

Type of Winding Connections: Isolation between source and load is provided by separate primary and secondary windings. The autotransformer (one in which the primary and secondary share a common winding), however, achieves the same power transfer at half the power rating.

DC Polarization: The cores of transformers supplying half-wave rectifiers become polarized by the unidirectional current in the secondary windings. To minimize this effect, larger cores are required.

transformer-
design
character-
istics

The foregoing discussion should illustrate the complexity of the transformer-design problem. In arriving at a final design all the conflicting requirements must be balanced against each other so that the transformer exhibits:

1. *Low External Field:* The alternating magnetic field surrounding a power-supply transformer must be held to a minimum to prevent induced AC modulation in adjacent circuits and consequent "jitter" in the CRT display.
2. *Low Temperature Rise:* Power-supply transformers must not generate excessive heat in the instrument. High temperatures shorten the life of electronic components and change their electrical characteristics. Semiconductors, in particular, require a constant-temperature operating environment for stable performance.
3. *Low Regulation:* The term "regulation," when applied to transformer characteristics, indicates the difference in voltage on the

secondary between "load" and "no-load" conditions. A good power supply transformer provides a stable secondary voltage under a wide range of load conditions and therefore exhibits the characteristic of *low* regulation.

4. *Compact Design:* Cubic dimensions and weight of a power supply transformer must be held to a minimum, especially in portable instruments. Even the *shape* of the transformer must be suitable to its location in the instrument.

line-
voltage
selection

Because the instrument containing a power supply may be used almost any place in the world, measures must be taken to make the power-supply transformer adaptable to a wide range of line-voltage conditions. A choice of two nominal line voltages is provided by constructing the primary winding in two sections (Fig. 6-9). These sections are connected in series for a nominal 115-V source, while for 230 V the windings are connected in parallel. The voltage on the secondary windings is the same in either case.

Line voltages also vary considerably within a nominal range. That is, a rated line voltage of 115 volts may, when measured, actually read 10% above or below that value. Therefore the primary winding of the power-supply transformer is usually provided with several taps. By connecting the line voltage to the appropriate tap, the secondary voltage can be brought to the required level, Fig. 6-10. For a given voltage on the secondary winding, low line voltages require more primary turns while high input voltages require fewer turns.

Unless otherwise specified, instruments are generally wired for 115 V, and adaptive changes must be made in the field. Several methods are provided for this contingency. In the older instruments the necessary changes are made by unsoldering the factory connections and making the required connections according to instructions in the instrument manual. Care must be exercised when making these changes to assure the proper working voltage (115 V) for the fan motor. In other instruments the necessary changes can be made by shifting square-pin connectors mounted on a power board. Sometimes the nominal voltage selection can be made by a simple snap-switch, and

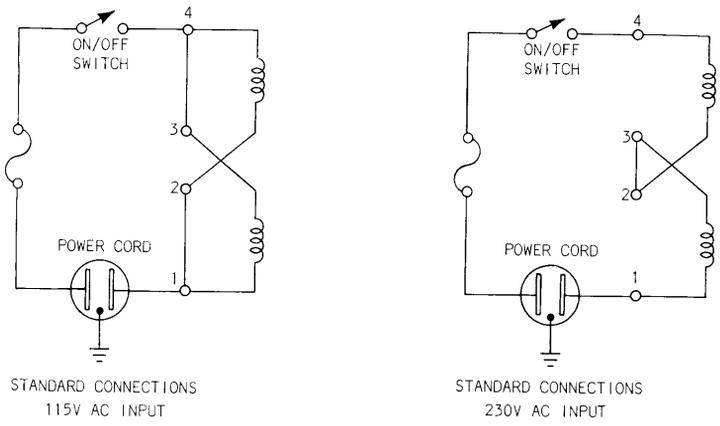


Fig. 6-9. Transformer primary connections.

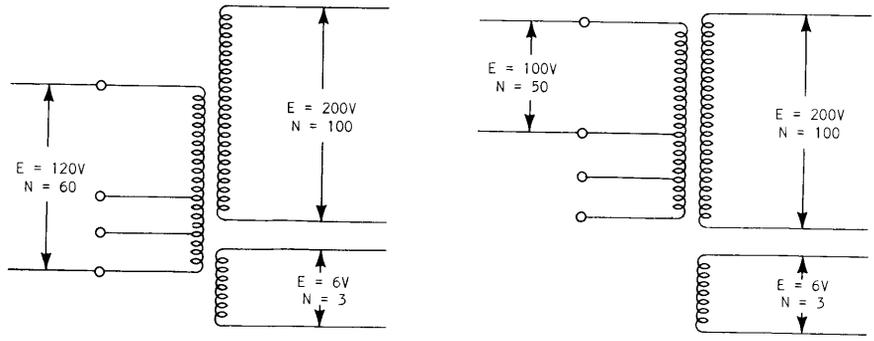


Fig. 6-10. Line-voltage compensation.

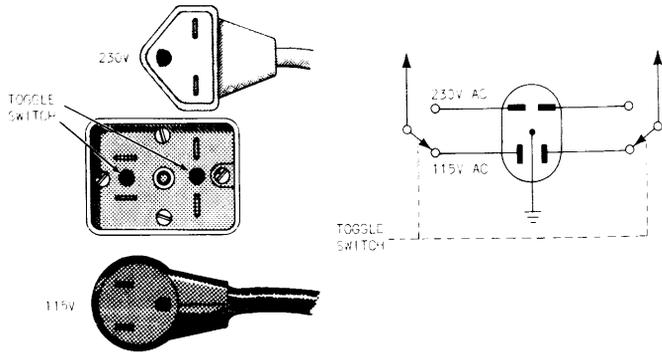


Fig. 6-11. 115 V - 230 V selector.

a number of instruments are fitted with a power receptacle which automatically makes the proper selection through a toggle switch when the appropriate power cord is plugged into it. (Fig. 6-11.)

The voltage selection device shown in Fig. 6-12 provides a rapid and convenient means for adapting an instrument to a wide variety of line-voltage conditions. Input circuit combinations are clearly displayed for visual examination. Appropriate line fuses are automatically selected when adaptive changes are made.

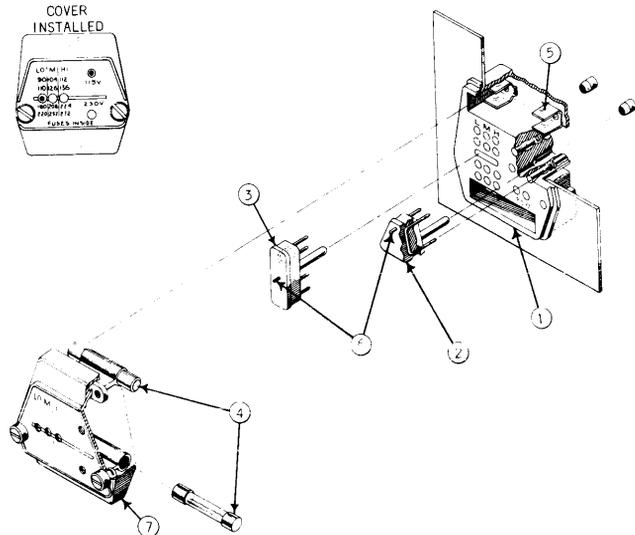


Fig. 6-12. Voltage-selection device.

Power transformer leads are terminated in the selector body (1). Rotating the 115 V-230 V selection plug (2) 180° connects the primary windings in series or parallel configuration. Similar rotation of the range-selection plug (3) places it in the proper voltage range, while movement to left and right selects the transformer taps which accommodate the instrument to high, low, or medium (normal) line conditions. 115-V and 230-V line fuses (4), installed in the selector cover, are snapped into clips (5) in the selector body when the cover is installed. Red stubs (6) on the selection plugs project through apertures in the selector cover (7), indicating the combination of connections selected.

voltage-
regulating
transformers

When line-voltage fluctuations exceed the instrument's regulating capabilities, a *voltage-regulating transformer* may be required. Several types are available commercially; but since typical power supplies are designed for a sinewave input voltage with less than 2% distortion, *saturable-reactor* voltage regulators (or any other type which appreciably affects the shape of the line-voltage waveform) are not acceptable for voltage regulation purposes. This limitation is due to the relationship between peak and RMS values in a sinewave (Fig. 6-13). Since the output voltage of a rectifier depends on the peak input voltage, a distorted sinewave will produce an output which taxes the voltage regulator beyond its specified limits, even though an RMS meter gives the proper indication.

NOTE: Power supplies are generally designed to provide regulation over a wide range of input voltage variations and voltage-regulating transformers are necessary only in exceptional cases.

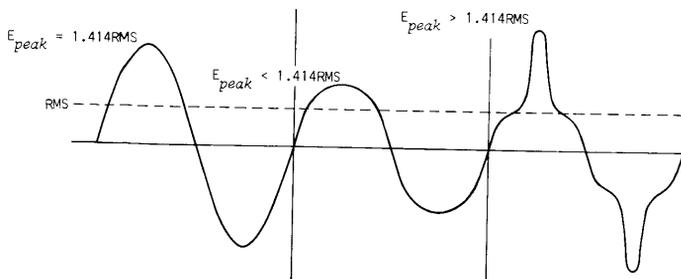


Fig. 6-13. Effects of distortion on E_{peak} .

Power-supply transformers furnish low voltages to filament heaters in tube-type power supplies, usually at 6.3 or 12.6 V AC. Since some circuits include tubes whose cathodes operate at high DC potentials, their filament-supply windings are connected to the output of a high-voltage power supply. This eliminates large potential differences between filament and cathode but also requires additional high-voltage insulation in the secondary filament winding.

filament
windings

DC
filaments

In certain cases filament current is rectified and regulated to prevent filament-to-cathode AC modulation and also to prevent line-voltage fluctuations from influencing cathode emission and thereby the tube's conductance.

line
filters

Silicon diodes generate transient currents as they are switched on and off in the power-supply rectifiers. These pulses contain high-frequency components which must not be allowed to feed back into the line and affect other instruments powered by the same source. Oscillator-driven power supplies also generate high-frequency components which must be isolated from the line. A line filter (i.e.; Radio-Frequency-Interference Filter, Electromagnetic-Interference Filter) is installed in some instruments to provide this isolation.

Fig. 6-14 is a schematic of a typical line filter. T1 is a small toroid transformer. C1 and C2 are small capacitors. High frequencies generated in the instrument are filtered out by the LC filter formed by C1, C2, and the inductance of the transformer windings. Line transients are kept out of the instrument (where they could cause erratic triggering and distortion of the display in very high-gain

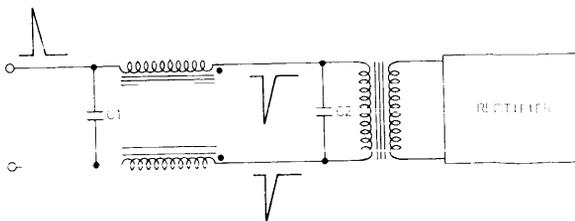


Fig. 6-14. Line (RFI) filter.

instruments) by the action of T1. Signals arriving at the instrument in push-pull (AC line voltage) create opposing flux fields; therefore no coupling takes place and the signal is passed on to the primary unaffected. Single-ended signals, however, are coupled in-phase to the other side of the line. Since current in the primary requires a *difference* in end-to-end potential the net effect in the transformer secondary is zero.

thermal
relays

To avoid damage from excessive internal temperatures, oscilloscopes generally are fitted with some form of *thermal relay*. Action of the relay depends on the difference in expansion coefficients of two different metals bonded together. When heated, a bimetallic strip of this sort will assume a curved shape to accommodate the difference in expansion of the two metals (Fig. 6-15).

A simple thermal relay is pictured in Fig. 6-16. At normal temperatures the contacts are held closed and current passes through them. However, when the bimetallic strip is heated to a given temperature the bimetallic strip bends and separates the contacts, opening the circuit.

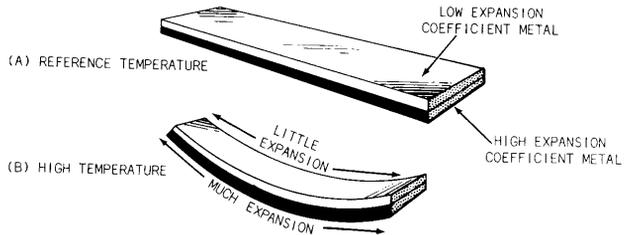


Fig. 6-15. Bimetallic strip.

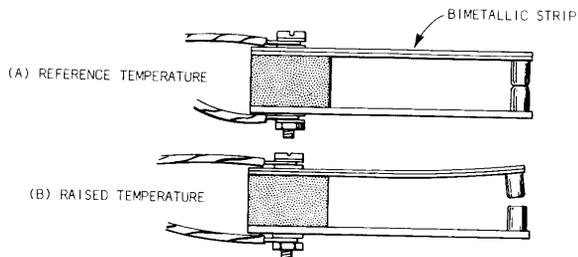


Fig. 6-16. Simple thermal relay.

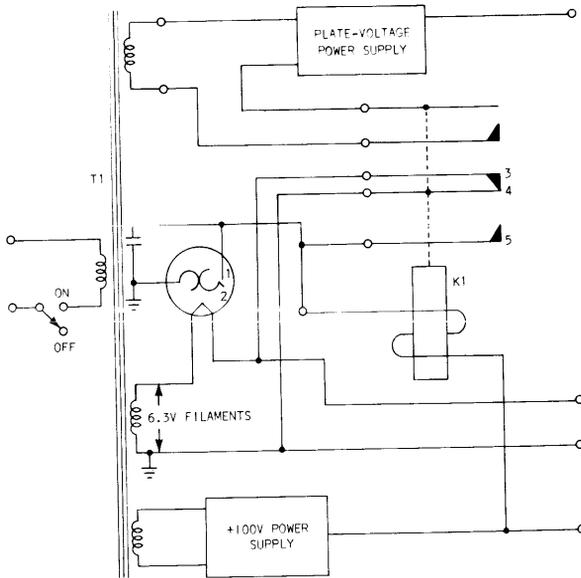


Fig. 6-17. Time-delay relay circuit.

time-delay
relay

When a highly positive plate voltage is applied to a cold vacuum tube some of the electron-emitting coating may be stripped off the cathode, substantially shortening the operating life of the tube. Therefore, most vacuum-tube power supplies are equipped with a relay which delays the application of plate voltages until the heater filaments have had time to warm up the tube's cathodes. Fig. 6-17 is a schematic diagram of a typical time-delay relay.

When power is applied to the primary of power transformer T1 the relay heater is energized by the 6.3-V filament transformer. This same filament voltage starts to heat the cathodes of the power-supply vacuum tubes. Plate voltage is withheld from these tubes by the open contacts of unenergized relay K1.

At a time determined by the setting of the thermal delay relay, the bimetallic strip bends enough to *close* contacts 1 and 2, completing the circuit between ground and K1's energizing supply. When K1 energizes, it closes contacts in the plate-voltage power supplies. The output of these supplies is then routed to the warmed-up tubes in the oscilloscope.

Relay K1 also closes contacts 4 and 5 and opens contacts 3 and 4. This action disconnects the relay heater which is no longer required, since K1's energizing circuit is completed through contacts 4 and 5. When the oscilloscope is turned off, however, contacts 4 and 5 open up and the relay is ready to function as described above.

INDEX

- AC-DC selector, 87-88
- AC-DC supplies, 85-107
- AC ripple, 5-7, 13-14, 17, 34, 52-53, 83, 114
- Battery pack, 87-88
- Binary multivibrator, 93
- Blocking oscillator, 93-107
- Bridge rectifier, 11-16, 109
- Buffer capacitor, 44-45
- Capacitive filters, 3-16
- Cascade voltage multipliers, 110-113
- Corona stabilizer, 125
- Center-tapped rectifier, 15-16
- DC-DC converter, 88-107
- DC filaments, 142
- Diode ratings, 6-7, 10-13
- Diodes, 4-15, 27-30, 33
 - electron tube, 4-15
 - germanium, 4
 - silicon, 4
 - vacuum tube, 4-11
 - zener, 27-30, 33
- Electron tube diodes, 4-15
- EMI filters, 142-143
- Energy storage, 88, 94, 98, 107
- Error amplifier, 39-51, 102
- Feedback rectifier, 102-103
- Filament windings, 142
- Filters, 3-16
 - capacitive, 3-16, 84
 - inductive, 84
 - line, 142-143
 - pi, 84
- Full-wave rectifier, 8-10, 108-110
- Gas diodes, 19-30, 33, 125
- Germanium diodes, 4
- Glow-discharge tube, 19-30, 33
- Half-wave rectifier, 3-8, 83, 115, 118
- High line-voltage protection, 75-76
- High-voltage filters, 113
- High-voltage oscillators, 75, 83, 115-120
- High-voltage rectifiers, 118-120
- High-voltage regulators, 113-120
- High-voltage supplies, 75, 109-125
- Input voltage selection, 138-141
- Line filters, 142-143
- Multivibrator, binary, 93-107
- Oscillators, 75, 83, 88-107, 115-123
 - AC-DC, 88-94
 - blocking, 93-107
 - high voltage, 75, 83, 115-123
- Overload protection, 72-76
- Pi filter, 84
- Plus and minus DC supply, 94
- Primary-regulated low-voltage supply, 83-84
- Pulse amplifier, 103-105
- Pulse-width control, 94, 100-107
- Rectifiers, 3-16, 83, 87, 102-103, 107
 - bridge, 11-16, 109
 - center-tapped, 15-16
 - feedback, 102-103
 - full-wave, 8-10, 108-110
 - half-wave, 3-8, 83, 115, 118
 - high-voltage, 118-120
- Regulators, 3, 17-80
 - high-voltage, 113-120
 - negative output, 54-55
 - series, 18-19, 31-80, 115, 125
 - shunt, 18-32, 125
 - transistorized, 69-80
- Reverse voltage breakdown, 4
- RFI filters, 142-143
- Ripple, 5-7, 13-14, 17, 34, 52-53, 83, 114
- Series regulator, 18-19, 31-80
- Shock regulator, 125

- Short-circuit protection, 72-76
- Shunt regulator, 18-32, 84
- Shunt resistor, 48-50
- Silicon diodes, 3
- Speed-up capacitor, 45, 71-72
- Stacked supplies, 57-61
- Start circuit, 95, 106
- Steering switch, 93-107
- Temperature compensation, 76-80
- Time-delay relays, 144
- Thermal relays, 87, 143
- Transformers, 3, 83-84, 127-142
 - characteristics, 127-138
 - core materials, 134-136
 - design considerations, 136-138
 - filament windings, 142
 - voltage-regulating, 141
- Voltage regulators, 3, 17-80,
113-120, 141
- Vacuum tube diodes, 4-15
- Voltage multipliers, 4, 83,
109-119, 125
- Voltage regulating (VR) tubes,
19-30, 33
- Voltage selection, input,
138-141
- Zener diodes, 27-30, 33, 102

CIRCUIT CONCEPTS
BY INSTRUMENT

Most concepts discussed in this book are applicable to all Tektronix products containing a power supply. However, the concepts listed on the following pages are used in only certain products. Applicability is indicated by a ●.

CIRCUIT CONCEPTS:

INSTRUMENTS:

OSCILLOSCOPES

	310A	317	321A	323	360	422	453	454	491	502A	503
CRT Grid-unblanking Correction (P120-121)								•		•	
Line-voltage Selection, Interlocked (p140-141)							•				
Line-voltage Selection, Plug-in (p139)								•	•		
Line-voltage Selection, Switchable, (p138)							•				
Multiplication, Cascade Doubler (p110)							•		•		
Multiplication, Cascade Multiplier (p110-113)			•	•		•					
Multiplication, Cascade Tripler (p111-113)		•				•		•			
Multiplication, Full-wave Bridge Doubler (p109-110)											•
Oscillator-driven Power Supplies, AC-DC (p85-108)			•	•							
Oscillator-driven Power Supplies, High-voltage (p109-125)	•	•		•	•	•	•	•	•	•	•
Oscillator-driven Power Supplies, Low-voltage (p83-84)											•
Plate-voltage Time Delay (p144)											
Rectification, Center-tapped Bridge (p15-16)											
Rectification, Full-wave Bridge (p109)	•	•		•			•	•	•		
Rectification, Full-wave Center-tapped (p15-16)			•			•	•	•	•	•	
Rectification, Half-wave (p3-8, 83, 115, 118)	•	•	•	•	•	•	•	•	•	•	•
RFI Filtering (p142-143)									•		
Thermal Protection (p87, 143)	•	•	•			•	•	•	•	•	
Voltage Regulation, DC-DC (p88-107)				•		•					
Voltage Regulation, Filament Supply (p142)											
Voltage Regulation, Shunt (p18-32, 125)						•					•
Voltage Regulation, Transistorized (p69-80, 85-107)			•	•		•	•	•	•	•	
Voltage Regulation, Vacuum Tube (p3, 17-69)	•	•			•					•	•

CIRCUIT CONCEPTS:

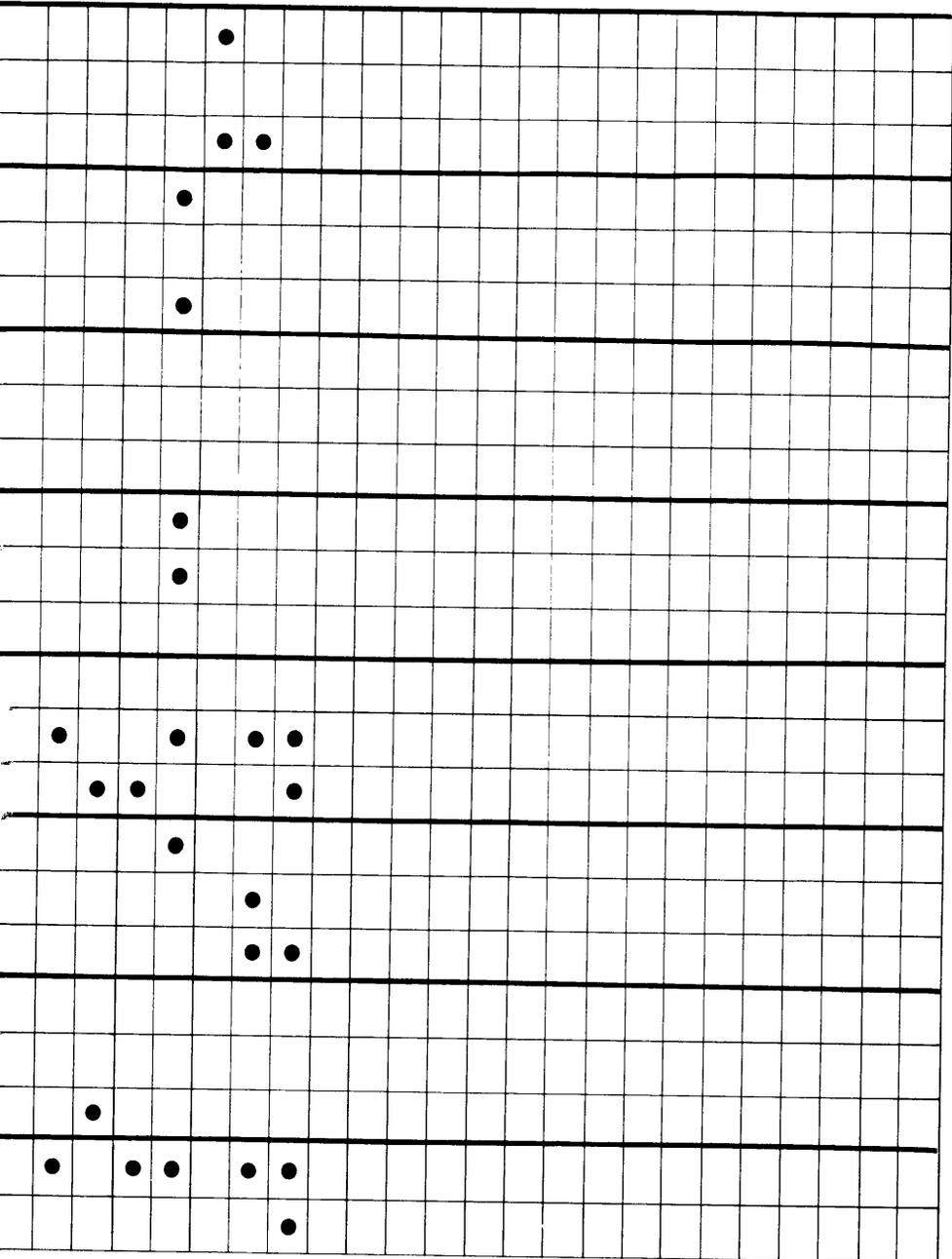
INSTRUMENTS:

OSCILLOSCOPES - cont.

	561A	564	565	567	568	575	581A	585A	647A	661
CRT Grid-unblanking Correction (P120-121)							•	•		
Line-voltage Selection, Interlocked (p140-141)										
Line-voltage Selection, Plug-in (p139)										
Line-voltage Selection, Switchable, (p138)										
Multiplication, Cascade Doubler (p110)										
Multiplication, Cascade Multiplier (p110-113)										
Multiplication, Cascade Tripler (p111-113)							•	•	•	
Multiplication, Full-wave Bridge Doubler (p109-110)										
Oscillator-driven Power Supplies, AC-DC (p85-108)										
Oscillator-driven Power Supplies, High-voltage (p109-125)	•	•	•	•	•	•	•	•	•	•
Oscillator-driven Power Supplies, Low-voltage (p83-84)										
Plate-voltage Time Delay (p144)							•	•		•
Rectification, Center-tapped Bridge (p15-16)			•			•	•	•		
Rectification, Full-wave Bridge (p109)	•	•	•	•	•	•	•	•	•	•
Rectification, Full-wave Center-tapped (p15-16)	•	•	•	•	•	•			•	
Rectification, Half-wave (p3-8, 83, 115, 118)	•	•	•	•	•	•	•	•	•	•
RFI Filtering (p142-143)				•					•	•
Thermal Protection (p87, 143)	•	•	•	•	•	•	•	•	•	•
Voltage Regulation, DC-DC (p88-107)										
Voltage Regulation, Filament Supply (p142)							•	•		
Voltage Regulation, Shunt (p18-32, 125)										
Voltage Regulation, Transistorized (p69-80, 85-107)	•	•	•	•	•		•	•	•	•
Voltage Regulation, Vacuum Tube (p3, 17-69)	•	•	•	•		•	•	•		•

MISCELLANEOUS

81A
130
292
410
601
611
1121



*This book is but one of a series.
The series consists of two groups,
circuits and measurements. These
texts present a conceptual approach
to circuits and measurements which
apply to Tektronix products.
Several "concept books" are in
preparation; those now available
are:*

Power Supply Circuits
062-0888-01

Oscilloscope Cathode-Ray Tubes
062-0852-01

Storage Cathode-Ray Tubes and Circuits
062-0861-01

Television Waveform Processing Circuits
062-0955-00

Information Display Concepts
062-1005-00

Semiconductor Devices
062-1009-00