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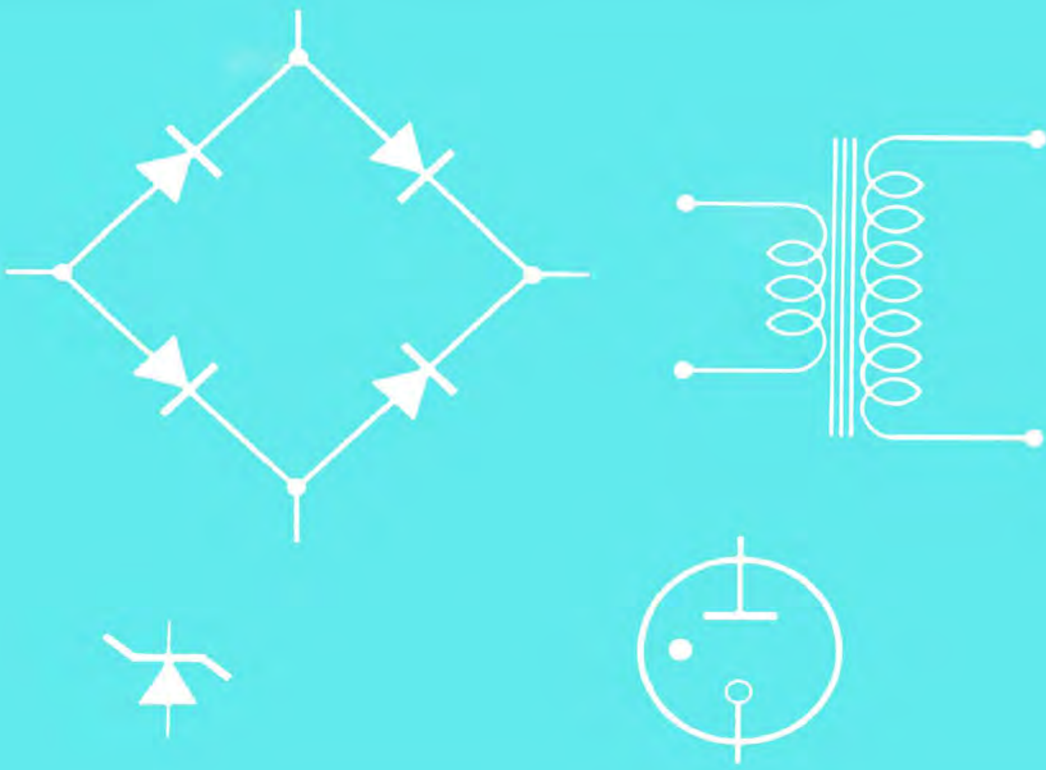

DESIGN & OPERATION OF REGULATED POWER SUPPLIES

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Preface

Present demands make it impractical to use power supplies merely capable of delivering nominal voltages and currents. Rather, the modern trend is toward the use of power supplies that provide a constant output despite variations in input voltage or load characteristics. The regulated power supply accurately establishes stable operating conditions, thus enhancing the performance of the equipment it powers. Unregulated supplies have a number of shortcomings which cannot be tolerated even in simple electronic systems, to say nothing of complex precision equipment. Although more common devices such as television receivers have often dispensed with such refinements, the higher quality performance obtained through their use would generally offset the additional cost of having a regulated power supply included.

Since the trend is moving toward the use of regulated power supplies to improve the operation of various electronic devices and systems, the subject of regulation is of major interest to all technical personnel in electronics. Service technicians will encounter these regulated power supplies in numerous pieces of equipment which they are called on to repair. Engineers instinctively will include them in their basic design in order to comply with rigid performance specifications. Even radio amateurs are no longer satisfied with merely any type of communications. They want the stability that regulated power supplies can give to their equipment which include transmitters, receivers, modulators and VFO's.

Precision measuring instruments and many other quality devices depend on the high degree of stability and increased life span which are obtained through the use of such supplies. Observation of a wide variety of equipments leads to the conclusion that an understanding of regulated power supplies is a necessary supplement to the knowledge of amplifiers, modulators, oscillators, and other functions that are elementary in applied electronics.

It should be understood that the circuits included herein are primarily prototype designs accompanied by nominal component values. These values will provide guidance in the breadboard design of preliminary models. However, in the final design it may become necessary to depart from these values in order to comply with the specific voltage and current demands of the equipment.

IRVING M. GOTTLIEB

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CHAPTER 6

CHAPTER 1

Basic Concepts of Regulated Power Supplies

Obviously, a most necessary prelude to our study of regulated power supplies is an understanding of the term "regulation."

In electrical engineering, "regulation" denotes the *change* in voltage occurring at the output terminals of a power source between no-load and full-load. The so-called voltage-regulated power supplies now common in electronics incorporate automatic means for maintaining the output-terminal voltage constant despite variations in the AC line voltage or DC load current (Fig. 1-1). In a very true sense these sup-

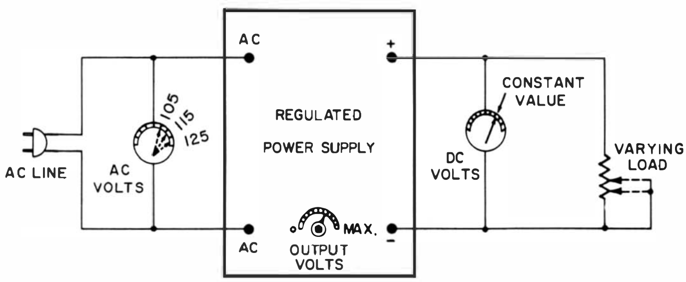


Fig. 1-1. A voltage-regulated power supply can maintain a constant load voltage despite variations in line voltage or load resistance.

plies are designed to eliminate or at least greatly reduce regulation. Therefore, when we employ the commonly accepted parlance, "voltage-regulated power supply," we pay homage to convention, but understand that such a supply is one in which the output voltage is stabilized with respect to changes in line and/or load conditions. The more effective such output-voltage stabilization is, the *smaller* the regulation of the power supply.

Regulation is a dimensionless quantity usually expressed as a percentage:

$$\% \text{ Regulation} = \frac{\text{No-load voltage} - \text{Full-load voltage}}{\text{No load voltage}} \times 100$$

As an example, consider a power supply which develops a no-load voltage of 250 and delivers 225 volts to a rated load. The regulation of this supply is 10%:

$$\frac{250 - 225}{250} \times 100$$

More precisely, this power supply has a regulation of *minus* 10%—the minus term conveying the information that the full-load voltage is less than the no-load, or open-circuit, volt-

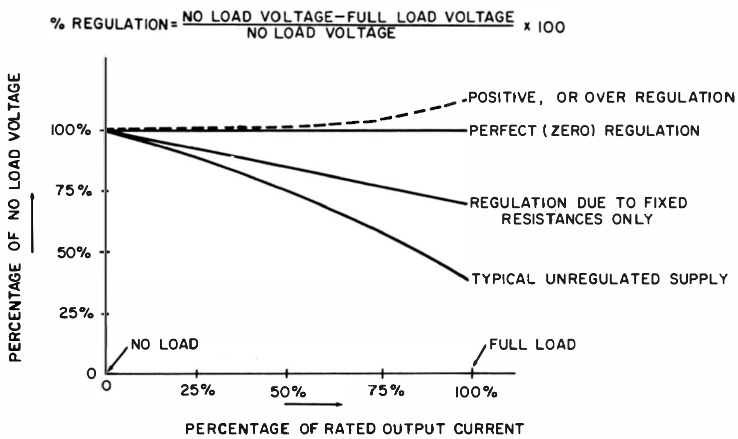


Fig. 1-2. Typical regulation curves.

age. Fig. 1-2 shows some typical regulation curves. You will see later that it is possible to achieve positive regulation with the load voltage rising as the load current is increased. Generally, when the minus or plus designation is not used, the implication is a voltage dropoff with load—that is, regulation in the "minus" direction. Regulation percentages between .001 and 1.0 are readily attainable with appropriate electronic circuits.

The need for stabilization of output voltage arises from the fact that almost any simple power supply has an inherent negative regulation. This is due to the inevitable resistance which acts in series with the load. When the load current flows through this inherent (internal) resistance, the resultant voltage drop subtracts from the voltage delivered to the load. In Fig. 1-3 we see that a 100-volt battery supplies a diminishing voltage to the load as the load is made to consume more current. Here, R_{int} represents the effective resistance, which actually comprises ohmic, electrochemical, and other effects within the battery. The regulation characteristic of such a battery is also typical of generators, rectifiers, transformers, photocells, and other energy converters for

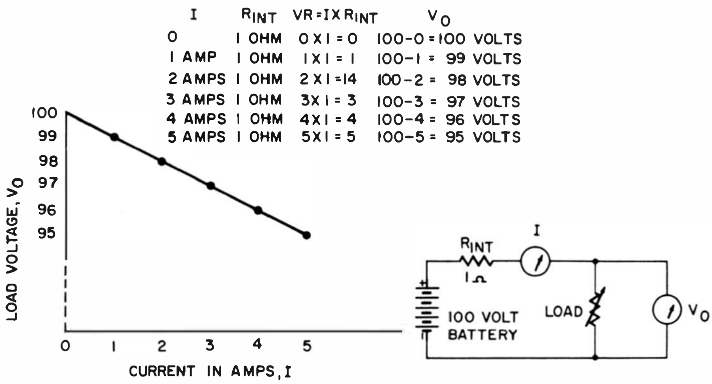


Fig. 1-3. Example of regulation due to resistance in series with load current.

supplying electrical power. Generally, the larger the batteries, rectifying devices, transformers, and other components used in simple supplies are physically, the less the internal resistance. However, any increase in bulk beyond that required from the standpoint of safe thermal operation is extremely costly and, more often than not, impractical as well.

CURRENT-REGULATED POWER SUPPLIES

Most loads encountered in electronics derive maximum benefit from operation with a voltage-stabilized source of power. This is not universally true, however. Sometimes it is of direct importance that the *current* be maintained constant. For example, sometimes a solenoid is used as a means of providing a steady magnetic field to an electronic device. The strength of such a field is determined by three factors—the number of turns of wire, the current passing through these turns, and the magnetic characteristics (permeability) of the medium exposed to the field. In most practical cases, current is the only factor which can suffer appreciable variation. This is particularly true because current is influenced both by the voltage impressed across the solenoid and by the temperature-dependent resistance of the windings. Energizing such a solenoid from a constant voltage supply will not stabilize the magnetic field strength, whereas operation from a stabilized current source will. An example of a current-carrying solenoid from which a constant magnetic field strength is desirable is the focusing solenoid employed in conjunction with certain traveling-wave tubes. When a current-regulated power supply is used, the current remains at a fixed value with respect to wide variations in load resistance and, generally, considerable fluctuations in line voltage. The voltage delivered by such a supply has such extremely poor regulation that its use is prohibited with voltage-sensitive loads.

The concept of percentage regulation is used in a slightly different manner with current-regulated power supplies. We cannot paraphrase the relationship for voltage regulation by stating that current regulation is equal to the difference between no-load and full-load currents divided by no-load current, because no-load current is always *zero* and hence a meaningful quotient is not mathematically obtainable. In order to specify the figure of merit of current regulation, we compute the percentage change in currents between two specified loads. One is generally zero resistance, or a short circuit. The other is then a certain numerical value of resistance. Let I_0 represent the current which flows when the output terminals of the supply are short-circuited, and I_R, the current which flows when resistance R is connected across the output terminals. Then we can designate percentage current regulation as follows:

$$\% \text{ Current Regulation (from zero resistance to R)} = \frac{I_0 - I_R}{I_0} \times 100$$

For example, suppose a current-regulated power supply delivers 5 amperes under a short-circuit condition—that is, when an ammeter is connected across the output terminals. When a 10-ohm load is connected across the output terminals, the current flow is reduced to 4.8 amperes. The percentage of current regulation between a short circuit and a load of 10 ohms is:

$$\frac{5 - 4.8}{5} \times 100 = 4\%$$

OUTPUT RESISTANCE

In *voltage-regulated power supplies*, the output resistance is the ratio of a change in output voltage to a change in output current, with the line voltage maintained constant. In Fig. 1-4, R is varied to change the output current; the resultant change in output voltage is derived from the corresponding voltmeter indications. A voltage-regulated supply with perfect (zero per cent) regulation would have an output resistance of zero ohms because a current change would not be accompanied by a voltage change. The magnitude of the current change should be stated. For example, we might specify the output resistance of a power supply as being 0.1 ohm from zero to full load. This value would then be understood as an average value that is valid within the specified load range.

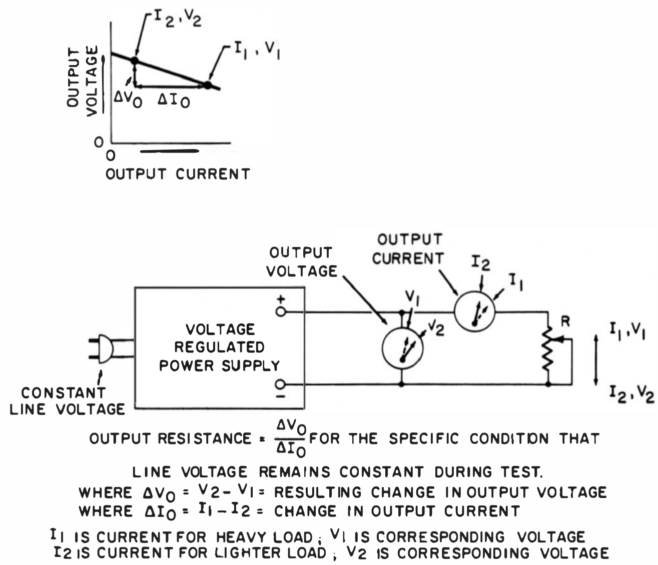


Fig. 1-4. Test setup for illustrating concept of output resistance.

The output resistance of a *current-regulated power supply* is similarly defined and is ideally infinite for a supply with perfect regulation. A difference in the practical determination of output resistance in the two types of supplies is that in a voltage-regulated supply, we change the load in order to change the load current and, in turn, the output voltage. Conversely, in the current-regulated supply, we change the load to change the load voltage and, in turn, the output current. Furthermore, a current-regulated power supply cannot be so tested down to zero current; a minimum load is required for current stabilization to exist. On the other hand, a short circuit of the output—with reduction of the output voltage to zero—is a valid test. For example, we could record a current and voltage of one ampere and zero volts for a current-regulated supply with shorted output terminals. Next, we might observe a current of 0.9 ampere and a voltage of 20 when the load is adjusted for its nominal operating resistance. The output resistance of this supply would then be 20 volts divided by 0.1 ampere, or 200 ohms.

Dynamic Output Impedance

The most common and relevant use of percentage regulation and output resistance is to define the DC output characteristics of a power supply. This, however, is not the whole story. As a matter of fact, a regulated DC power supply is often employed primarily for its AC characteristics. By AC characteristics, we are not at this time concerned with the AC line voltage. Rather, we wish to know how the power supply appears to a small AC voltage impressed across the output terminals. This is a very practical parameter because many, if not most, electronic loads involve circuitry in which AC is superimposed on the steady DC operating current. Amplifiers, oscillators, and modulators are examples of electronic "building blocks" which fall into this category. Switching circuits, multivibrators, and logic systems involve pronounced AC components in their operation. It is not sufficient merely to be concerned with the degree of DC voltage stabilization provided by a power supply when such circuits constitute the load. The power supply is more than an incidental requirement for energizing, or establishing, the operating bias of an electronic circuit. The power supply is actually part of the circuit for both DC and AC components. As an example, consider a two-stage audio amplifier. If the power supply appears as a high impedance to audio frequencies, it will then constitute a common impedance tending to couple the stages together in an undesired way. Such unintentional coupling is actually a feedback path allowing the intermingling of input and output voltages. This can cause oscillation, degradation of frequency response, or loss of gain. Ideally, the power supply for such an amplifier should appear as a short circuit to the whole spectrum of AC frequencies involved. Normally, a very low impedance is obtainable from a voltage-regulated supply, and this consideration may be of greater importance than the DC voltage stability defined by regulation percentage.

Dynamic output impedance can be made low, even in an unregulated supply, by means of output filter capacitors.

Thus, regulation percentage and dynamic output impedance have some relationship to each other, for it is well known that large output capacitors tend to maintain the load voltage at a constant value. Indeed, a capacitor with many millions of farads of storage capacity would be practically indistinguishable from a large battery and would provide excellent regulation for relatively light loads. In practical power supplies, considerable DC voltage stabilization is often provided, for loads of a small fraction of one ampere, by capacitors ranging from several tens to several hundreds of microfarads. The contribution to low dynamic output impedance made by any size of capacitor becomes progressively better as the frequency is increased. Consequently, an inordinately large capacitor may be required in order to provide a low dynamic output impedance at low frequencies. Often this situation pertains to the ripple frequency (generally 120 cps). These considerations make electronic regulation extremely desirable—for not only is excellent (low) DC voltage regulation readily attained, but low AC dynamic output impedance is maintained through the low frequencies, all the way down to zero frequency (DC). At the same time, many designs provide a low dynamic output impedance up to quite high frequencies as well. Often the *combination* of electronic voltage regulation and a relatively small output capacitor permits a low dynamic output impedance from DC to radio frequencies.

REGULATED SUPPLIES AS SEEN BY THE LOAD

A brute-force approach to zero regulation leads to an awkward predicament. If, for example, a negligible voltage drop is desired from a 12-volt battery over the load-current range of 0-to-20 amperes, it will not do merely to choose a battery capable of supplying such a current. The internal resistance of such a battery would be too high to permit the required low regulation; a battery of many tens of times the rating needed to supply 20 amperes would be required. Indeed, if our regulating percentage were specified too low, the battery could well crowd us out of the laboratory! Herein lies the beauty of the voltage-regulated power supply; it need only be designed for the required load. Yet it "looks" like a source of tremendous current capacity to the load (insofar as regulation percentage and dynamic impedance are concerned; see Fig. 1-5).

A similar situation prevails for current-regulated supplies. One having nearly zero current regulation appears to

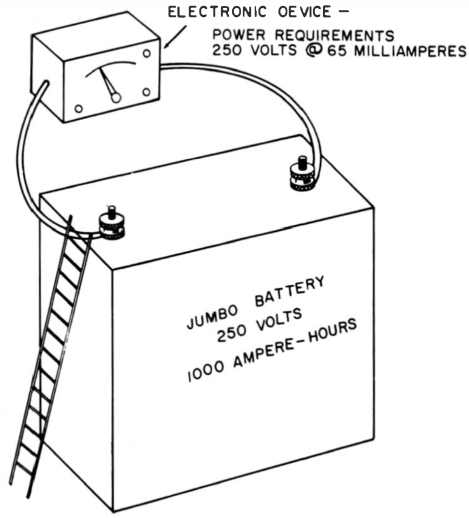


Fig. 1-5. A voltage-regulated power supply appears to the load as a tremendous source of current, although actually it is not.

the load as a power source followed by an extremely high resistance (Fig. 1-6). It is obvious that to actually make such a circuit would lead to gross inefficiency, for the voltage drop and therefore the power dissipation in such a resistance would greatly exceed that in the load itself. The current-regulated power supply "looks" like a source with extremely high internal or series resistance. But this is where the resemblance ends, for the current-regulated supply *does not* incorporate an inordinately high resistance in the load-current path.

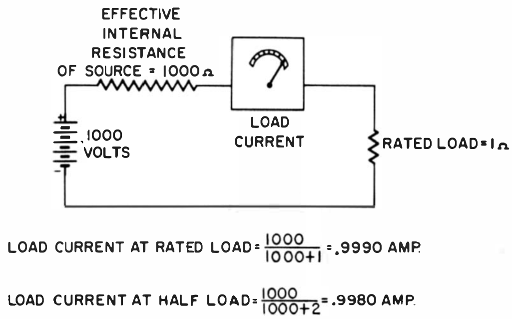


Fig. 1-6. A current-regulated supply acts as a source with a very high series or internal resistance insofar as its current-regulating ability is concerned.

Response Time

If the load to a regulated power supply is abruptly changed, a transitory state will exist before the output becomes stabilized. There are several reasons. Shock excitation of various inductances and capacitances is frequently encountered and takes the form of damped oscillations. Another reason is that energy-storage elements impose time constants which simply do not permit an instantaneous change in level. The inductance of the DC control winding of magnetic amplifiers is an example. Still another limitation in the rapidity of response is the relatively low frequency response of certain transistors. Instantaneous response theoretically implies the capability of infinite frequency response. In practical supplies, amplifiers and control elements must have frequency responses in the megacycle range in order to provide a "rapid" response. When transistors having frequency cutoff characteristics of several kilocycles are used, the response time from this alone is much longer than one could obtain with a vacuum tube (assuming the current-handling capability of the tube was suitable).

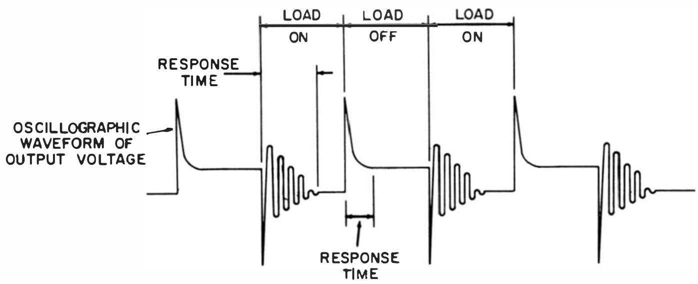


Fig. 1-7. Response time.

Fig. 1-7 provides an insight into the meaning of response time. The frequency at which the load is changed should be slow enough that the final extinction of all transient phenomena is distinctly observed. Response times for adding a load and relieving it will generally be different. The longer of the two is the more important specification because it imposes the frequency limit at which the power supply behaves as a dynamically regulated source of power to the load. It is often stated that an output capacitor can be employed in conjunction with a voltage-regulated supply to extend this frequency (that is, to decrease the response time). This is true in many practical situations, but may prove misleading in others. A large output capacitor does lower the dynamic output impedance, and it can decrease the response time by preventing or attenuating transients. However, the time constant involved in charging and discharging the output capacitor can, if the capacitor is large enough, slow down the response time. A similar line of reasoning applies to response time in a current-regulated power supply. Here, a series inductance can increase the dynamic output impedance, which should be as high as possible. In particular, such an inductance can take over at those frequencies at which the regulating circuitry falls off in response. However, the response time of the supply can be adversely increased by such an inductance, because time is required to change the current level in an inductance.

Response time can be specified for load variations other than from zero to full load. In such instances, the conditions should be stated; for example, one could specify the response time of a voltage-regulated power supply as being, say, 600 microseconds when the load is changed from one-quarter to full load.

Line-Voltage Considerations

Regulation percentage is made more meaningful by relating it to a permissible change in line voltage as well as to different load conditions (see Fig. 1-8). We might specify,

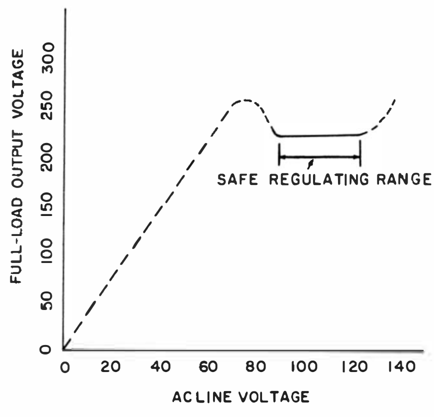


Fig. 1-8. The line-voltage range is an important consideration with regulated power supplies.

for example, that a certain power supply exhibits a maximum regulation of 1% for a ±10% change in the line voltage. The regulation of this supply between no-load and full-load could be more or less than 1%, depending on circuit details. Although there is a tendency for low load regulation to be accompanied by low line regulation, the two modes of stabilization are not rigorously, nor even necessarily, related. The most difficult operating conditions for voltage-regulated supplies usually occur for the combination of low line voltage and full rated load, for it is then that the greatest tendency exists for a drop-off in the output voltage. High immunity to line-voltage variations is desirable, not only to assure a constant output voltage, but also to prevent transients and noise from being transferred from line to load. Some regulating circuits produce large voltage transients when the line switch is opened or closed. This is undesirable—particularly for operation of semiconductor devices, wherein the surge can assume destructive proportions.

Stabilization Ratio—This defines the factor by which a regulated power supply reduces the effect of variations in line voltage, divided by the resultant percentage change in DC output voltage. As an example, suppose the DC output of a 500-volt supply undergoes a total excursion of 2 volts as a result of a +10-volt change in the AC line voltage from a nominal 115 volts. The percentage change in line voltage is 2x10 ÷ 115 = 17.4. The percentage change in DC output voltage is 2 ÷ 500 = 0.4%. The stabilization ratio is then 17.4 ÷ 0.4 = 43.5-to-1. Stabilization ratios in the tens of thousands are attainable from closely regulated power supplies.

RIPPLE

The ripple output of a supply is another definitive operating feature. If, as is usually true, full-wave rectification from a single-phase line precedes the regulating circuitry, the main ripple component will be at twice the line frequency. Sometimes an output ripple voltage is specified simply as a maximum of so many volts rms or peak-to-peak. For instance, we might state that a particular supply developed a maximum of five millivolts rms. This would imply any line voltage, load, or combination of the two within the ratings. On the other hand, we could specify the ripple as being no more than a specified percentage of the DC output voltage. Thus, a supply might be said to have a peak-to-peak ripple voltage of no more than 0.1% of the DC output voltage. If the DC output voltage happened to be 50 volts, then the maximum ripple voltage would be .05 volt peak-to-peak. Peak-to-peak is a better way to define ripple than rms, because the ripple voltage is often nonsinusoidal; in fact, sometimes the ripple waveform is characterized by spikes of fairly high amplitude. In such an instance the spikes, having a low energy content, would not contribute much to the rms value and their presence would be virtually concealed by this measurement method. A power supply rated for ripple voltage no greater than one millivolt might appear suitable for operation of sensitive equipment. However, if the one-millivolt rms rating included 20-millivolt spikes, such a supply would be likely to inject considerable noise into the equipment. A common source of such spikes is a semiconductor rectifier as it switches into and out of conduction. In current-regulated supplies, current ripple—not voltage ripple—is of primary importance. Other than this, the basic ideas are similar.

Most regulated supplies exhibit low ripple, because the automatic regulating circuitry acts on the ripple in the same

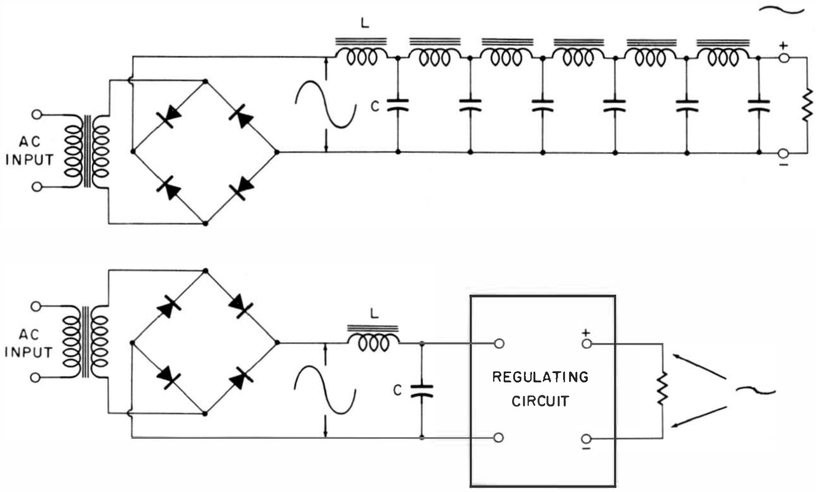


Fig. 1-9. The regulating circuit has a ripple smoothing ability equivalent to a complex network of filter elements.

manner as for any other tendency for the output voltage (or current in current-regulated supplies) to change (Fig. 1-9). In a very real sense, automatic regulation constitutes electronic filtering. This, indeed, is often a justifiable reason for incorporating automatic regulation. A regulated power supply can produce the same ripple suppression as several stages of large, heavy inductor-capacitor sections.

The figure of merit of ripple suppression contributed by a regulator circuit or device is given by the ripple factor. Ripple factor is the ratio, expressed as a percentage, of output ripple voltage to input ripple voltage. That is:

$$\text{Ripple Factor} = \frac{E_{out}}{E_{in}} \times 100$$

where,

E_{out} is the peak-to-peak amplitude of ripple voltage appearing at the *output* (load) terminals of the voltage-regulated power supply,

E_{in} is the peak-to-peak amplitude of ripple voltage impressed by the rectifier and filter system (or other source) at the input of the regulating circuit.

For example, suppose a full-wave bridge rectifier—in conjunction with a single capacitor as the filter element—develops 100 volts DC with a 0.1 volt peak-to-peak ripple at 120 cps. This 120-cps ripple voltage is impressed at the input of a voltage regulator. Its peak-to-peak value across the voltage-regulator output terminals is .001 volt. The ripple factor is then $.001 \div 0.1 \times 100$, or 1% under stated load conditions.

Ripple factor is usually meant for "worst conditions," this implying full-load, but has a similar meaning for current-regulated supplies—provided we substitute currents for voltages. Here, we would be interested in the peak-to-peak currents which would flow into a stated load—with and without benefit of the current-regulating circuit.

COMPONENTS EMPLOYED FOR REGULATION

Varistor

A varistor is an element in which the relationship between impressed voltage and the resultant current is something other than the simple linear proportionality indicated by Ohm's law. However, not all nonlinear resistances are suitable for use as a voltage-regulating element in the shunt regulator circuit. To be useful for this application, it is necessary that the resistance of the element *decrease* in response to either the higher voltage impressed across it or the higher current passed through it. Thus a tungsten lamp, although a nonlinear resistance, would not perform as a shunt regulating element. The resistance of the tungsten filament increases with impressed voltage—or more precisely, with the current passed through it. On the other hand, a carbon filament lamp would provide voltage regulation when connected into a shunt regulator circuit (Fig. 1-10). The resistance of the carbon filament fulfills the requisite of decreasing with impressed voltage. (Here again, we of course realize that the decrease of resistance is due to the elevated temperature,

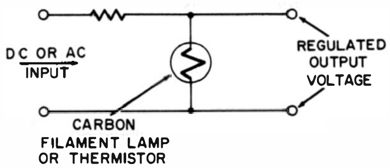


Fig. 1-10. Voltage regulation obtained by using a varistor in a shunt arrangement.

which in turn is due to the increased current brought about by the increased voltage.)

A varistor in which the resistance *increases* with current flow can provide voltage regulation when used in the series regulator circuit (Fig. 1-11.) However, this arrangement provides regulation against a varying line voltage only; voltage regulation with respect to a varying load resistance is poorer than if a linear resistance were used in the series arm.

The forward-conduction characteristics of semiconductor diodes such as point-contact, selenium, copper-oxide, and germanium and silicon junction diodes, can provide a regu-

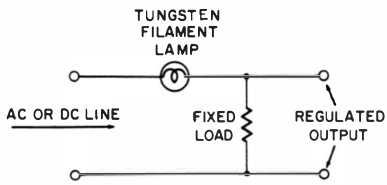


Fig. 1-11. Regulation against line-voltage variations by means of a series varistor.

lation percentage low enough to be useful in low-voltage applications. Although the regulated voltage provided by a single diode is generally less than one volt, any number of these diodes can be connected in series—in which case the nominal value of the regulated voltage is multiplied by the number of such series-connected diodes. However, at four or five volts the zener diode—because of its ability to provide much closer regulation—merits consideration.

Thyrite is the trade name of General Electric for a silicon carbide varistor. Unlike the semiconductor diode, *Thyrite* is nonpolar. *Thyrite* elements are available in a wide variety of power ratings, voltage ranges, and degrees of nonlinearity. The equation defining the nonlinearity of *Thyrite* is:

$$I = KV^n$$

where,

I is the current through the element,
K is a constant determined by the resistivity and geometry of the material,
V is the voltage impressed across the element,
n is an exponent, generally between 3 and 8.

Thyrite behaves like an element in which resistivity is controlled by the impressed voltage. In most other varistors, the change of resistance is a function of current.

Current Ballast Tubes

Current ballast tubes, also known as *Barretters,* were among the earliest means of immunizing electronic equipment against the adverse effects of changing AC line voltage. They remain popular as a simple and economical approach to the regulation problem, and presently are used for heater-current stabilization of series-connected tubes in AC-DC

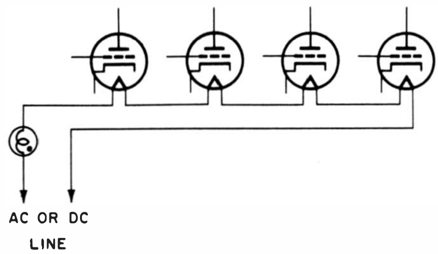


Fig. 1-12. A current ballast tube connected as shown will provide constant current to series heaters.

radio equipment (Fig. 1-12), for preregulation in conjunction with more complex and precise regulating techniques, and for indirectly providing voltage regulation in applications where close regulation is not necessary. This device is essentially an incandescent lamp incorporating an iron filament in a gaseous atmosphere of hydrogen. By proper choice

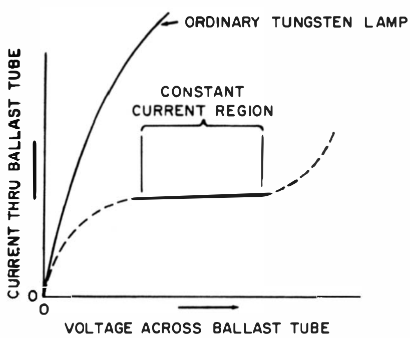


Fig. 1-13. Characteristic curve of a current ballast tube.

of filament geometry and gas pressure, it is possible to design such a lamp so that its voltage-current characteristic will be as depicted in Fig. 1-13. When operation is within the constant-current region of the device, a considerable change in terminal voltage results in a relatively small change in current. Consequently, when connected in series with the AC line, the current ballast tube will provide a nearly constant current to equipment powered from the line. When used in the primary circuit of a power transformer connected to a rectifier (Fig. 1-14), the DC load voltage will then be substantially independent of variations in the AC line voltage. (This type of regulation is not suitable for applications where appreciable variations occur in the load resistance. Load-voltage regulation with respect to load current is adversely affected by the insertion of the ballast tube into the AC line.)

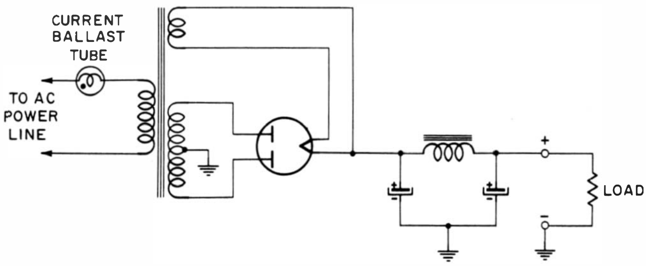


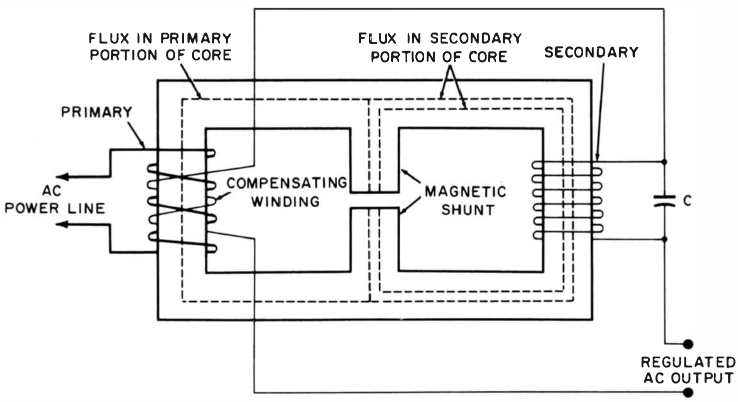
Fig. 1-14. Current ballast tube used in conjunction with a power supply.

The iron filament is subject to damage by external magnetic fields. For this reason, a magnetic shield is often used with ballast tubes. Filament temperature is below the values attained in ordinary light bulbs; nevertheless, considerable power dissipation is involved, and attention must be given to the matter of heat removal by free convection. If ventilation is poor, the tube may not operate within its intended constant current region. These tubes are also made with filamentary materials other than iron, and with gases other than hydrogen. Unless otherwise stated by the manufacturer, these tubes should be mounted vertically, base down.

Static-Magnetic Regulating Transformers

Special transformers are also available for reducing the effects of varying AC line voltage. These transformers, when utilized in conjunction with modern silicon rectifying elements, can provide good regulation with respect to both line and load (AC line regulation of 1% and a load regulation of about 3% are readily acheived). When such transformers are employed in addition to the regulation techniques described in this book, the effect of the AC preregulation greatly enhances the over-all performance of the regulated power supply. Such "constant-voltage" transformers, as they are sometimes called, are particularly useful with solid-state regulator circuits, because the transistor or controlled-rectifier regulating element need not absorb the excess power whenever the line voltage rises. Further protection is conferred by limiting the short-circuit current of the constant-voltage transformer to about twice the normal full-load current.

Pictorial and schematic diagrams of the basic constant-voltage (also called static magnetic regulating) transformer are shown in Figs. 1-15A and B, respectively. This trans-



(A) Wiring arrangement.

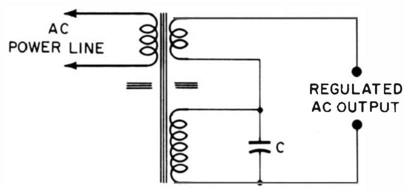


Fig. 1-15. The static-magnetic regulating transformer.

(B) Schematic symbol.

former differs from conventional power types in that a magnetic shunt is incorporated in the core geometry and a capacitor is connected across the secondary. These two features provide the common objective of causing more flux to circulate in the secondary portion of the core structure than in the primary portion. The capacitor consumes a large reactive current, and the resultant ampere-turns in the secondary winding produces a high flux. Because of the magnetic shunt, much of this flux is retained in the secondary portion of the core. The magnetic shunt functions as a magnetic isolator, to reduce the coupling between the primary and secondary windings. The significant feature is that the primary magnetic circuit operates *below* the "knee" of the core magnetization curve, whereas the secondary magnetic circuit operates *above* the knee—that is, in the magnetic saturation region. See Fig. 1-16. A change in the AC line voltage impressed across the primary winding will accordingly produce a proportionate change in flux in the primary portion of the core structure. Because of magnetic saturation, however, the accompanying flux change in the secondary portion of the core structure will be relatively small. The change in induced voltage appearing across the secondary winding will therefore be proportionately smaller than the change in voltage

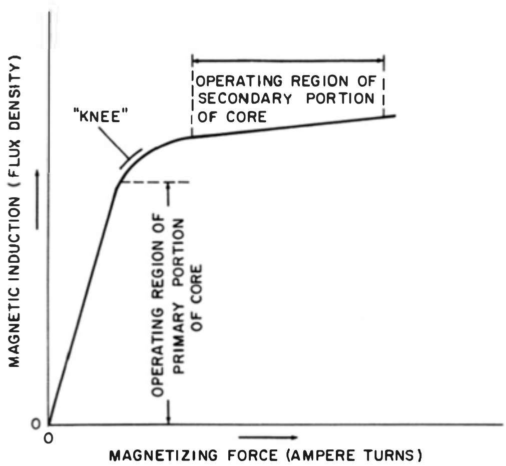
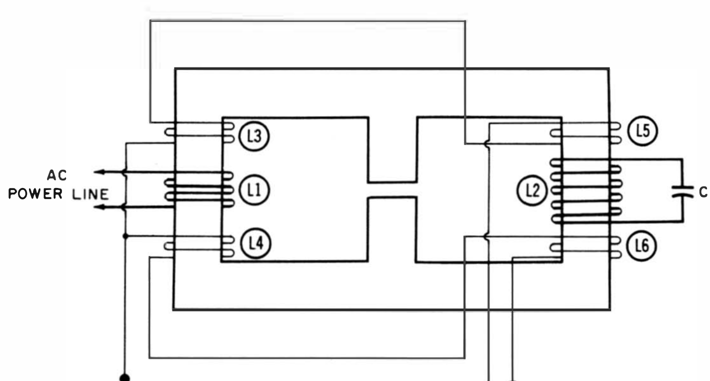


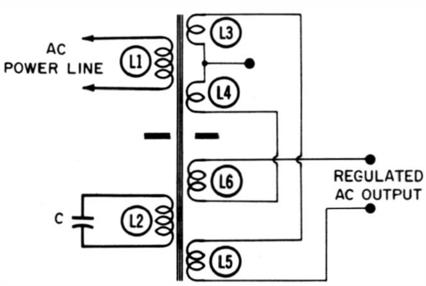
Fig. 1-16. Curve showing operating regions of primary and secondary core sections in a static-magnetic regulating transformer.

across the primary. Thus, this type of transformer regulates its own secondary voltage. A further refinement consists of a so-called compensating winding over the primary and connected in series-opposing with the secondary. The voltage change developed across this winding improves the regulation by introducing a small voltage, to compensate for the departure of the saturated region of the magnetization curve from the ideal horizontal slope (that is, from complete saturation). One of many possible modifications of this arrangement is shown in Fig. 1-17. Here, separate secondary windings are employed for providing capacitor and load currents. This technique permits electrical and physical utilization of the capacitor without regard to the independent requirements of the load. A similar method, often encountered, involves a single tapped winding.

Because the flux in the secondary portion of the core is increased by the capacitor reactive current, the induced voltage in the secondary winding obviously is higher than would be indicated by the numerical ratio of secondary to primary turns. This, however, does *not* mean that resonance has been established by the capacitor with respect to the inductance of the secondary winding. Resonance is utilized in other



(A) Wiring arrangement.



(B) Schematic symbol.

Fig. 1-17. Another type of static-magnetic regulating transformer.

types of regulating transformers, but not in the self-regulating transformer described here. The waveshape of the secondary voltage is closer to a square than a sinusoidal wave. This waveshape is actually desirable in many instances, because the rectifying elements are subjected to less peak AC voltage than when a sine wave is being converted to DC.

Transformer Regulation by DC Saturation

Fig. 1-18 depicts a transformer-rectifier arrangement which provides DC voltage regulation as a result of the DC load current flowing through saturable core transformers. In this circuit, T1 is a conventional power transformer, and its core operates below the knee of its magnetization curve. Therefore, any induced secondary voltages will be propor-

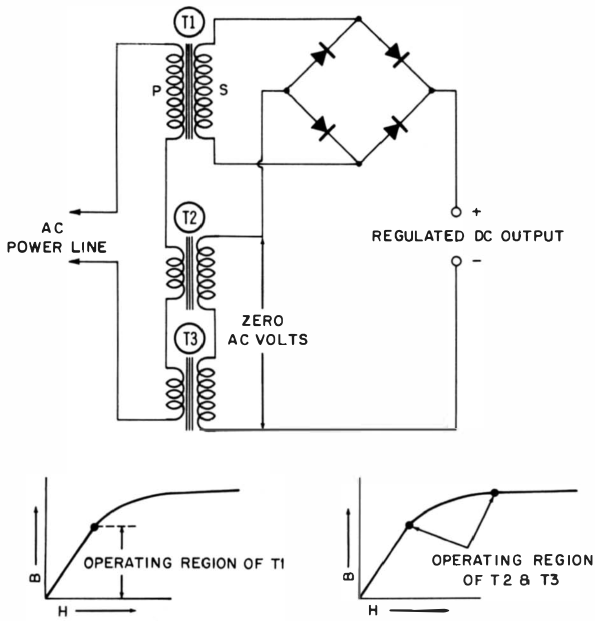


Fig. 1-18. A transformer regulating arrangement using the variable saturation method.

tionate to the impressed primary voltage. As a result, no voltage or current limiting takes place. However, the cores of transformers T2 and T3 operate in a region of their magnetization curves wherein core permeability can be controlled by the amount of current passing through their DC windings. Accordingly, the inductance of the AC windings of T2 and T3 is a function of the DC load current.

If the load current tends to increase because of a change in line or load conditions, the resultant increase in the core saturation of T2 and T3 will *lower* the effective inductance of their AC windings. This, in turn, will permit more AC voltage to be impressed across the primary of T3. The tendency for load voltage to fall because of the increased load current will thereby be counteracted. Should the lead current tend to fall, the converse sequence of events will take place and thereby counteract the accompanying tendency of the lead voltage to rise.

Since the secondary windings of T2 and T3 are connected in series opposition, any AC voltages induced in these windings will cancel each other in the DC load circuit. Hence, it follows that saturable transformers T2 and T3 must have identical characteristics; otherwise, AC ripple will be injected into the load.

General Aspects of VR Tubes

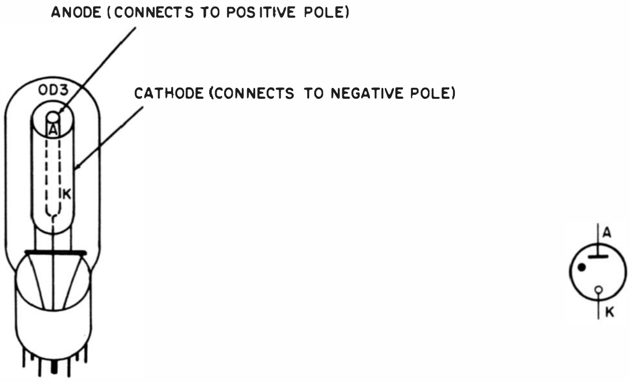
Voltage-regulating tubes are gaseous diodes making use of various mixtures of inert gases such as neon, argon, krypton, and xenon to provide a selection of several ionizing voltages in the normal glow portion of their conduction characteristics. Ordinary neon lamps with two identical pinlike electrodes also display constant-voltage characteristics over a portion of their operating curves. These lamps are inferior to VR tubes, however, because a relatively small current will project them into the abnormal glow region, which is unsuitable for voltage stabilization. The unique structure of the elements of a VR tube enable a relatively high current to be consumed before the abnormal-glow mode of operation is reached. As will be shown, the more current a shunt regulating element can consume while still exhibiting the type of current-resistance relationship of Table 1-1, the heavier the load that can be handled.

Fig. 1-19A shows the basic construction of a voltage-regulator tube. Observe that the *anode*—that is, the positively polarized element—is the much *smaller* element. The cathode, or negatively polarized element, resembles in both structure and orientation the plate (anode) of an ordinary electron tube. This fact plus widespread usage of the symbol in Fig. 1-19B—which depicts the anode as the platelike, or *larger*, element—naturally leads to confusion. As a consequence, it is not uncommon for technicians or even engineers to sometimes connect VR tubes incorrectly.

VR tubes have either internal jumpers or multiple-pin connections, which enable wiring the power-supply circuit in such a manner that no voltage is delivered to the load when the VR tube is not in its socket. This is a protective feature—with no VR tube, the voltage applied to the load will be considerably higher, and there is a possibility of damage. The socket connections of the common VR tubes

Table 1-1. Voltage-regulator tube characteristics

Type	V_{out} Nom. DC Operating Voltage	I_{min} Min. DC Operating Current MA	I_{max} Max. DC Operating Current MA	Approx. DC Starting Voltage	Momentary DC Starting Current MA	V_{in} Min. DC Supply Voltage	Approx. Regulation
OA3	75	5	40	100	100	105	6.7%
O82	108	5	30	115	75	133	0.93%
OC3	108	5	40	115	100	133	1.9%
OA2	151	5	30	156	75	185	1.3%
OD3	153	5	40	160	100	185	2.6%



(A) Construction. (B) Schematic symbol.

Fig. 1-19. The voltage-regulator tube.

are shown in Fig. 1-20. When an OA3, OC3, or OD3 is used, it is possible to utilize the internal jumper at pins 3 and 7 (Fig. 1-20A) as a means of breaking the AC supplied to the primary of the power-supply transformer. With the OA2 and OB2 (Fig. 1-20B), multiple-element connections rather than an isolated jumper are provided, so that the positive or negative DC output load is interrupted whenever the tube is removed from its socket.

(A) OA3, OC3, OD3. (B) OA2, OB2.

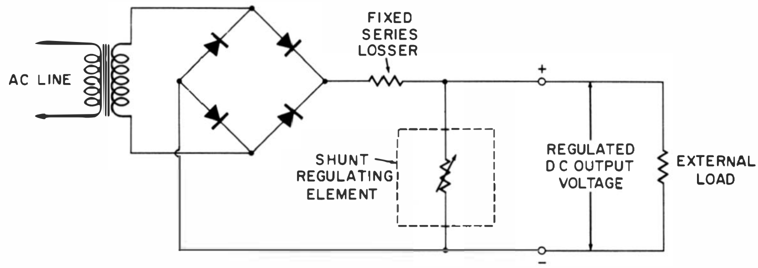
Fig. 1-20. Voltage-regulator tube pin connections.

BASIC TYPES OF AUTOMATIC REGULATION

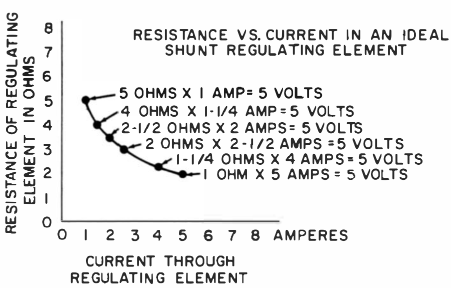
There are a number of ways to classify regulating circuits, and various circuits are employed. For our purpose, we will (at least initially) find it advantageous to assign voltage- or current-regulating techniques to one of the two basic family groups.

Open-Loop Regulators

The first of these, the open-loop regulator, utilizes a voltage- or current-sensitive element which automatically changes its internal resistance in such a way that the load voltage (or in some instances, the load current) does not change. Some of the elements which enable this to be accomplished in suitable circuits are gaseous diodes such as VR (voltage-regulator) and corona tubes, reverse-biased junction (zener) diodes, forward-biased diodes (*Stabistors, Varistors,* and vacuum diodes), and elements having negative coefficients of resistance with respect to voltage or current (carbon filament lamps, *Thyrite* cartridges, and *Thermistors*). The general circuit used for obtaining stabilized voltage with such elements is depicted in Fig. 1-21A.



(A) Circuit.



(B) Characteristic curve.

Fig. 1-21. Typical open-loop regulating circuit utilizing a shunt element to stabilize the output voltage.

The curve in Fig. 1-21B, which depicts the behavior of an ideal variable-resistance element, provides considerable insight into the operation of simple open-loop voltage-regulated power supplies. This curve has the interesting property that the product of the two quantities defining any point on the curve is of constant value. Thus, the product of the horizotnal units (amperes) and vertical units (ohms) is *five* in this particular example. Specifically, we see that five times one, four times one and one-quarter, two times two and one-half, etc., all yield the product five. Inasmuch as this product is derived from multiplying current by resistance, we have the condition wherein 5 volts is maintained for any current on the curve. From this mathematical property, we deduce that an element displaying such a relationship must maintain a constant voltage across itself and therefore in a load connected across it. Gaseous diodes known as VR (voltage-regulator) tubes, and reverse-biased silicon junction diodes known as zener or breakdown diodes, closely approach the above described relationship. They are therefore the most frequently encountered shunt regulating elements.

Closed-Loop Regulators

The other basic regulating technique is represented by the *closed-loop,* or "error-sensing," circuit. Although such circuits can be subdivided, the essential ingredient of closed-loop automatic regulation is control of the output voltage

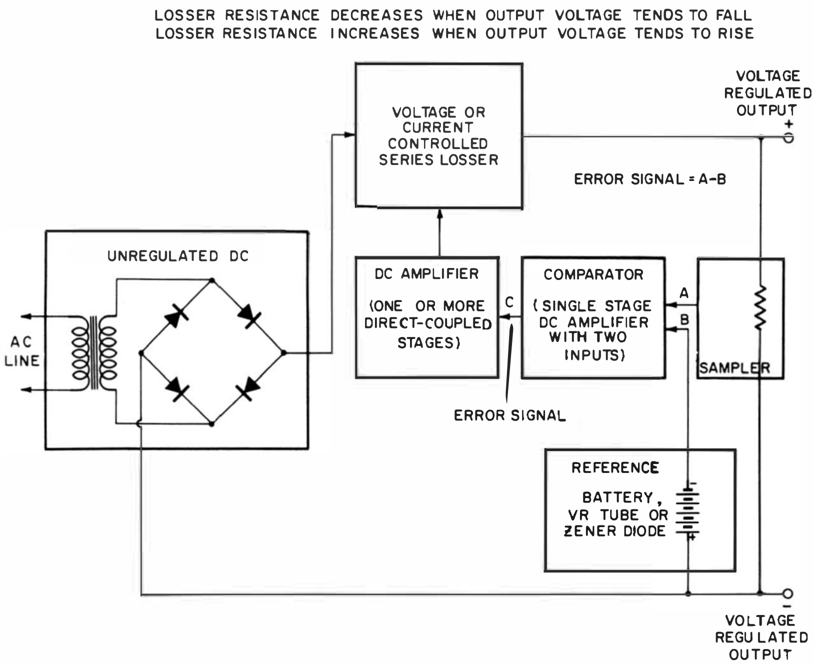


Fig. 1-22. Typical closed-loop arrangement for producing automatic voltage stabilization.

(or current) in response to an *error signal* resulting from comparison of a portion of the output with a stable voltage source known as the reference. The reference-voltage source is often a simple open-loop regulator circuit. Thus, a shunt regulator circuit of the type shown in Fig. 1-21, which uses a VR tube or zener diode, is one of the building blocks of the more complex closed-loop regulator.

The closed-loop regulator illustrated in the block diagram of Fig. 1-22 makes use of a series "losser" element, the resistance of which is controlled by the amplified error signal. Should the DC output voltage tend to rise, the resultant error signal will *increase* the resistance of the series losser and thereby counteract the voltage rise. The converse events occur if the output voltage tends to fall. In this way, the output voltage is maintained almost constant. The error voltage always tends to extinguish itself by fixing the output voltage at such a value that a sampled portion of it is equal to the reference voltage.

The power demand from the voltage-reference source is generally a small fraction of that which would exist in a shunt-type voltage regulator supplying a similar external load. Consequently, it is relatively easy to provide electrical and thermal operating conditions for the shunt element which optimize its voltage stability and thus enhance the stability of the over-all regulating circuit.

SOME CONSIDERATIONS ABOUT THE RECTIFIER-FILTER SYSTEM

Because of the relative ease with which regulation can be accomplished electronically, the rectifier-filter system often is not given design consideration much beyond that pertaining to production of the nominally required voltage and current. However, it should be appreciated that the ultimate regulation and ripple suppression depends upon *both* the rectifier-filter system and the regulator. Indeed, when exceedingly good performance is required in these respects, it may be more feasible to enhance the operation by improving the rectifier-filter system rather than by attempting improvements upon the regulator proper. A technique which should not be overlooked is the use of a swinging choke in the power supply. The regulation of a choke input, capacitor output, full-wave power supply is shown in a general way in Fig. 1-23. The major contributor to regulation in region A-B is resistance in the transformer, rectifier bridge, and choke winding. The major contributor to regulation in region B-C is capacitor C1 because L1 does not have any appreciable effect. Note the abrupt change in regulation encountered as load current is decreased from point A (which might correspond to maximum rated current) to point B. This is the consequence of the choke becoming abruptly ineffective when no longer carrying current beyond a critical value. At these low currents the choke no longer prevents the rectifier

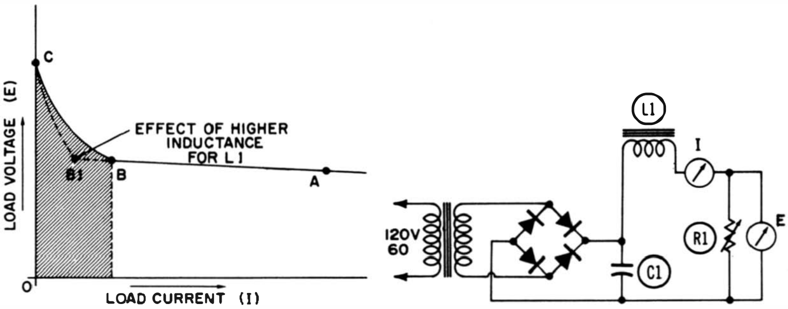


Fig. 1-23. Graph showing poor regulation region at low currents with choke input filter.

current pulses from reaching peak and zero values. Therefore, the output capacitor tends to charge to peak voltage; this manifests itself as a rapid rise in output voltage as load current is decreased to zero.

Higher inductance will cause point A to occur at a lower current. Unfortunately, practical difficulties impose limitations before this design approach can be pursued very far. A choke with higher inductance also has higher resistance if the core size is maintained. This adversely affects regulation and defeats the basic objective even though the regulating curve can be made somewhat more consistent throughout the load range. Off-setting this by increasing both wire and core size obviously only increases weight and cost. A practical compromise can be affected by taking into account that the flywheel effect of the choke is a function of both inductance and current. This indicates that less inductance will suffice at higher rather than at lower load currents to provide good regulation.

Such reasoning leads to the concept of the so-called *swinging choke*, one that will swing its inductance in such a way that high inductance will prevail at low load currents and low inductance at higher currents. The low inductance at higher currents is not the desirable operational mode; rather it constitutes the practical solution in that lower inductances *suffice* at higher load currents. The important feature is high inductance at low load currents because this permits good regulation over a wide load current range. Such a choke is much smaller and less costly than one designed with a nearly constant value of inductance equal to the highest value displayed by the swinging choke. The change in inductance in the swinging choke is due to magnetic saturation of the core and is the consequence of greatly reducing or

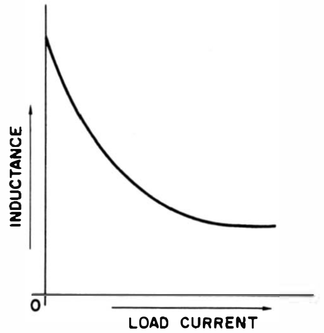


Fig. 1-24. Graph showing inductance versus load current of a swinging choke.

eliminating the air gap ordinarily provided in conventional constant inductance chokes. Fig. 1-24 depicts the inductance of a swinging choke as a function of load current.

Even with the use of the swinging choke, an artificial load or bleeder may have to be connected across the output of the power supply in order to prevent voltage rise at load currents approaching zero. However, the power dissipated by such a bleeder will be very much less than the bleeder needed to pacify the low current regulation with a conventional choke. It is desired that bleeder power should be low; for such power dissipation subtracts from the current available to the load and degrades the efficiency of the system. Fig. 1-25 illustrates the fact that critical inductance occurs at a lower value with a heavy load than with a light load.

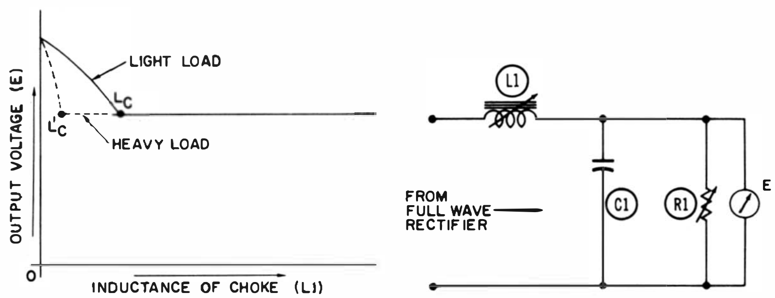


Fig. 1-25. Graph showing the effect of choke inductance on the output voltage for light or heavy loads.

L1 in the diagram is not a swinging choke, but a manually adjusted variable inductance to make it easier to determine L_C and voltage reduction.

The critical inductance (L_C) of a choke is readily calculated in terms of several parameters of the power supply as shown in the following formula:

$$L_C = \frac{R}{p\pi f_p\ (p^2 - 1)} \text{ henrys}$$

where,

R is the combined resistance of the transformer, rectifiers, and load,

f is the frequency in cycles per second

p is the number of rectifiers firing per input cycle of power line (p equals 2 in single-phase full-wave and bridge rectifier currents).

(With semiconductor rectifiers, rectifier resistance can generally be neglected.)

In the common situation where f corresponds to 60-cps AC source and where full wave, single phase rectification is used, this formula reduces to:

$$L_C = \frac{R}{1130} \text{ henrys}$$

In actual practice it is always wise to design in the direction of greater inductance; often circumstances permit a minimum inductance several times greater than L_C. Such *excessive* inductance is not wasted, but confers greater attenuation of ripple.

A swinging choke is not useful for half wave rectification because infinite inductance is required to achieve L_C for any load current. Half-wave rectification has inherently poor regulation and thereby requires correspondingly better stabilizing performance from the regulator. Its use is limited to high voltage and/or low current systems, or where the specifications are not demanding.

For reasons of economy, the choke is generally less critically chosen in terms of its inductance. Thereafter, filtering action is most cheaply acquired by using the output capacitor C1. Size and cost become prohibitive here too. At this point, the voltage regulator contributes *electronic filtering* equivalent to the effect of the impractical size of the chokes and capacitors. However, the initial filtering must be provided by the rectifier-filter system, otherwise, the amplitude of the ripple impressed upon the voltage regulator will carry it beyond its operating range. In such a case, the regulator can neither suppress ripple nor regulate effectively.

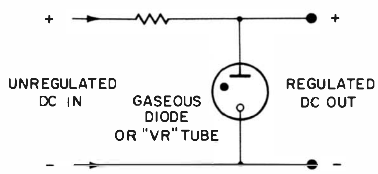
When a single output capacitor is used, with or without a series resistance, a brute force approach is generally employed in that the capacitor is made very large. This has become feasible through the availability of relatively small and inexpensive electrolytic capacitors. Because of the high capacitance which can be so used and because of the low voltage drop in semi-conductor rectifiers, the regulation provided by such a technique is adequate for many practical purposes. It is often necessary to include a small resistance in series with the rectifier output in order to protect the rectifiers from the current surge when the capacitor is initially charged.

CHAPTER 2

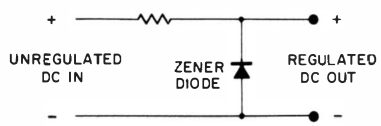
Open-Loop Circuits Using VR Tubes

GENERAL CONSIDERATIONS

Most nonfeedback, or open-loop, regulating circuits are shunt-type voltage stabilizers. Generally, a gaseous diode or a semiconductor (zener) diode is utilized as the shunt element. Additionally, a series element comprising a fixed resistance appears in these circuits. The two most common shunt-type voltage-regulating circuits are shown in Fig. 2-1. They can be briefly described as variable voltage dividers in which the shunt arm automatically adjusts its resistance to maintain a constant voltage across itself and the load. Although this simple description is entirely valid, we will gain even more insight into open-loop regulation by investigating this principle in detail.



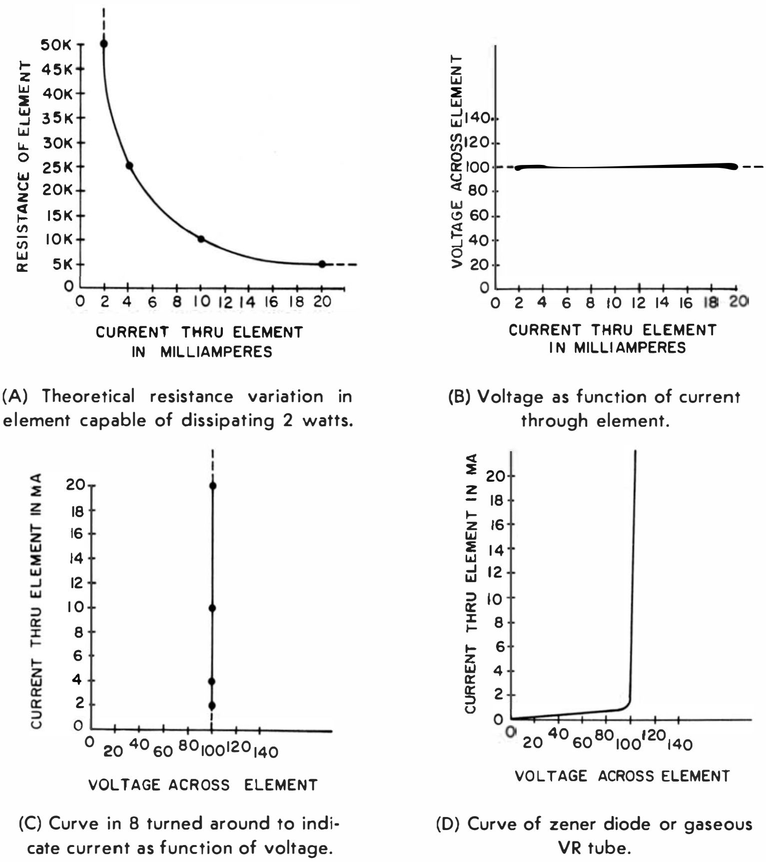
(A) Using voltage-regulating tube.



(B) Using zener diode.

Fig. 2-1. Shunt-type voltage-regulator circuits.

The ideal shunt regulating element is one which changes its internal resistance in such a way that the current flowing through it does not change the voltage across it. In Fig. 2-2A is depicted a resistance-versus-current curve which illustrates the basic concept of such an element. Note that the application of the Ohm's-law relationship for voltage, $E = I \times R$, always yields the same voltage, no matter what current is selected. For example, the product of 2 milliamperes and 50K ohms, 10 milliamperes and 10K ohms, 4 milliamperes and the corresponding 25K ohms, etc., is always 100 volts. The voltage constancy despite current change is shown in Fig. 2-2B. In a very true sense, the shunt regulation element is a current-sensitive nonlinear resistance which is *inversely proportional* to the current flowing through it.



(A) Theoretical resistance variation in element capable of dissipating 2 watts.

(B) Voltage as function of current through element.

(C) Curve in B turned around to indicate current as function of voltage.

(D) Curve of zener diode or gaseous VR tube.

Fig. 2-2. Characteristics of a 100-volt shunt regulating element.

Mathematically, such an element would be defined as conforming to the equation:

$$R = \frac{k}{I}$$

where,

R is the resistance of the element,
I is the current flowing through the element,
k is a proportionality factor numerically equal to the voltage developed across the element.

Our equation can be transformed into R = E ÷ I, which to all appearances is one of the Ohm's-law derivatives. However, voltage E in this application is *fixed,* thereby forcing resistance R to become a function solely of current I.

As a result of supplying current to such a shunt element, its behavior—from the viewpoint of the external load—is suggestive of a battery with perfect (zero) regulation. Fig.

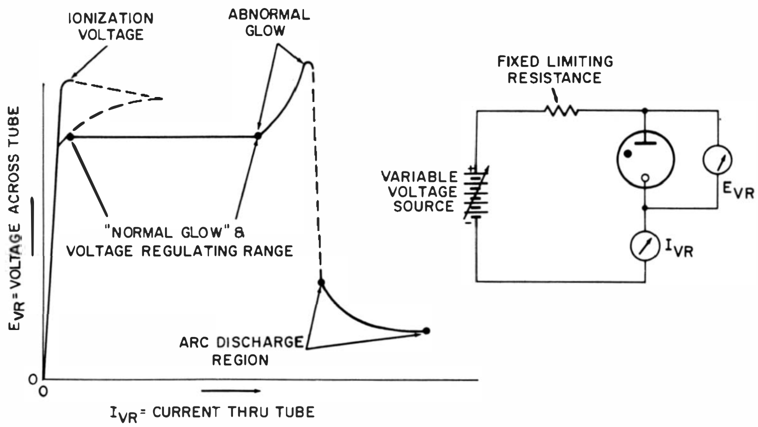


Fig. 2-3. The voltage-current characteristics of a gaseous VR tube.

2-3 shows the voltage-current characteristic of a VR tube. The normal glow region has the constant voltage previously demonstrated with idealized shunt regulating elements.

The Need for Series Limiting Resistance

The series element provides a function beyond acting as the fixed section of a variable voltage divider. Fig. 2-2C is a duplication of Fig. 2-2B, but with the axis of the graph restated. *Current* through the shunt element is now shown as a function of the *voltage* impressed across the element. We see that for less than 100 volts, the current is zero. At 100 volts, the current flow theoretically is infinite for our ideal element. In a practical element, the current flow is still extremely large—many tens or hundreds times more than the device can withstand. In other words, an element of this type tends to be *self-destructive*—the flow of current raises the temperature, lowering the resistance and enabling still more current to flow. A cumulative cause-and-effect relationship would lead to a catastrophic thermal runaway if the current flow were not limited somehow. Fortunately, the series element can be designed to have a resistance which limits the current through the shunt regulating element to values which are simultaneously safe and within the proper region of the characteristics, insofar as voltage stabilization is concerned.

Another Effect of Series Limiting Resistance

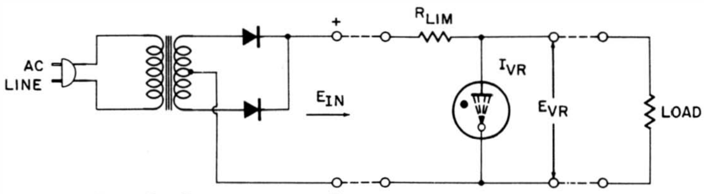
An immediate consequence of series resistance is that the input voltage must be *higher* than the stabilized output voltage. The current flowing through the series element divides between the shunt element and external load. The voltage drop due to this total current flowing through the series element is deprived from the shunt element and load. If this voltage drop is excessive, the shunt element will not receive enough voltage to operate within its voltage-stabilizing region. Although it is convenient to speak of "current-actuated" devices such as transistors, electromagnetic solenoids, etc., we can thereby find ourselves at odds with the physical facts of electricity. Whatever occurs in these devices is always primarily due to the application of a voltage; i. e., current flow is caused by electrical pressure. This is true even in a series circuit, wherein we say that the "voltage drop across such and such a resistance is 'due' to such and such a current flowing through it," but it is truer to say that whatever current flows through the resistance is *due to the voltage impressed across it.* Thus, lest we misconstrue the implication of Fig. 2-2A, we must realize that this characteristic exists only if 100 volts can be impressed across the element. In Fig. 2-2C we therefore see that the device would be *inactive* unless a minimum of 100 volts were available.

Unfortunately the accompanying voltage drop, as the load current flows through the series element, ultimately deprives the shunt regulating element of its operating voltage as the load is made heavier. The series element thus establishes the upper limit to the amount of current the load may consume; once this limit is exceeded, the shunt regulating element can no longer stabilize the load voltage. Fig. 2-4 depicts the basic equation relating the relevant parameters of a VR-tube shunt regulating circuit.

VOLTAGE STABILIZATION IN OPEN-LOOP CIRCUITS

Voltage Stabilization with Fixed Input and Varying Load

Figs. 2-5, 2-6, and 2-7 show how voltages and currents change in order to permit the VR tube to maintain a constant output voltage under the condition of fixed input voltage and varying load current. Voltage stabilization is maintained as long as the VR tube is supplied with current within the flat (horizontal) portion of its operating characteristics (Fig. 2-5). It is interesting to observe that current stabilization also exists in this circuit (Fig. 2-6). I_A, the total current consumed from the source of input voltage, remains constant



$$R_{LIM} = \frac{E_{IN} - E_{VR}}{I_{VR}}$$

WHERE: E_{IN} IS THE MINIMUM SUPPLY VOLTAGE ANTICIPATED, AT THE HEAVIEST LOAD & LOWEST LINE VOLTAGE,
E_{VR} IS THE OPERATING VOLTAGE OF VR TUBE,
I_{VR} IS THE MAX. ALLOWABLE CURRENT IN VR TUBE-IN AMPS.

NOTE: E_{IN} SHOULD EXCEED E_{VR} BY AT LEAST 1.3 IN ORDER TO ASSURE STARTING, EVEN UNDER HEAVIEST LOAD & LOWEST LINE VOLTAGE.

Fig. 2-4. Calculating the series limiting resistance of a shunt regulating circuit.

despite variations in the load current (within the regulating range of the VR tube). There is another quantity which remains fixed as long as output-voltage stabilization exists; this is the equivalent resistance of the load and VR tube in parallel. It is this quantity which is responsible for the constancy of output voltage and supply current. That this is so follows from the basic voltage-divider configuration of the regulator circuit (Fig. 2-7). Table 2-1 shows the relationship between various values of load resistance and currents within the circuit.

Voltage Stabilization with Varying Input Voltage and Fixed Load

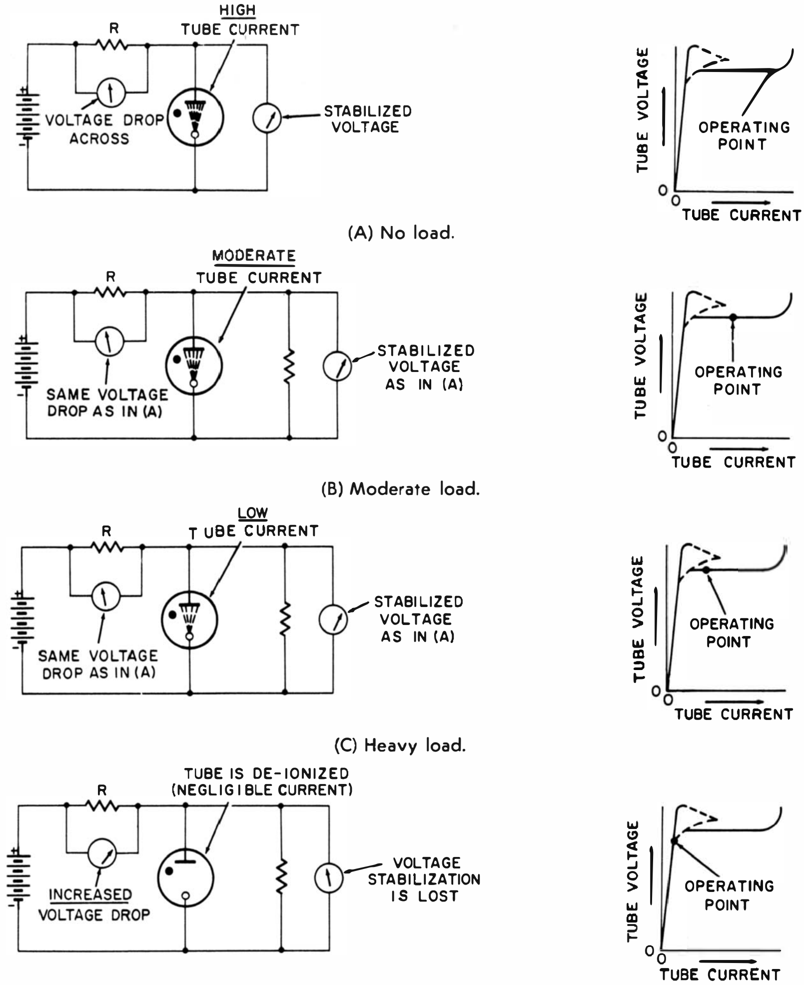
When the load current is maintained constant but the input voltage is permitted to vary, the circuit action is somewhat different from that described for the converse set of conditions. Fig. 2-8 shows the effects of variations in the source voltage on the output voltage. From Table 2-2 and Fig. 2-9 we now see that the output-voltage stabilization is accompanied by a constant load current I_C. The variation in VR tube-current consumption is exactly what is required to enable the shunt arm (load and VR tube) to relate to the

Table 2-1. Effect of fixed source voltage and various values of load resistance on circuit of Fig. 2-7A

Load Resistance in Ohms	Current in VR Tube in MA (I_b)	Current In Load in MA (I_c)	Current From Supply in MA $I_a = I_b + I_c$	Equivalent Resistance of VR Tube in OHMS	Resistance Due to VR Tube and Load in Parallel in OHMS	Voltage Across Load V_{load}
Infinite	43.2	0	43.2	2500	2500	108
20,000	37.8	5.4	43.2	2857	2500	108
10,000	32.4	10.8	43.2	3333	2500	108
5,000	21.6	21.6	43.2	5000	2500	108
3,000	7.2	36.0	43.2	15000	2500	108

Table 2-2. Effect of fixed load resistance and various source voltages on circuit of Fig. 2-9A

Supply Voltage (V supply)	Voltage Drop Across Series Resistance (V_R)	Voltage Across Load (V_{load})	Load Current (I_c) in MA	Tube Current (I_b) in MA	Supply Current (I_a) in MA	Equiv. Resist. of VR Tube and Load in Parallel Ohms
150	42	108	21.6	27.8	49.4	2186
145	37	108	21.6	21.6	43.2	2500
135	27	108	21.6	10.2	31.8	3396
130	22	108	21.6	4.3	25.9	4066



(A) No load.

(B) Moderate load.

(C) Heavy load.

(D) Excessive load.

Fig. 2-5. Effect of various loads on regulator circuit voltages.

series-dropping resistance in such a way that a constant load voltage is maintained in the face of a changing input voltage.

Variations in Both Input Voltage and Load

In general usage, the input voltage and the load both vary. As a consequence, both circuit actions previously described take place simultaneously. Despite this more complex operation, the load voltage is stabilized, for each of the described actions responds uniquely to the conditions responsible for it. The load current and the voltage drop across the series limiting resistance will no longer be constant, as they were in the two experiments. Nevertheless, the *resistance* of the VR tube still adjusts to values which always result in a constant output voltage.

THE VR TUBE AND ITS APPLICATIONS

Load Capabilities of VR Tubes

Most VR tubes are rated at currents between 5 and 40 milliamperes. It is an interesting fact that, as long as the

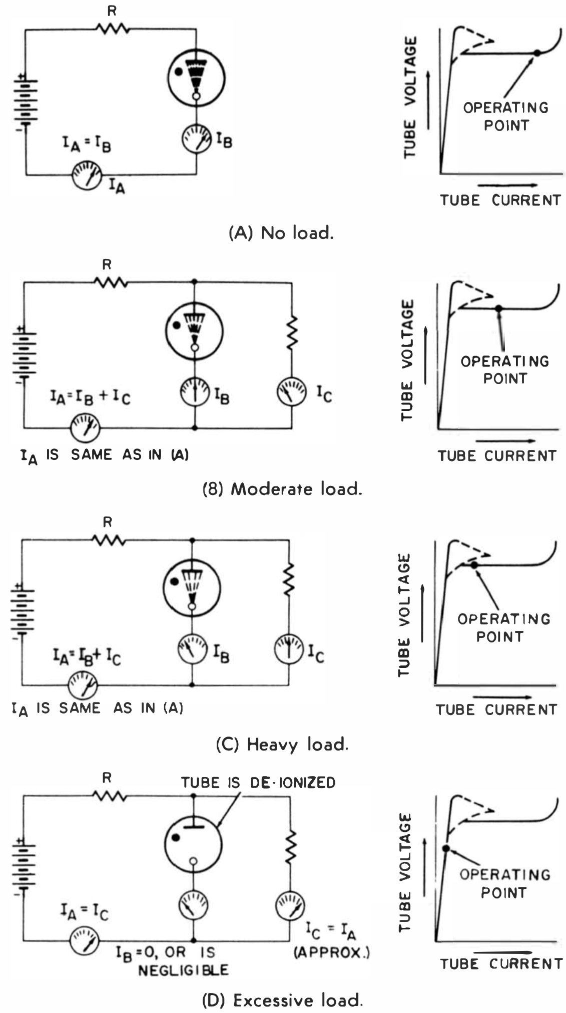
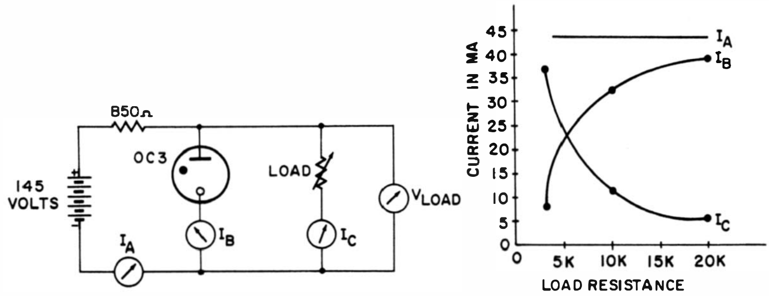


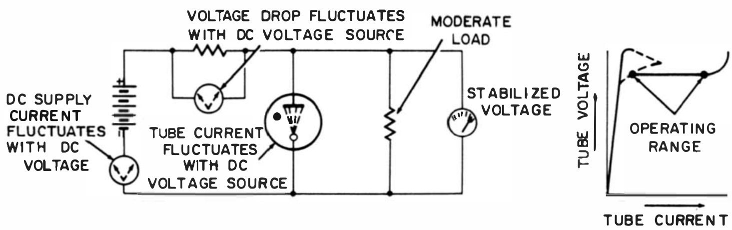
Fig. 2-6. Effects of various loads on regulator circuit currents.



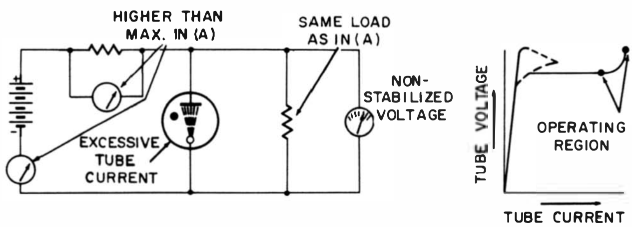
(A) Circuit.

(B) Current-resistance curve.

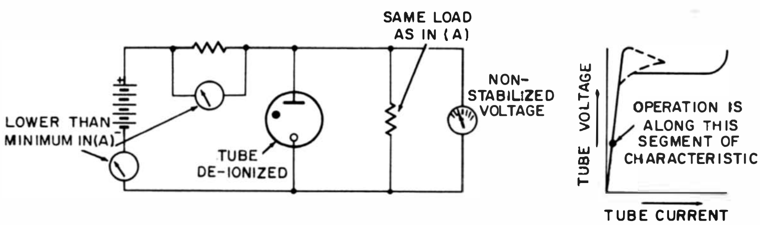
Fig. 2-7. Regulator circuit action when source voltage is fixed and load is variable.



(A) Moderate DC voltage variations.



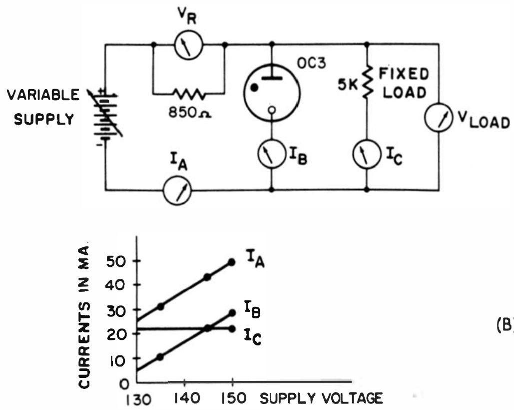
(B) DC voltage too high.



(C) DC voltage too low.

Fig. 2-8. Effects of source voltage variations on voltage-regulator circuit.

VR tube current is maintained within this range, no limitation is imposed on the current consumed by the load. Certain practical matters make it difficult or undesirable to attempt voltage stabilization for high-current loads, however. A heavy current load necessitates a relatively low series dropping resistance; otherwise, the DC supply voltage from the rectifier would have to be inordinately high. If such a load became disconnected from the output terminals, the VR tube would be damaged or destroyed by the heavy cur-



(A) Circuit.

(B) Current-voltage curve.

Fig. 2-9. Regulator circuit action when load is fixed and the source voltage is variable.

rent which would then flow through it. Allowing for this possibility, however, there is no reason why the VR tube could not stabilize the voltage across a load drawing much more current than the tube itself. For example, if the intention is to stabilize against line-voltage fluctuations and it is ascertained that the load, although heavy, is relatively constant and always present, a VR tube can be used for higher load currents than would be permissible with a varying load.

Ignition Voltage for Gaseous Tubes

Gaseous diodes such as VR tubes require *higher* than the operating potential to start ionization of the gas. Now another requirement is imposed on the rectifier or unregulated DC supply—the minimum voltage available under the worst conditions must be sufficient to fire the tube. Otherwise, tube conduction will be negligible and voltage stabilizing action will not be attained. "Worst conditions" generally involves the *simultaneous* occurrence of full load and low line voltage. Sometimes other conditions aggravate the problem of starting. Experiments have shown that a gaseous diode with a normal glow voltage of approximately 70 volts requires over 200 volts to fire it when placed under refrigeration and in the absence of light. The energy of light photons is imparted to the gas molecules and also produces photoelectric emission from the electrodes. As a consequence, it is good practice to provide a suitable margin of surplus starting voltage—particularly if separation is to take place in total darkness. Under ordinary conditions (at or above room temperature and with even the small amount of light available from the tube filaments), the starting voltage of VR tubes is generally from 8 to 20 volts higher than the operating voltage. Some tubes now contain a radioactive isotope, which brings about reliable starting at lower potentials.

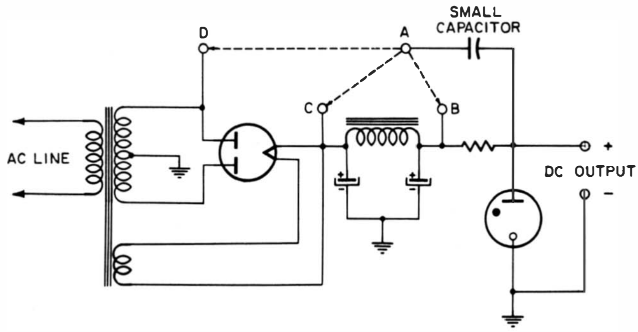


Fig. 2-10. Methods of connecting a small capacitor to aid in initial ionization of VR tube.

It is possible to initiate ionization with a lower DC supply voltage if a small capacitor is connected across the series dropping resistor. The capacitor bypasses a portion of the otherwise undesirable ripple component around the resistor, making it available as additional starting potential. If the rectifier output passes through a filter prior to being impressed across the VR tube, the bypass capacitor should then be connected between the VR-tube anode and the positive output terminal of the rectifier, or to one of the high-voltage terminals of the power transformer. See Fig. 2-10. Here, lead A of the capacitor can be connected to point B, C, or D. The last is the most effective of the three. Sometimes the voltage stabilizing action or the filtering effect of a VR tube is desired for a circuit not previously designed to furnish the required starting voltage. If sharp pulses of sufficient amplitude, or audio or radio frequencies of suitable voltage level, are available elsewhere in the equipment, the introduction of such energy through a small capacitor connected to the anode of the VR tube can facilitate firing. Sometimes, better results are obtained by connecting the small capacitor to one of the "blank" pins of the VR tube. The electrified point of the pin within the tube then becomes a copious producer of ions.

Combinations of VR Tubes

Parallel-Connected VR Tubes—It is possible to obtain certain special results by appropriate combinations of VR tubes. One quite likely to come to mind would be parallel operation in order to increase the current-handling capability of the load. However, if two or more VR tubes are simply paralleled, it is highly probable that only *one* will ignite—the resultant lowering of output voltage by the ionization of one tube will tend to prevent the others from starting. This is undesirable in another respect. Since the circuit is designed for a high current output, its series limiting resistance would

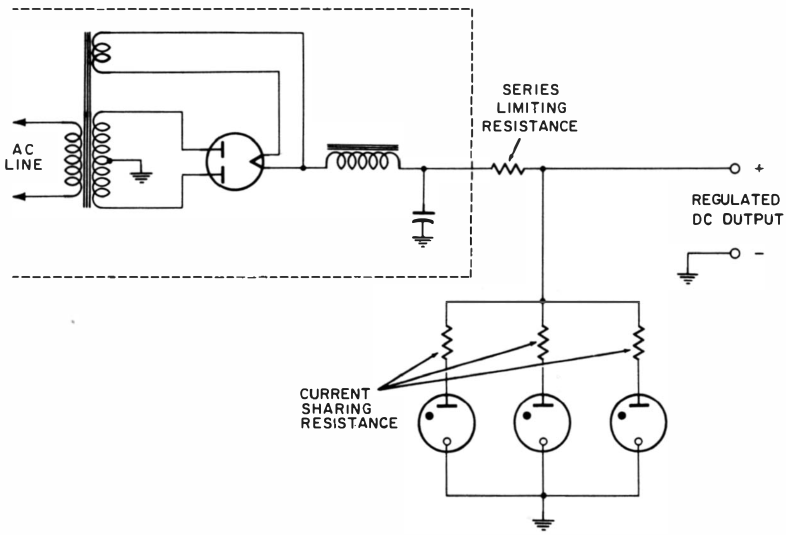


Fig. 2-11. Method of connecting VR tubes in parallel.

be too low to prevent damage to the tube "lucky" enough to ignite at the expense of the others. Even if two selected tubes were utilized in such a fashion that both were assured sufficient starting potential and current, subsequent operation would tend to be unstable as one tube or the other hogged most of the available operating current. Current sharing can be enforced by means of individual series resistances, as in Fig. 2-11. In this way, parallel operation can be employed to provide a stabilized voltage to heavy loads. However, insertion of the current-sharing resistances will seriously degrade the regulation and dynamic output impedance. Hence, except where poor stabilization is better than none, paralleling of VR tubes in this manner is not recommended.

Series-Connected VR Tubes—VR tubes connected in series constitute a practical circuit configuration with several operational advantages, such as stabilization of *higher* voltages than can be accomplished with one VR tube alone. Although it is preferable to employ tubes with identical current ratings, any combination of voltage breakdowns may be selected. If one tube has a lower rated maximum current, the series limiting resistance and the input voltage should be so related that *this* current is never exceeded in *any* tube. Another advantage of series-connected VR tubes is the availability of more than one stabilized DC voltage, as shown in

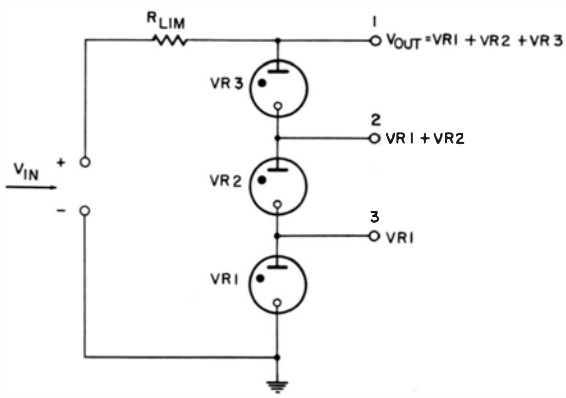


Fig. 2-12. VR tubes connected in a series arrangement.

Fig. 2-12. Here we see that, in addition to ordinary VR-tube action, such an arrangement functions as a voltage divider. The number of stabilized voltages provided is equal to the number of series-connected VR tubes.

In Fig. 2-12, due consideration should be given to the fact that *upper* tubes are forced to carry the current consumed by any loads connected to the *lower* taps. For example, if a load is connected between ground and tap 3, VR2 and VR3 must supply this current, even though the voltage to this load is stabilized by VR1. As a consequence, the allowable load currents available from taps 1 and 2 are lower. Similarly, if taps 2 and 3 are supplying current to loads, the permissible current output from tap 1 is diminished by the sum of the currents supplied by taps 2 and 3.

Cascaded VR Tubes—In Fig. 2-13 we see a cascade, or tandem, arrangement of VR tubes. In this circuit, the VR150 constitutes a stabilized input-voltage source for the operation of the VR75. This technique greatly enhances the stability of the VR75, which is now virtually relieved of the

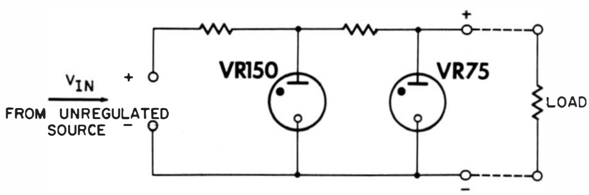


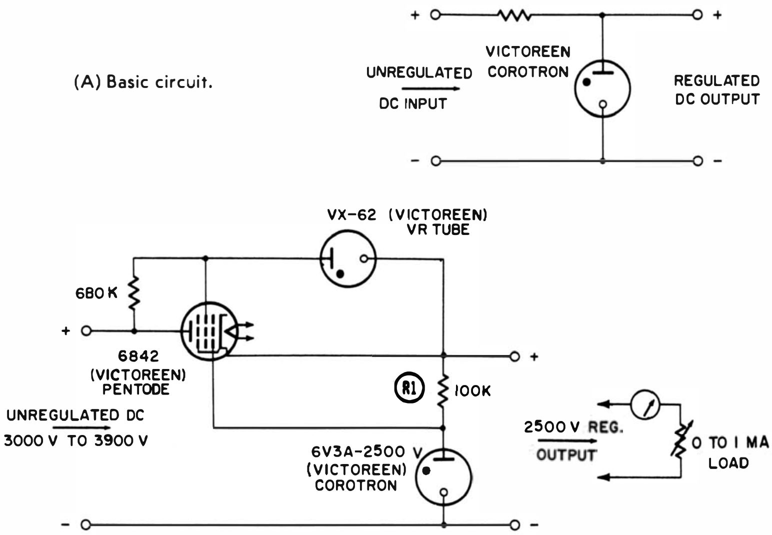
Fig. 2-13. Cascaded VR tubes.

need to regulate with respect to the varying input voltage. A requirement of this circuit is that the tube receiving its operating power from the unregulated source must have a higher breakdown voltage than the tube which stabilizes the load voltage. Because of the limited number of breakdown voltages available in VR tubes, this arrangement is not as flexible with VR tubes as it is with zener diodes, which may be had in a wide variety of breakdown voltages.

High-Voltage Regulated Supplies Using Corona-Type Tubes

By suitable choice of gas pressure and electrode geometry, a gaseous diode can be designed to display a voltage-regulating region at much lower currents than those required to produce a glow discharge in a conventional VR tube. At the same time, this regulating region can be made to correspond to electrode potentials at least as high as 25 kilovolts. Ionization at these relatively high voltages and low currents is essentially a *corona* phenomenon. These tubes are known as *Corotrons* and are manufactured by the Victoreen Instrument Company. The simplest type of regulator circuit is identical in configuration to the conventional VR-tube circuit (Fig. 2-14A). The same manufacturer also makes high-voltage pentode tubes, for use with the *Corotron* in closed-loop regulator circuits. A typical circuit of this type is shown in Fig. 2-14B. Observe that a separate voltage amplifier is not employed to boost the level of the error signal before it is applied to the grid of the pentode losser tube. This in itself is not a novel circuitry feature. There is, however, more than initially meets the eye. The voltage drop across R1 (Fig.

2-14B) is a small fraction of the total output voltage. Therefore, the feedback factor is not divided down appreciably by the sampling network (R1 and the *Corotron*). This fact, in conjunction with the high voltage gain of the pentode, results in a loop gain as high as might be obtained in conventional lower-voltage regulators utilizing one or two stages of DC voltage amplification prior to the losser element. By dispensing with the cascaded DC amplifiers, this circuit achieves low dynamic impedance with very little of the ordinarily accompanying drift.



(B) High-voltage pentode used in conjunction with corona tube.

Fig. 2-14. High-voltage corona-tube regulator circuits.

Disadvantages of Gaseous Diode Voltage Regulators

Before the advent of the semiconductor reference element (zener diode), it would have served no useful purpose to cite the shortcomings of the gaseous VR tube, for nothing better was available. Since this is no longer true, it should prove instructive to summarize the undesirable characteristics of the VR tube:

1. The requirement of a higher-than-operate starting voltage.
2. The possibility of spurious signal generation (relaxation oscillations, ionic or molecular oscillations, and "hash" or noise voltage).

3. The availability of the device in only a few operating voltages.
4. Limited life span.
5. Unpredictable and erratic tendencies with regard to both the starting and the operating voltages.

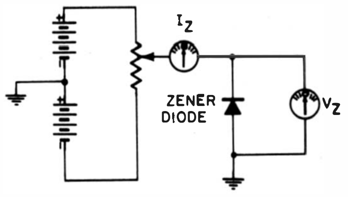
CHAPTER 3

Open-Loop Circuits Using Zener Diodes

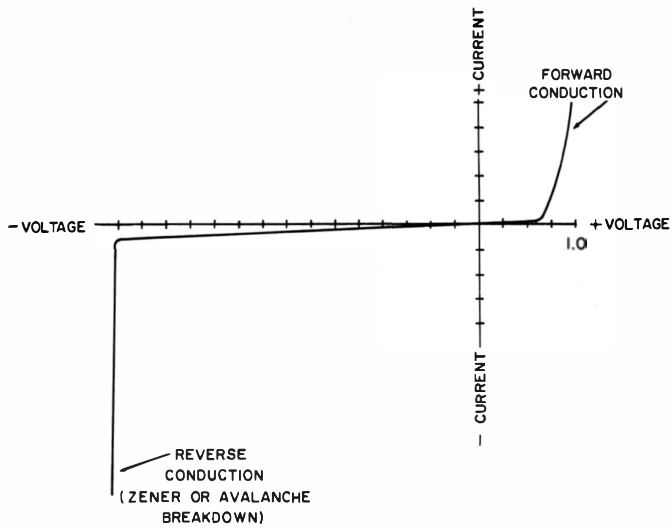
THE ZENER DIODE

The zener diode essentially is a silicon junction rectifier which exhibits abrupt conduction at a discrete *reverse bias.* Unlike semiconductor diodes used for rectification, wherein reverse conduction must be *avoided,* the zener diode is deliberately used in this operating mode. When so used, its effective DC resistance varies with the current, in a manner closely approaching the ideal characteristic of the shunt regulating element depicted earlier in Fig. 2-1A. Thus, the zener diode can provide a constant voltage to a load, in much the same way previously described for the gaseous tube. There are numerous applications where high current capability takes precedence over extremely close regulation. This requirement is most simply and, often, most economically met by means of a shunt regulator employing a power zener diode. Also, back to back power zeners make satisfactory regulators for AC voltage in many cases. A considerable art has evolved about the zener diode, and, to thoroughly understand this fascinating device, one must become acquainted with the physics of semiconductors (which is outside the domain of this book). Rather, we shall be concerned only with the external characteristics of the device, and, in particular, with those related directly to its use as a shunt-type voltage regulator.

From the basic circuit in Fig. 3-1A you can obtain a better understanding of the operation of a zener diode. The characteristic curve of Fig. 3-1B is generally representative of any semiconductor rectifying diode. However, the relatively abrupt reverse conduction and nearly vertical slope of the curve in the reverse-conduction region are *specifically* characteristic of the zener diode. Disregarding the sign of the polarizing voltage and omitting the forward-conduction region, we can depict the curve as in Fig. 3-2. This clearly shows the similarity of the zener diode to the gaseous VR tube and thus enables us to visualize both devices as practical elements approximating the voltage-current relationship of the ideal element represented by the curve of Fig. 2-2D.



(A) Circuit.



(B) Characteristic curve.

Fig. 3-1. General characteristics of zener diodes.

Operating Voltages

Unlike with gaseous diodes, a wide selection of breakdown voltages—ranging between 3.9 and 200 volts—is available in zener diodes (see Table A-3 in the Appendix). As the semiconductor art progresses, even higher voltage units will probably become practical. The physics of the breakdown phenomenon is such that the closest approach to ideal shunt regulating characteristics is obtained for diodes designed for reverse-conduction voltages between 6½ and 8½ volts. For this reason, it is sometimes desirable to employ series-connected strings of such low-voltage units in order to achieve a higher breakdown voltage. Table A-3 shows Motorola power zener diodes in 10 watt and 50 watt capacities. These have breakdown voltages ranging from 3.9 volts to 200 volts. In all instances, the diodes have reverse polarity when the letter R is appended to the type number. Reverse

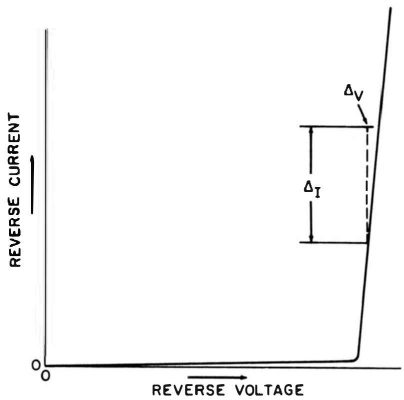


Fig. 3-2. Simplified method of depicting the breakdown voltage characteristics of the zener diode.

polarity corresponds to packaging in which the negative connecting terminal (anode) is insulated from the stud or case. Such an expedient results in more precise DC regulation and lower dynamic output impedance than can be obtained from a single element rated at the higher voltage. At the same time, such practice can entail greater expense. In Fig. 3-3 we see a typical family of curves for zener diodes having various breakdown voltages. Keeping in mind that a *vertical slope* in the reverse-conduction region would provide perfect regulation and a zero dynamic output impedance (at least at low frequencies), we can acquire a feeling for the relative "goodness" of such diodes in terms of their zener breakdown voltages. Inasmuch as zener diodes are employed most often in conjunction with transistors and other solid-state devices, it fortunately happens that breakdown voltages in the 6½- to 8½-volt region are practical for circuit operation. Notice that the curves for diodes B, C, and D of Fig. 3-3 are nearly vertical; this signifies low dynamic impedance.

Table A-1 in the Appendix comprises a family of General Electric one-watt zener diodes. It is recommended that such

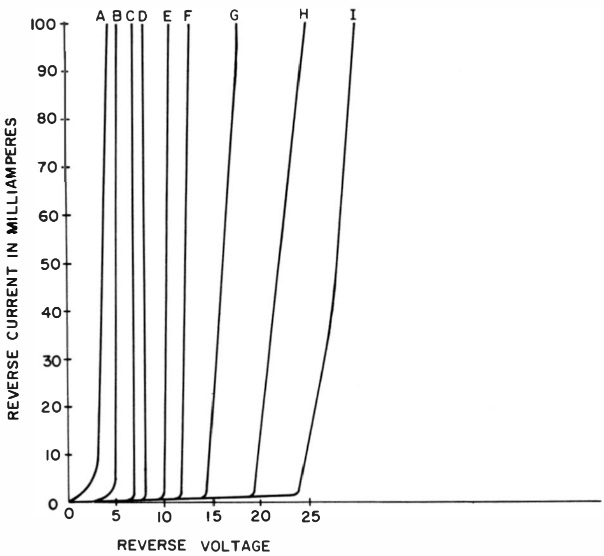


Fig. 3-3. General trend of reverse-conduction characteristics for zener diodes with different breakdown voltages.

one-watt units be substituted for their voltage counterparts in schematic diagrams calling for 200 mw, 400 mw, and 750 mw zener diodes. The use of one-watt units is advantageous for construction of experimental projects. The ad-

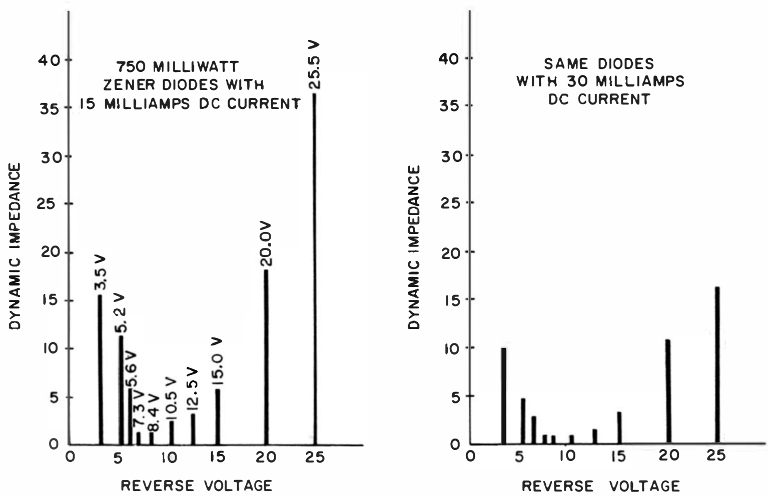


Fig. 3-4. Effect of operating current on dynamic impedance of various zener diodes.

ditional electrical ruggedness of these one-watt diodes provides increased protection against destruction with the often accompanying chain reaction burn-out of associated semiconductors. An additional bonus is obtained if the diode current is increased, for then a lower dynamic impedance results. The one-watt zener diode has been so reduced in price that initial selection cannot be rejected on cost alone.

Dynamic Impedance

Effect of DC Current on Dynamic Impedance—In a given zener diode, the dynamic impedance decreases as the DC operating current rises. This is illustrated in Fig. 3-4, which shows the dynamic impedances of various diodes corresponding to two DC currents. Of course, the DC operating current cannot be increased without ultimately exceeding the permissible temperature rise of the junction. Inasmuch as the zener diode sometimes must dissipate considerable ripple power, the DC current alone may not constitute sufficient consideration in designing for safe operation. Safe operation may not actually be the limiting factor in deciding the maximum DC current to pass through a zener diode. The temperature coefficient of zener diodes with breakdown voltages greater than about six volts becomes increasingly positive with more DC current. Unless compensating components with suitable negative temperature coefficients are present in the circuit, it may not be desirable to pass too high a DC current through a zener diode because the voltage breakdown could shift considerably with temperature variations. Aside from the effect of more or less DC current, conditions accounting for variations in the operating temperature of a zener diode are (1) an initial warm-up period, (2) proximity of heat-generating components, and (3) changes in the ambient temperature.

In order for manufacturers' values of dynamic impedance to be meaningful, it has become standard practice to designate the dynamic impedance of a zener diode for DC operating currents of 20% of maximum permissible DC current at 25°C. Additionally, the measurement is carried out under the stipulation that the AC superimposed on the DC operating current is 10% of that value and is a low-frequency sine wave, generally 60 cps. For example, a zener diode capable of withstanding a maximum DC current of 200 ma would be tested for dynamic impedance by superimposing 4 ma of 60-cps alternating current on 40 ma of DC operating current. The dynamic impedance would then be calculated by dividing by 4 ma the value of 60-cps voltage needed to produce the 4-ma AC current. If, in this particular example, 8 millivolts is required, the dynamic impedance of this zener diode would be quoted as .008 ÷ .004, or 2 ohms.

Variations in Dynamic Impedance with Rated Power Dissipation—It has been shown that increased DC current in a given zener diode tends to *lower* the dynamic impedance. Graphically, this is manifested by the fact that the characteristic curve is almost a vertical slope. If the DC current is made too high, however, excessive heat will be produced. This results in an unstable coefficient of temperature with respect to voltage—that is, the breakdown voltage tends to lose even its short-term stability. A further increase in DC current will damage the junction. This behavior would, however, lead us to suspect that a zener diode with a junction having a large area—that is, one designed to carry more current—would have a lower dynamic impedance than a small diode. From a highly different viewpoint, we could liken a large-area junction to an equivalent parallel connection of an appropriate number of smaller junctions. Now we would expect the dynamic impedance of the large junction to be less than that of any smaller junction, because impedances in parallel result in a smaller net impedance than that of any of the constituents. As a matter of fact, zener diodes of *high power rating* do tend to have *lower dynamic impedances* than diodes with like breakdown voltages but smaller power or current ratings.

In Tables 3-1, 3-2, and 3-3 we see a significant comparison of zener diodes belonging to three different families of power ratings. In all three, the maximum permissible current is the power rating divided by the breakdown voltage, and the DC test current for evaluation of dynamic impedance is approximately 20% of the maximum permissible DC current.

Dynamic Impedance and Series-Connected Diodes—Figs. 3-3 and 3-4 indicate that dynamic impedance tends to in-

Table 3-1. Dynamic impedance versus power rating—typical 750-milliwatt zener diodes

Type	Nominal Breakdown Voltage	Maximum DC Current MA	Test Current for Z_z MA	Dynamic Impedance Z_z Ohms
1N1507	3.9	180	35	14.0
1N1510	6.8	110	22	1.5
1N1512	10.0	75	15	1.8
1N1514	15.0	50	10	5.0
1N1517	27.0	26	5	50.0

Table 3-2. Dynamic impedance versus power rating—typical 3.5-watt zener diodes

Type	Nominal Breakdown Voltage	Maximum DC Current MA	Test Current for Z_z MA	Dynamic Impedance Z_z Ohms
1N1588	3.9	850	150	2.6
1N1591	6.8	525	100	0.6
1N1593	10.0	350	70	0.7
1N1595	15.0	225	40	3.4
1N1598	27.0	125	25	13.0

crease quite rapidly for diodes with breakdown voltages greater than 10 or 15 volts. Other things being equal, it is generally possible to obtain lower dynamic impedances by allowing more DC current to flow through a diode. This can

Table 3-3. Dynamic impedance versus power rating—typical 10-watt zener diodes

Type	Nominal Breakdown Voltage	Maximum DC Current MA	Test Current for Z_z MA	Dynamic Impedance Z_z Ohms
1N1599	3.9	2500	500	0.84
1N1602	6.8	1500	300	0.20
1N1604	10.0	1000	200	0.55
1N1606	15.0	650	140	1.50
1N1609	27.0	350	70	4.50

even be exploited to the extent that a diode with a much larger power capacity is used than the demands of the external load would dictate. This is a practical technique for lowering the dynamic impedance for diodes with breakdown voltages higher than 20 volts or so, particularly when 50 to 100 or more volts must be stabilized. However, a better approach consists of connecting the appropriate number of low-voltage diodes in series. For example, three 8.4-volt diodes in series will result in an effective breakdown and sta-

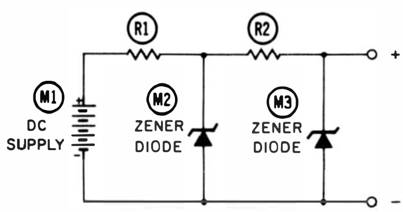


Fig. 3-5. Two-stage cascaded shunt regulator.

bilization voltage of about 25 volts. The dynamic impedance of the combination will be approximately the same as for a single diode, but the permissible power dissipation will be three times as much. Any number of low-impedance diodes can be series-connected to provide stabilization at a higher voltage.

Dynamic Impedance and Cascaded Shunt Regulators— The dynamic output impedance of a zener-diode voltage can also be lowered by cascading two or more stages. A two-stage arrangement is depicted in Fig. 3-5. The diode used in the first stage (M2) must have a higher breakdown voltage than the output stage. This arrangement significantly improves regulation with respect to the line or input voltage.

Stabilization of Small Voltages

Below 3.5 volts, the characteristics of zener diodes deteriorate to such an extent that these units are not generally available commercially. Even between 3.5 and 4 volts, the onset of conduction is "soft"; and to attain a low dynamic impedance, such diodes must be operated with appreciable current. This is undesirable, for as we shall see, the already excessive temperature drift of low-voltage diodes is further degraded by the high current.

A very effective way of obtaining low-voltage stabilization is by means of the parallel network shown in Fig. 3-6. The output voltage is the difference between the breakdown voltage of the two diodes. Inasmuch as the two diodes will be similar, they may be expected to have fairly closely-matched temperature drift characteristics. The output voltage will then be relatively immune to temperature, because the fact that the two diodes change in the same direction tends to maintain the same difference voltage. As an example, if 8.4- and 7.2-volt diodes are used, a stabilized output of 1.2 volts will be provided.

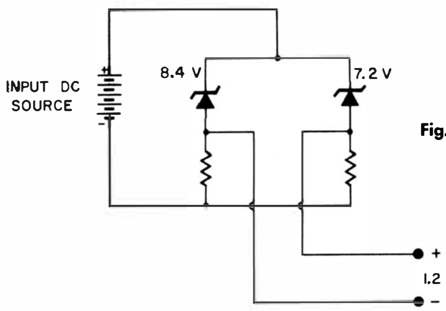


Fig. 3-6. Parallel network for obtaining low-voltage stabilization.

Temperature Coefficient

The zener breakdown voltage tends to be temperature-dependent. For many applications, the displacement of voltage with respect to temperature is too small to prove troublesome. It is fortunate that, generally speaking, the temperature dependency and the dynamic impedance are both at or close to a minmum at breakdown voltages between 5½ and 8 volts. The exact voltage at which either or both the temperature dependency and the dynamic impedance are a minimum is partially a function of the manufacturing process and the operating current. The temperature coefficient is the measure of voltage change with respect to temperature. Temperature coefficient is given in per cent per degree centigrade (%/°C.). Inasmuch as the temperature dependency may not necessarily be a linear relationship for wide temperature variations, the temperature coefficient implies the average rate of change with respect to an ambient temperature of 21° centigrade. Suppose, for example, a zener diode is rated at 6 volts breakdown at 20 milliamperes and at 21°C. ambient; under these conditions, the manufacturer designates the temperature coefficient as +.05%. This means that a 10° rise in temperature will change the zener breakdown voltage 0.18 volt, or 180 millivolts ($10 \times .05\% \times 6$). The new breakdown voltage at the elevated temperature will therefore be 6.18 volts.

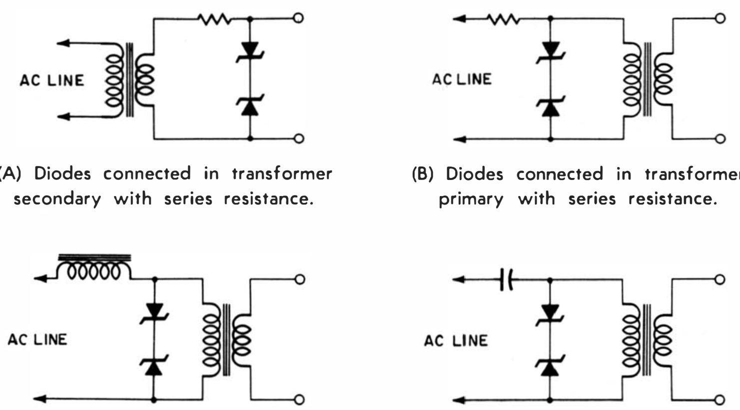
Thus, we see that the effect of temperature tends to degrade the precision at which DC voltages can be stabilized. For certain applications it is imperative that attention be given to this matter. It is not always the best approach to design the zener diode for zero temperature coefficient. Sometimes it is easier to insert or to take advantage of another element with an equal and opposite temperature coefficient. The over-all temperature coefficient will then be zero, and the temperature dependency will be neutralized throughout the tracking range of the zener diode and the compensating element. An important application of this technique is found in the substantially zero temperature-coefficient voltage-reference elements made by a number of semiconductor firms. These consist of one or more series-connected zener diodes with positive temperature coefficients; also connected in series are one or more forward-biased junction diodes. That is, the polarities of the diodes are such that the zener diodes operate in their normal reverse-current region. They therefore exhibit the abrupt breakdown-voltage characteristic of such operation. The polarity connections of the compensating diodes are such that they operate in their forward-conduction region with the same current which flows through the zener diodes. Silicon junction diodes generally develop a potential drop of ¾ to 1 volt when biased in the forward-conduction region. At the same time, they exhibit a *negative* temperature coefficient. The object of this arrangement is, of course, to attain exact neutralization of the temperature coefficient by the opposing temperature dependencies of the forward-biased compensating diodes and reverse-biased zener diode(s). The diodes are packaged as an integral unit. Even greater stability of temperature immunity is obtained by stabilizing the current permitted to flow through the reference unit. In extremely critical applications, the reference unit is operated inside a thermostatically regulated oven, much like quartz crystals.

The dynamic impedance of these reference units is somewhat *higher* than can be obtained from a single zener diode having the same breakdown voltage. This is due to the presence of the forward-biased diode(s). However, these units are not intended for use in simple shunt-type voltage regulators. Rather, they are designed for use in feedback, or closed-loop regulator, circuits wherein amplification is employed to effectively *reduce* the dynamic impedance of the reference element insofar as the "viewpoint" of the *load* is concerned. Thus, the achievement of a stable and substantially zero temperature coefficient at the expense of dynamic impedance constitutes a good bargain in the over-all picture for the intended application.

AC Regulation with Zener Diodes

Fig. 3-7 illustrates several ways of stabilizing AC voltage by means of zener diodes. Here, reverse-voltage breakdown is utilized in conjunction with series resistance—much as in DC circuits. However, two diodes are necessary in order that both the positive and the negative excursions of the AC wave will be stabilized. The two diodes exchange conduction roles with the alternations of the AC wave. Whichever diode happens to be forward-biased at a particular portion of the cycle will behave essentially as a straight-through connection to bias the other diode in its reverse, or zener, mode of conduction. The output wave is a trapezoid by virtue of the clipping action of the diodes. Fig. 3-7A provides AC regulation in the secondary winding, whereas regulation takes place in the primary in Figs. 3-7B, C, and D. In the latter two circuits, a reactance is substituted for the series resistance. The advantage of using reactance instead of resistance is that the necessary series voltage drop is obtained with negligible expenditure of power. In all circuits the zener diodes consume an appreciable portion of the AC input power. In Figs. 3-7B, C, and D, the transformer is not burdened with this extra power drain. Even so, there is always an additional core loss due to the nonsinusoidal shape of the impressed voltage wave.

The AC regulator circuits of Fig. 3-7 are very useful for supplying filament power to tubes involved in critical circuit functions. For example, the triggering point of multivibrators can be made less erratic; again, the frequency stability of oscillators is generally improved when a regulator source of filament power is used. A regulated filament supply additionally tends to extend tube life. These AC regulation cir-



(A) Diodes connected in transformer secondary with series resistance.

(B) Diodes connected in transformer primary with series resistance.

(C) Diodes connected in transformer primary with series inductance.

(D) Diodes connected in transformer primary with series capacitance.

Fig. 3-7. Methods of obtaining AC regulation with zener diodes.

cuits may also be employed to provide preregulation for the various closed-loop regulators to be described in Chapters 4 and 5. Such a technique is most applicable to smaller capacity supplies.

CHAPTER 4

The Closed-Loop Regulation Circuit

Many voltage and current regulators use tubes or transistors as "losser" elements. These, in response to an error signal, change resistance in the appropriate direction to maintain the output voltage or current at a fixed value. Regulated supplies of this type are closed-loop systems—that is, essentially *feedback amplifiers.* An effective way to become acquainted with the circuit techniques used in these supplies is to investigate a number of schematic diagrams and thereby "acquire a feel" for the basic objectives involved in circuits having diverse configurations.

In Fig. 4-1 we see four basically similar arrangements of series-connected losser elements and voltage references. In these circuits the losser element functions as a cathode (or emitter) follower, with a tendency to maintain the output voltage close to the voltage developed across the reference element. An immediate advantage over the zener diode and VR tube regulators described in Chapter 2 is that the voltage-reference element is relieved of carrying the load current. Consequently, it becomes practical to stabilize large load power with small reference elements. The reference element and its associated circuitry can then be designed primarily for thermal stability. The dynamic output resistance of the power supply is much less than that of the voltage reference element, the improvement being directly proportionate to the current-amplification factor of the series losser element.

Most voltage regulated power supplies derive their operation from the action of a series losser element inserted in one lead. The losser element may be used in either of two ways and this determines the design and characteristics of the associated component sections of the supply. The two configurations are shown in Figs. 4-2A and 4-2B. In Fig. 4-2A the series losser operates as an emitter follower. This

(A) PNP transistor with battery voltage reference.

(B) PNP transistor with zener-diode voltage reference.

(C) NPN transistor version of circuit in B.

(D) Vacuum-tube circuit with VR tube.

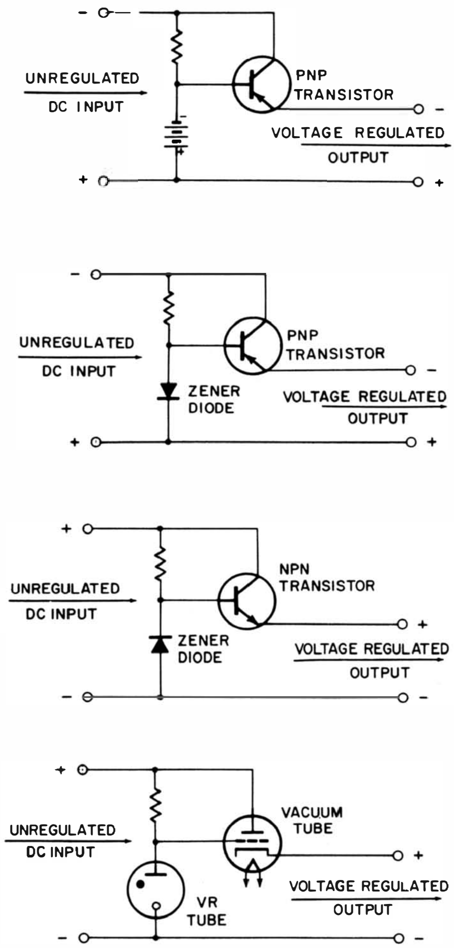
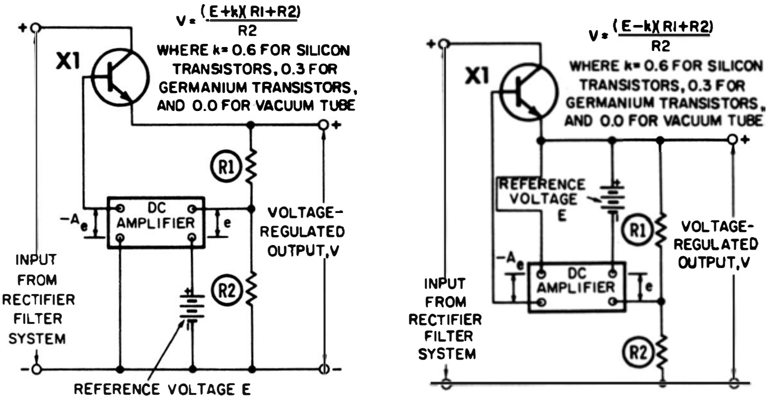


Fig. 4-1. Basic emitter- or cathode-follower voltage-regulator circuits.

is because the amplifier error signal $-A_e$ is effectively applied to the base with respect to the collector and the output voltage is derived from the emitter with respect to the collector of X1. This may not be immediately obvious inasmuch as neither the amplified error signals nor the voltage regulated output involved direct connections to the collector of X1. However, insofar as the dynamic operation of X1 is concerned, both input and output voltages vary with respect to the collector as *common element*.

The circuit in Fig. 4-2B makes use of X1 as a common-emitter amplifier. Here again, a second look may be neces-



(A) Series losser operated as an emitter-follower.

(B) Series losser operated as a common emitter amplifier.

Fig. 4-2. Two basic circuit arrangements for series losser regulated power supplies.

sary. X1 does not appear as a conventional common emitter amplifier inasmuch as the output circuit is associated with the emitter rather than the collector. Nonetheless, it can be readily seen that all criteria for common emitter operation are satisfied; the amplifier error signal is introduced at the base with respect to the emitter of X1. Moreover, from a dynamic standpoint, output variations actually occur in the collector circuit with respect to the emitter. The emitter is *common element* to one side of both input and output voltage variations.

It is interesting to compare the two configurations. It might initially appear that the common emitter arrangement, by virtue of its inherent voltage amplification must make better use of a given number of power supply elements than the emitter follower which provides *no* voltage amplification. However, things do not actually work out this way. The common-emitter stage exceeds the common-collector stage both in voltage gain and in output impedance by about the same factor. Voltage gain of the common emitter circuit is used up in reduction of its relatively high output impedance. Thus, dynamic output impedance of both circuits becomes about the same in the closed-loop regulator.

Often there are practical reasons which favor the use of one circuit over the other. For example, the comparator, DC amplifier stage(s), and particularly, the last DC amplifier stage (the DC stage which drives either the losser transistor or its associated Darlington stage) can be designed to handle voltages in the neighborhood of the emitter-base voltage of the *common-emitter* type losser. This is true even though the output terminal voltage of the regulator may be quite high. Conversely, with the *common-collector* (emitter follower) type of series losser, the DC amplifier stages must either be designed to withstand approximately output terminal voltage or must be protected from it. The latter-design technique can be accomplished for the last DC voltage amplifier stage by inserting a high-voltage zener diode in its collector lead. The zener diode "soaks up" most of the DC voltage without degrading amplification.

On the other hand, for low and medium output voltages, say under 30 or 40 volts, the common-collector series losser may lead to more economical design because the overall phase requirements of the regulator can be met with one less DC stage (in most designs) than required in the regulating circuit employing a common-emitter series losser.

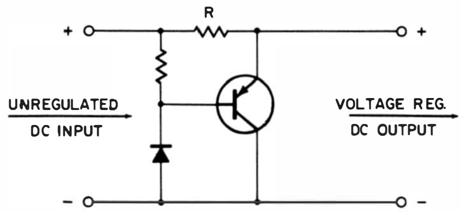
In Fig. 4-2 k is the nominal forward-bias voltage required by the comparator transistor (or tube) associated with the divider network R1 and R2.

Often the nature of the series losser is obscured in the complexities of the regulator schematic diagram. When one terminal of the voltage reference element connects directly to the emitter of the series losser, the losser operates as a common-emitter type. When one terminal of the voltage reference element connects to the collector or to the supply lead not interrupted by the series losser, the losser operates as an emitter follower. Moreover, the common emitter series losser incorporates zero, or an *even* number of polarity-inverting DC amplifiers (excluding the losser itself), whereas the emitter follower series losser achieves proper phase relationship with an odd number of such stages. (However, when complementary symmetry is involved, the odd or even criteria loses its validity.)

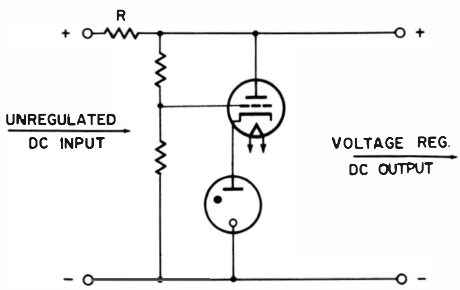
In most voltage regulators employing vacuum tubes as series lossers, the voltage requirements of all tubes are readily met with the "common collector" (cathode follower) configuration. For very high voltages, however, recourse is again made to common emitter mode of operation. With tube circuits the main precaution which ordinarily must be observed involves the insulation of the transformer filament windings. When the losser tube has a directly heated filament it is best to use a separate filament supply.

SHUNT VOLTAGE REGULATORS

Instead of being connected in series with one of the supply leads, the variable losser can be circuit-oriented to function as the *shunt* arm of a voltage divider. Two possible arrangements are depicted in Fig. 4-3. In the transistor circuit (Fig. 4-3A), any increase in output voltage will likewise increase the base-to-emitter bias in the negative direction. This increases the emitter-to-collector conduction of the transistor and thereby counteracts the increased output voltage. The tube circuit shown in Fig. 4-3B is not the exact counterpart



(A) Transistor circuit.



(B) Vacuum-tube circuit using VR tube.

Fig. 4-3. Shunt voltage regulators.

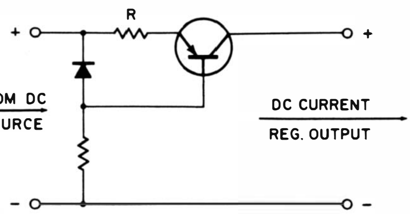
of the transistor circuit, because here the VR tube is located in the cathode lead. Although the tube is not directly across the output terminals, the operation is essentially similar to that of the transistor circuit. An increase in output voltage increases the grid-to-cathode voltage in the positive direction, thereby causing heavier cathode-to-plate conduction. In the transistor configuration (Fig. 4-3B), such action increases the voltage drop across series resistance R, enabling the output voltage to remain almost constant. In these circuits, the variable losser functions as a cathode (or emitter) follower, the input being essentially the voltage drop across the reference element.

An important operating difference between series and shunt losser circuits is the no-load or light-load efficiency. In the shunt type, the losser itself must assume the role of load in the absence of the external load. Thus, the efficiency of this type of regulator is very low when the current demand of the external load is not an appreciable fraction of the full-load current. Between half- and full-load, the efficiency is comparable to that obtainable from a similarly loaded series-type regulator. In transistor regulators, the shunt circuit has the advantage of being inherently short-circuitproof—a short across the output terminals merely removes the operating voltages. In contrast, a short circuit applied to a series regulator subjects the losser element to high voltage and high power dissipation. The voltage alone is often sufficient to destroy the series control transistor and, by a chain reaction, the smaller transistors if a DC amplifier is incorporated between the reference element and series control transistor.

CURRENT REGULATORS

Configurations of some simple current regulators are shown in Fig. 4-4. Like the simple series and shunt voltage

(A) Transistor as series element.



(B) Vacuum tube as series element.

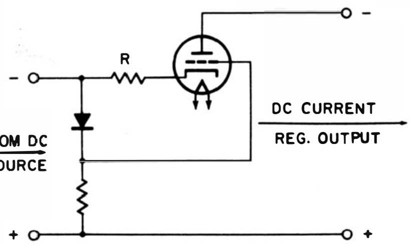


Fig. 4-4. Current regulators.

regulators described previously, these current regulators depend for their operation on a fixed reference voltage, feedback of an error signal, and power amplification in the losser element. Note that although current is the stabilized parameter, it is the *voltage drop* that is actually "sensed" and compared with the reference voltage. This voltage drop is developed across a small resistance, R. Being directly proportionate to the load current, this voltage drop can be said to faithfully represent the current flowing through the external load. Should the load current rise, the increased voltage drop across R constitutes bias of the proper polarity to counteract the current increased. The converse sequence of events occurs if the load current decreases. The nominal value of stabilized load current can be varied by changing the value of R or inserting voltage reference sources with different fixed voltages.

A somewhat different approach to current regulation is shown in the representative circuit of Fig. 4-5. Here the stabilizing action is not dependent upon a voltage reference element. The configuration comprises a two stage, direct-coupled amplifier with heavy negative feedback. The flow

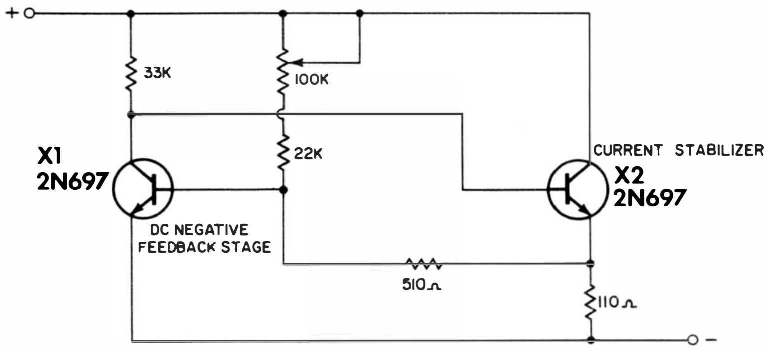


Fig. 4-5. Representative current regulator using DC negative feedback stage.

of current through current-stabilizer transistor X1 develops a voltage drop across the emitter resistance of that stage. This voltage drop is amplified and inverted by the AC negative feedback stage X2 and then fed back to the base of X1. The resulting closed loop action is control of X1 so as to oppose any change in its emitter-collector current. The result is that the applied voltage can vary over a considerable range with negligible change of current in the series connected load. The variable resistance is empirically adjusted to provide constant load current with respect to applied voltage. (By means of this adjustment it is possible to have the load current increase or decrease with respect to an increase in applied voltage.)

As shown in Fig. 4-5, the circuit will maintain a load current of approximately 5 milliamperes for applied voltages in the 5- to 25-volt range. Higher currents can be regulated by decreasing either or both the 110 ohm emitter resistance of stage X2 or the 33K collector resistance of stage X1. The plus and minus polarity designations are relative. The current regulator, voltage source, and load comprise a series circuit; their connection sequence is not important as long as the regulator "sees" the indicated relative polarity. A useful circuit feature of this type of current regulator is that it is effectively a *two* terminal circuit. Thus, it can be packaged and utilized as a constant current "diode" in an analogous manner to the constant voltage zener diode. PNP transistors can be substituted if the appropriate changes are made in terminal polarity. With larger transistors, particularly for X2, the circuit can be designed to regulate much heavier currents much more closely.

A family of high current germanium power transistors (PNP) are depicted in Table A-2 in the Appendix. These Honeywell units maintain a minimum current gain of 15 up to collector currents of 65 amperes. This is a very worthwhile characteristic, for it is often found that a power transistor begins to suffer appreciable loss in current gain long before its power dissipation rating is exceeded. When this happens, the transistor becomes less responsive to the amplified error signal and the performance of the regulator is degraded.

Commercial two-terminal current regulators are available. These are manufactured both from discrete components and from monolithic semiconductor material. One such device, known as the *Currector* is made by the Circuitdyne Corporation. A typical current-voltage relationship of

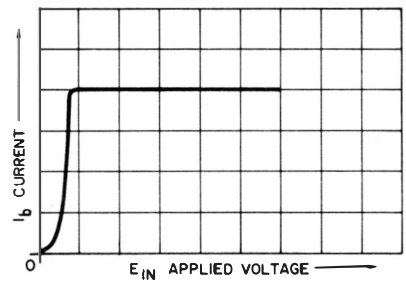


Fig. 4-6. The volt-ampere characteristic of the Currector.

a *Currector* is shown in Fig. 4-6. Such devices are very useful for providing constant current to differential amplifiers. When so used, the constant current device is connected in place of the mutual emitter resistance. A differential amplifier so connected will perform well from a relatively low-voltage supply. Superior common mode rejection and improved temperature stability are obtained compared to operation with the emitter resistance. The basic idea is depicted in the representative circuit of Fig. 4-7.

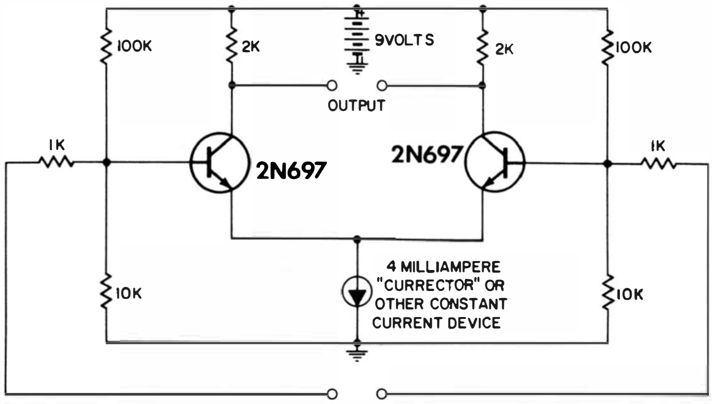


Fig. 4-7. Representative differential amplifier with constant current source in place of resistance.

Two-terminal current regulating devices can be used to upgrade the performance of voltage-regulated power supplies. In the simplest of such applications, the current regulating device substitutes for the series dropping resistor in an open-loop zener diode supply. In Fig. 4-8 the voltage

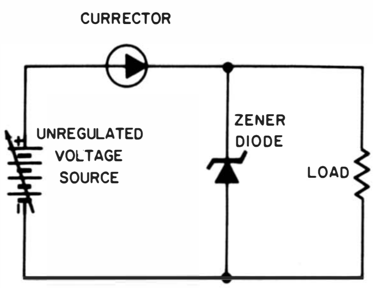


Fig. 4-8. Replacement of series loss resistance by Currector element in zener-regulated constant-voltage supply.

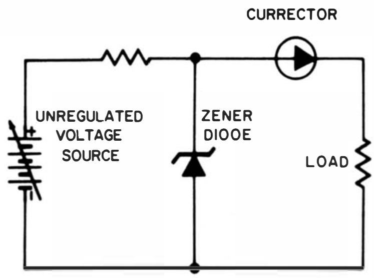
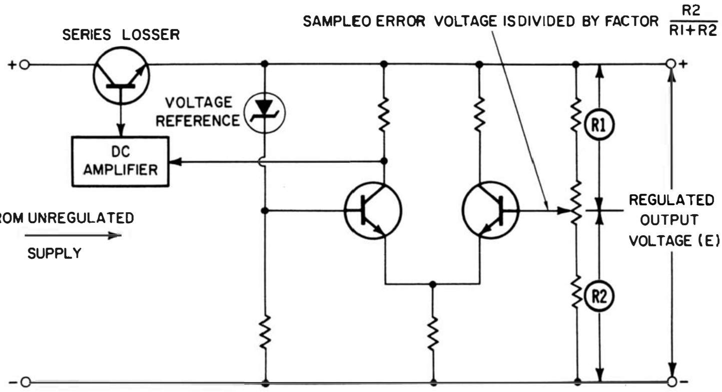


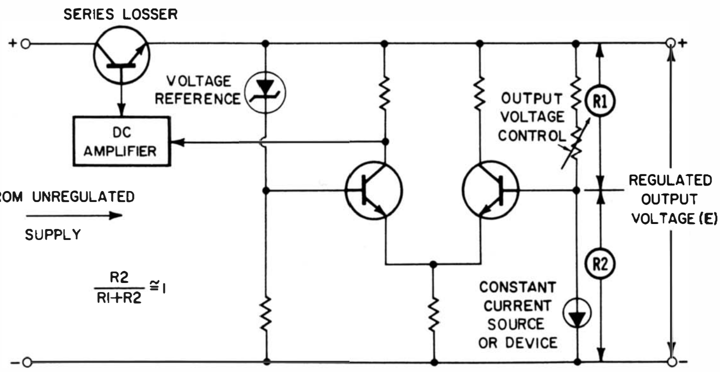
Fig. 4-9. Improvement of current regulation by means of zener diode preceding the Currector element.

regulation seen by the load is better than would be obtained with the conventional series dropping resistor. This is particularly true of load voltage with respect to variations in the unregulated supply. Probably the best use of this current regulator-zener diode combination is for achieving a more stable voltage reference in closed loop regulated supplies.

The converse situation is shown in Fig. 4-9. Here, enhanced current regulation ensues from the constant-voltage



(A) Conventional circuit.



(B) Circuit with constant current source.

Fig. 4-10. Simplified sampler/comparator circuits showing advantage of constant current source in comparator section.

property of the preceding zener diode. Other useful combinations of constant-current "diodes" and constant-voltage diodes will, no doubt, suggest themselves to the imaginative experimenter.

In conventional closed-loop regulated supplies, the resistance voltage divider network provides a desired and an *undesired* effect. The potentiometer action of this divider permits an appropriate sample of the output voltage to be impressed at the comparator/DC amplifier for comparison with the stable reference voltage. However, this network *also* divides the available loop amplification. This situation can be circumvented by the arrangement shown in Fig. 4-10 where a two-terminal, constant-current source substitutes for the resistive arm of the sampling network which conventionally is connected to the unsensed side of the supply.

Even though a resistance, R1 is used for control of output voltage, the sampled voltage is always substantially unattenuated. The constant current source is the equivalent of an extremely high resistance insofar as concerns potentiometer action of the sampling circuit. Nearly 100 percent negative feedback exists in the regulating circuit as a consequence of this. Another way of stating this is that virtually no amplification is wasted when such an arrangement is used. This is reflected as a lower, dynamic output impedance, generally the most important performance parameter of voltage-regulated power supplies. To derive optimum results with this technique, the emitter-base current of the sensing transistor should be a very small part of the stabilized current provided by the constant current source.

High performance voltage-regulated supplies usually have some provision for operating the collector of the last DC voltage amplifier from a constant-current source. This prevents injection of ripple into the base of the series losser or its Darlington connected driving stage. The most common constant-current circuit employs a "three" terminal transistor/zener-diode arrangement such as shown in Fig. 4-4A. In some cases, the pentode-like characteristic of a grounded-base transistor provides constant-current behavior for this purpose. (With a transistor so-used, no zener diode is needed). It is probable that a two-terminal constant current source would serve this function equally well as the three terminal circuits.

ERROR-SIGNAL AMPLIFICATION

Up to now we have investigated several simple configurations for regulating the voltage or current of load powers beyond the practical or economical range of zener diodes or VR tubes. The power amplification contributed by the variable losser element improves the dynamic characteristics of the supply in all these circuits. Further improvement is readily attainable by increasing the power amplification in the feedback loop. This is done by inserting voltage amplifiers—and often in transistor regulators, current amplifiers—between the voltage reference and variable losser. With such a scheme, the control function becomes sensitive to very small changes in output voltage or current. Other advantages are also obtained—both design and operational flexibility prosper from such an arrangement. For example, the output voltage can readily be made controllable and independent of the fixed reference voltage. Thus, a 6-volt zener diode can stabilize output voltages from zero to several hundred volts, any voltage within such a range being selected by a potentiometer. Contrary to the situation which prevails in simple supplies (where a potentiometer is provided for output control), the dynamic output impedance and output resistance of the regulator-amplifier combination can remain extremely low over the entire range of adjustable voltages.

DC-AMPLIFIER PROBLEMS

Although the insertion of amplification between the sampled portion of the output voltage (or current) and the variable losser enhances the performance of the power supply, this scheme is not without its problems. One of the undesirable characteristics of DC direct-coupled amplifiers is their tendency to drift. A slight shift in operating point in the input stage is indistinguishable from an error signal. Such a shift will therefore change the output of the power supply, this change being unrelated to the stabilization. Under certain conditions, drift can become cumulative and result in a runaway condition. In any event, the presence of appreciable drift causes erratic hunting of the output level. This problem generally does not impair the performance of capacitor- or transformer-coupled AC amplifiers. For this reason, when precise stabilization is required, the error signal sometimes is interrupted by an electronic or electromechanical chopper and the resultant AC signal amplified by an AC rather than DC amplifier. This type of regulated power supply will be dealt with in more detail later. Our present concern is with techniques for minimizing drift in a DC amplifier.

You will find that the inherent drift problems generally begin to assert themselves whenever two or more DC amplifier stages are cascaded. In such instances the drift in the input stage inflicts the most serious consequences, because it receives *more* amplification than the drift occurring in a second or third stage. For this reason, a great deal of attention has been given to the design and operation of the input stage. Transistors are particularly susceptible to drift from temperature changes. A very effective circuit for minimizing the input-stage drift of transistors and vacuum tubes is the differential amplifier in Fig. 4-11. Note that the two amplifying elements are physically close together—that is, in the *same thermal environment.* Thus, a change in the common filament temperature will not change the plate current and therefore the output-voltage level. Suppose the filament temperature is decreased, with the attendant tendency of the

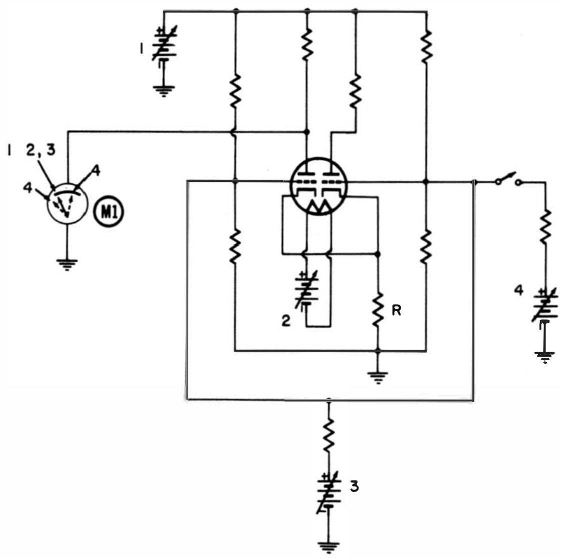
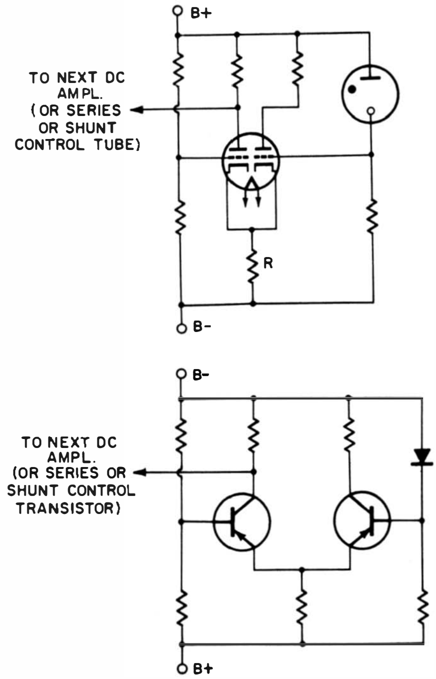


Fig. 4-11. Effect of operating voltage variations on output of differential amplifier.

plate current in both tube sections to decrease. Because of the common cathode resistance, R, each tube will control the cathode-to-grid bias of the alternate tube. Any tendency of *both* tube sections to decrease their plate currents by the same amount will result in both cathodes being simultaneously driven *less positive* than their grids. In turn, conduction will increase in both tube sections. Because of the amplification in the tubes, the increased conduction (due to the cathode-to-grid bias) very nearly offsets the decrease in conduction due to the lowered filament temperature. Therefore, the conduction of the tube sections remains fixed, and their output does not change. Here we see a regulatory action in which equal changes in the amplifying elements are canceled. Significantly, in a properly designed differential amplifier, most drift tends to take place within both amplifying elements.

Let's look at other "common mode" changes. The same signal, when applied simultaneously to *both* grids, is similarly prevented from changing the output-voltage level at the plates. The same is true for B+ variations. In summary, variations in voltages from sources 1, 2, and 3 in Fig. 4-11 produce no appreciable change in output voltage as seen on meter M1. A change in output voltage does occur, however, if the bias on *either* one or the other (but not both) grids is changed. Now, in Fig. 4-12, B+ variations *will* change the output voltage because *one* of the control elements is connected to a voltage reference element. Such an arrangement does not result in both grids or both bases receiving a common-mode signal. An output signal then is developed which is the amplified difference between the voltage levels applied to the two grids or bases. Such operation is desired when the differential amplifier is used as an error-signal amplifier in the feedback loop of a regulated power supply. Fig. 4-12A



(A) Vacuum-tube circuit.

(B) Transistor circuit.

Fig. 4-12. Typical examples of differential amplifiers with input voltage reference connected to one input.

illustrates the vacuum-tube circuit and Fig. 4-12B the equivalent transistor configuration.

Transistors present more of a problem in the differential amplifier than vacuum tubes. This is primarily due to the matching and tracking problems inherent in transistors. In the first place, transistors bearing the same type designations often differ from one another more than two vacuum tubes of different type designations. Even if a pair of transistors are matched at room temperature, some of their characteristics can diverge considerably as the temperature rises. This is particularly true of the collector saturation current;

it is generally far better to use transistors with very *low* collector saturation currents than to attempt to match two transistors having appreciable collector saturation currents at room temperature. For this reason, silicon transistors are preferable to germanium transistors for the differential amplifier. Additionally, the two transistors should have as nearly identical current gains and base-to-emitter voltage characteristics as possible. Both transistors should be mounted in a common heat sink—despite the fact that dissipation will normally be quite low—because the heat sink ensures an identical thermal environment for both transistors.

The Fairchild transistors shown in Table A-4 and A-5 in the Appendix may be considered as "work-horse" types for regulator circuits. These transistors are of silicon planar construction and possess excellent characteristics. Note that they are available in both NPN and PNP polarities. The author particularly recommends the 2N966 and 2N967 NPN transistors and the 2N1131 and 2N1132 PNP transistors. These transistors can be directly substituted for a great number of germanium transistors specified for such circuit functions as comparator, differential amplifier, voltage amplifier, and Darlington driver.

Complete differential amplifiers within a single package are now available. These have the important feature that both elements are exposed to very nearly the same thermal environment. A list of such differential amplifiers made by Fairchild appears in Table A-6 in the Appendix. It will be observed that differential amplifiers corresponding to both NPN and PNP transistors are available.

Somewhat similarly, Darlington amplifiers are packaged within one container. This gives the effect of a single transistor with extremely high current gain. Three Fairchild units are shown in Table A-7 in the Appendix.

Epoxy encapsulated silicon planar transistors are available which, in many instances, provide all the attributes of silicon transistors but at considerably lower price than conventional TO-5 and TO-18 packaging. These transistors are "passivated" by means of a silicon-oxide surface film, so that they are not primarily dependent upon the epoxy seal for protection from moisture and atmospheric gasses. Some of these types have exceedingly high current gain (beta) ratings. Epoxy has good thermal conductivity and for a few cents additional outlay, a simple clip-on heat sink will provide increased power dissipation capability. Table A-8 lists a series of General Electric NPN types. Fairchild makes both NPN and PNP types with epoxy encapsulation; these are shown in Tables A-9 and A-10 in the Appendix.

Table A-11 (in Appendix) lists a series of Fairchild temperature compensated reference diodes. These have very low temperature coefficients. By their use, great stability of regulated output voltage or current is readily obtained with respect to temperature.

Although elaborate precautions are prescribed for balancing the two amplifiers in the differential circuit, the two plate-load resistors need not be equal in resistance, because the plate or collector current is determined primarily by the bias at the cathodes or emitters. Consequently, we often see unequal plate or collector resistances, and sometimes one of them is omitted.

OSCILLATION SUPPRESSION

The DC amplifier is an AC amplifier as well. This is desirable, for it enables the over-all regulator circuit to provide electronic filtering of residual ripple from the unregulated power supply. However, the effect of the amplification in lowering the dynamic output impedance cannot extend to higher frequencies than those to which the amplifier will provide appreciable gain. Inasmuch as high amplification in the error-signal feedback loop is a relatively cheap way to enhance the performance of a regulated supply, it is commonplace to provide fairly high gain in all but the most rudimentary regulators. This, however, introduces the possibility of oscillation. Although the feedback is negative at DC and low frequencies, cumulative phase shifts in the amplifying elements generally are such that the criterion of oscillation is met at some sufficiently high frequency. When such oscillation occurs, the AC generated can exceed the permissible voltage swing at the input of one or more amplifying elements. As a result, rectification takes place, with the attendant shift in operating point of the amplifier stages. This amounts to a condition of enforced drift that incapacitates the regulator. In transistor amplifiers, *destruction* of transistors is not an uncommon occurrence. Even if the oscillation is weak, it will still impair the regulator performance by appearing across the output terminals. In the interest of proper regulator operation, it is most important that any oscillation or tendency toward oscillation be suppressed.

Oscillation can be either caused or inhibited by actual, stray, or equivalent reactances in certain portions of the amplifier circuit. The most straightforward remedy is to connect capacitors at the appropriate junctions, to prevent phase-gain conditions for oscillation. Such capacitors will be observed in the ensuing circuits of regulated power supplies. Where feasible, they are connected to high-impedance points; this permits the use of a comparatively small capacitor. Sometimes it is necessary to connect a large electrolytic capacitor directly across the output terminals in order to discourage oscillation at all frequencies.

Oscillation problems are generally agitated in voltage-regulated supplies designed to provide a considerable range of control over the output voltage. The reason is that zero output voltage is approached as the negative feedback is increased (by adjusting the potentiometer in the output-voltage sampling network). The potentiometer is adjusted in the direction which makes the sampled voltage a larger fraction of the output voltage than would be true for higher output voltages. It is well known, from audio-amplifier techniques, that a practical limit is reached in the amount of negative feedback which can be incorporated. This practical limit corresponds to the occurrence of oscillation at those frequencies where gain and phase conditions actually produce positive feedback, thus converting the amplifier to an oscillator. Other things being equal, such conditions tend to manifest themselves as the amplifier gain and feedback factor are increased. In the voltage-regulated supply, the feedback factor increases as the supply is adjusted toward a lower output voltage.

CHAPTER 5

Typical Closed-Loop Regulated Supplies

OPERATION AND DESIGN

A 250-Volt, 200-MA Vacuum-Tube Voltage-Regulated Supply

In Fig. 5-1 we see the schematic diagram of a vacuum-tube voltage-regulated power supply, which is very useful for numerous audio, amateur, and experimental projects. Note that the double-triode input amplifier is a cascaded two-stage DC type rather than a differential amplifier. The added amplification thus obtained produces a low dynamic output impedance, which is considered more desirable for the intended applications than extreme constancy in the output voltage. Sufficient constancy is brought about by operating VR tube V4 from the voltage which the tube helps stabilize. The price paid for this, however, is that the output voltage cannot be continuously adjusted down to zero volts by adjustment of R13, the 10K potentiometer. If carried too far, the operating voltage across the VR tube will disappear. Also, the two-stage voltage amplifier will be deprived of its proper operating voltage before the output voltage can be reduced to zero. The 12K fixed resistance (R12) prevents lowering the output voltage below about 125 volts.

Capacitor C3 (0.25 mfd) in the cathode circuit of the second DC amplifier stage provides AC bypass action in order to prevent AC degeneration due to R7. It is desirable to have high AC amplification in order that the voltage amplifier can effectively filter out the residual ripple in the output. Capacitor C2 modifies the phase characteristic of the amplifier and thereby discourages oscillation. For this connection, capacitances between .01 and 0.1 mfd generally work well. The series losser element (V2) is a type 6080 dual triode. The two triode sections are connected in parallel through current-

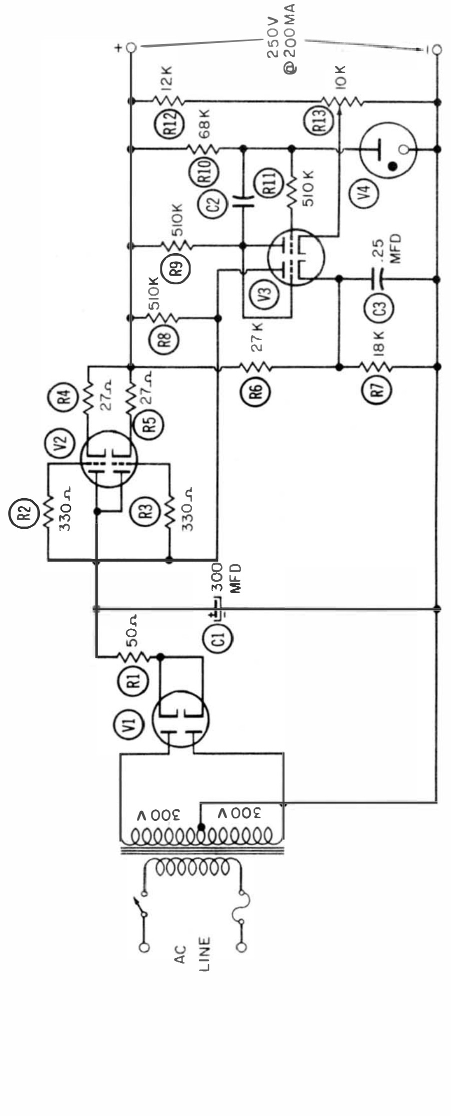


Fig. 5-1. A vacuum-tube voltage-regulated supply rated at 250 volts, 200 ma.

equalizing resistors R4 and R5 and operate essentially as a single cathode follower.

In this power supply, the grid of the input amplifier is held at a fixed positive voltage with respect to the negative line. Hence, if AC line or load conditions change in such a direction that the DC output voltage tends to increase, the cathode of the input amplifier will become more positive than the fixed potential at the grid. Such a change in grid-to-cathode bias will decrease the plate current in the input tube. The later in turn will increase plate conduction in the second DC amplifier (the left half of V3). The grids of the 6080 control tube (V2) will be driven in the negative direction by the increased plate current in the second amplifier stage. The final result is lower conduction in the control tube—which is tantamount to saying that the change in DC resistance of the 6080 tube nearly restores the DC output voltage to its previous value. The converse sequence of events occurs if the output voltage tends to drop. In this way, the output voltage is stabilized at a discrete value controlled by the adjustment of the 10K potentiometer.

Transistorized Shunt-Voltage Regulator

Fig. 5-2 is an example of a closed-loop voltage regulator with the variable losser connected *across* the load rather than in series with it. If the output voltage tends to increase, the losser element (transistor X5) will receive an error signal which increases the conduction through its emitter-to-collector circuit. This increases the voltage drop across R11, thereby depriving the load of the voltage change responsible for the error signal. The converse conditions apply when the output voltage tends to decrease. In this way, the voltage across the load is maintained at a fixed value. Although a closed loop is involved, the ultimate action is similar to that in a simple VR-tube or zener-diode circuit. Note that the effective resistance of the shunt element must change in the *opposite* direction from that of a series-connected losser in order to accomplish like results.

An interesting feature of the circuit in Fig. 5-2 is that voltage reference M1, differential-amplifier stage X1-X2, and voltage-amplifier stage X3 all have individually regulated power sources, which contribute toward stability and also enable control of the regulated voltage output down almost to zero volts. This design is well suited for developmental work with transistor circuitry. Short-circuited output terminals do not damage or overload any elements. The parts data for this circuit are provided in Table 5-1.

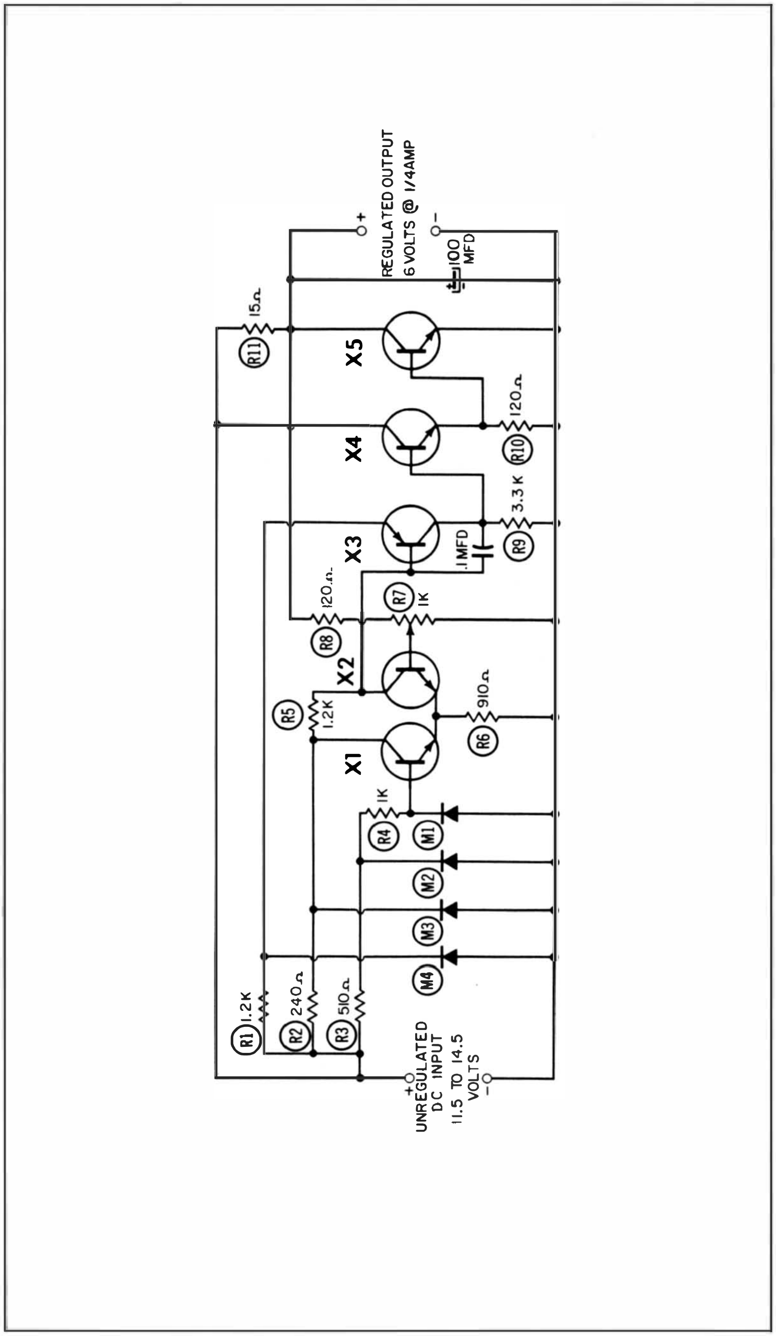


Fig. 5-2. A low-voltage shunt regulator with an output of 6 volts at 0.25 ampere.

Table 5-1. Transistor and semiconductor diode data for circuit of Fig. 5-2

Component	Description	Circuit Function	Type
X1, X2	NPN silicon	Differential amplifier-comparator and voltage amplifier	2N338
X3	PNP germanium	2nd stage of DC voltage amplification	2N1374
X4	NPN silicon	Emitter-follower driver stage	2N343
X5	NPN silicon	Variable-resistance shunt control element	2N1722
M1	750-MW zener diode	Voltage reference for differential amplifier	MZ5.6T5
M2	750-MW zener diode	Voltage regulator for M1	MZ8.2T5
M3	750-MW zener diode	Voltage regulator for differential amplifier	MZ10T5
M4	750-MW zener diode	Voltage regulator for emitter bias of X3	MZ8.2T5

A High-Voltage Regulated Supply

The voltage-regulated supply shown in Fig. 5-3 is capable of providing up to several milliamperes to the collector of a traveling-wave tube, with voltage regulation on the order of 0.1% (line or load). Output voltage is adjustable from 1,600 to 2,100 volts. Additionally, VR tubes provide a selection of less precisely regulated voltages for the anode of the traveling-wave tube. Tubes V8, V9, V6, and V7 constitute the voltage reference, comparator and DC amplifier, and series losser for regulation of the high voltage provided by rectifiers V10 and V11.

The circuit as described thus far would be operative and would be typical of many vacuum-tube voltage regulators. However, in order to obtain closer regulation and higher stability, the voltage applied to the VR tubes is also regulated. This function is provided by tubes V2, V3, and V4, which stabilize the voltage delivered by full-wave rectifier V1. It is interesting to note that VR tubes V8 and V9 belong as much to the regulator circuit associated with V2, V3, and V4 as to the regulator circuit associated with V5, V6, and V7. V4, a dual triode connected as a differential amplifier, is the comparator and first DC amplifier stage of the voltage regulator for the VR tubes. V3 is the second stage and has a similar circuit configuration. V2 is the controllable series losser.

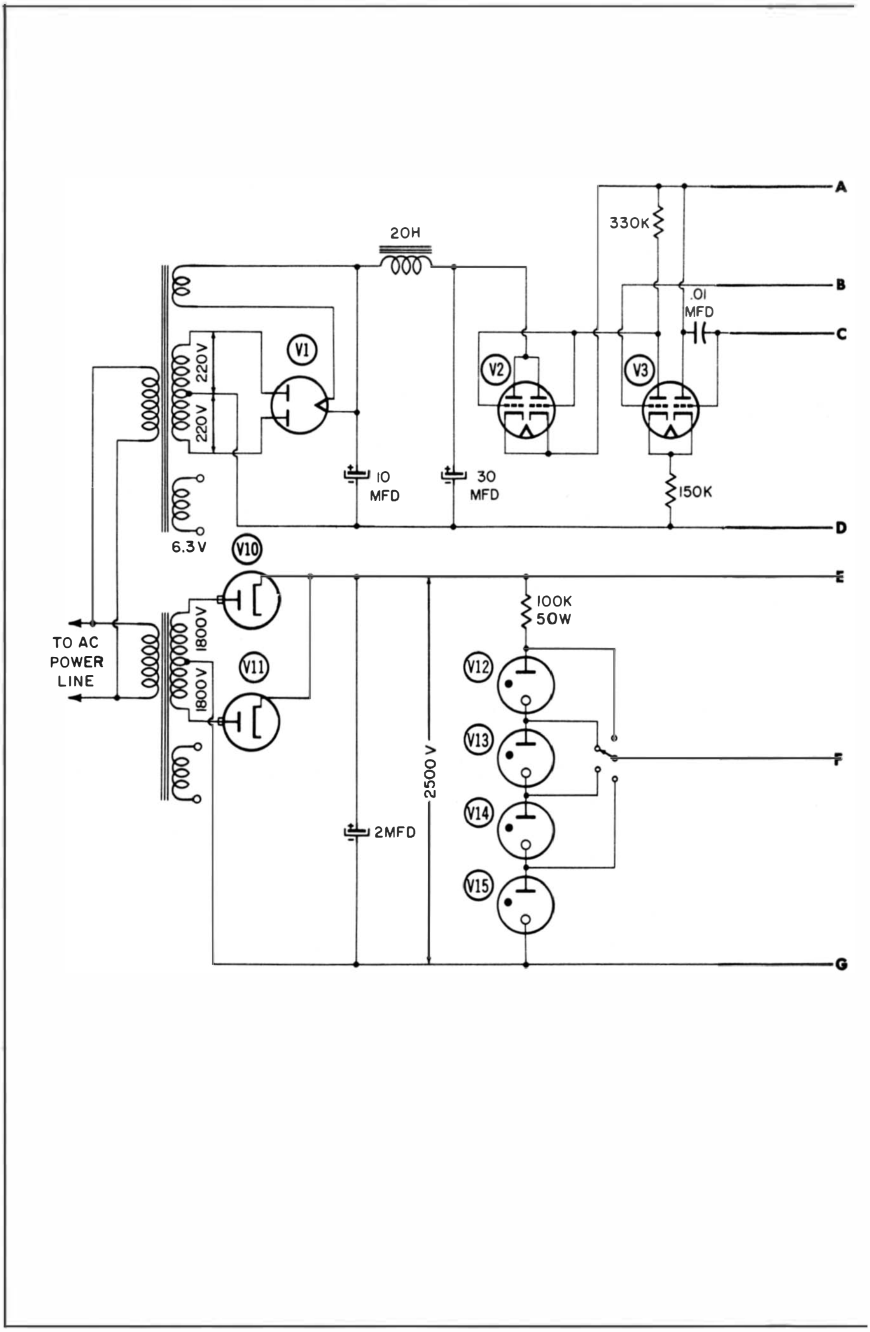
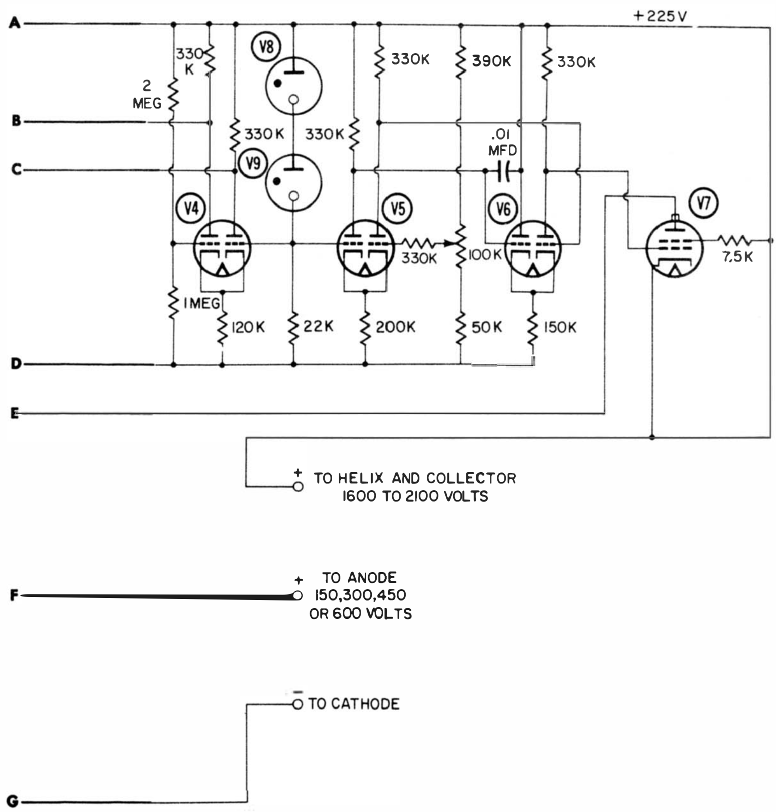


Fig. 5-3. A high-voltage regulated power supply for



traveling-wave tubes.

Table 5-2 gives the tube types and their circuit functions for the power supply in Fig. 5-3.

A 2.5-Ampere Voltage-Regulated Power Supply

Despite the extremely simple circuit configuration of the voltage-regulated power supply shown in Fig. 5-4, the performance is more than adequate for the requirements of many high-current loads. It would be both impractical and difficult to obtain high current and low voltage from tube circuits; yet this type of output is a natural for transistors. An unusual aspect of this supply is that there is no voltage amplification of the error signal. Transistors X1 and X2 are both emitter followers and provide current amplification only. In essence, they improve the impedance match between reference diode M5 and the base of series losser X3. This enables X3 to develop optimum power sensitivity as a variable series control element in this circuit. Consequently, the regulation characteristics are much better than they would be without X1 and X2. Although voltage amplification is generally provided in the error-signal amplifier of regulator

Table 5-2. Vacuum-tube data for the circuit in Fig. 5-3

Tube	Description	Circuit Function	Type
V1	Full-wave rectifier	Provides unregulated DC for DC voltage amplifiers	5Y3
V2	Dual triode	Series control tube for B+ requirements of DC voltage amplifiers	12AU7
V3	Dual triode	Differential amplifier (2nd stage) for regulation of voltage to V7 and V8	12AX7
V4	Dual triode	Differential amplifier (1st stage) for regulation of voltage to V7 and V8	12AX7
V5	Dual triode	Differential amplifier (1st stage) for regulation of high voltage	12AX7
V6	Dual triode	Differential amplifier (2nd stage) for regulation of high voltage	12AX7
V7	High-voltage Tetrode	Series control tube for regulation of high voltage DC	6146
V8, V9	High-stability VR tubes	Provide reference voltage for high voltage regulated supply (V5, V6, V7)	5651
V10, V11	High-voltage half-wave rectifiers	Develop 2500 volts DC across 2-mfd capacitor	2X2A
V12, V13, V14, V15	VR tubes	Provide regulated voltages at 150, 300, 450 and 600 volts	OA2

circuits, it is always the over-all *power amplification* which enables closer regulation to be achieved. In tube-regulator circuits, the power amplification is usually obtained by *voltage gain* alone, because the losser tube seldom operates deeply in its positive-grid (current-consuming) region. With transistor losser elements, input current is consumed; and even when the error signal is amplified considerably, at least one emitter-follower stage is required in order to satisfy the current demand of the losser transistor.

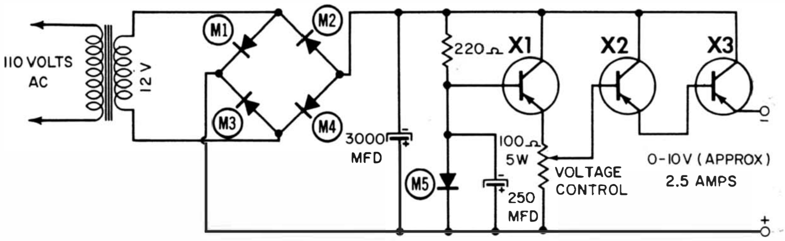


Fig. 5-4. A series regulator with a continuously variable output voltage from 0 to 10 volts and a current rating of 2.5 amperes.

This voltage-regulated power supply is well suited for energizing the DC control winding of magnetic amplifiers. As a source of filament power, it will prove beneficial in minimizing hum problems in audio equipment and delicate instrumentation techniques. Table 5-3 gives the transistor and diode types used in this circuit, and their function.

Table 5-3. Transistor and semiconductor diode data for the circuit of Fig. 5-4

Component	Description	Circuit Function	Type
M1, M2, M3, M4	Silicon rectifiers	Provide full-wave bridge rectification	1.5 or 2 AMP Units
M5	Zener diode	Voltage reference for 1st DC stage, X1	MZ9.1T5
X1	GE power transistor	1st DC amplifier stage and comparator	2N554
X2	GE power transistor	2nd DC amplifier stage	2N554
X3	GE power transistor	Series control element	2N554

A 150-Volt, 0.5-Ampere Solid-State Voltage-Regulated Power Supply

The fully transistorized voltage regulator shown in Fig. 5-5 is unique in that the voltage output is considerably higher than the several tens of volts often obtained from solid-state

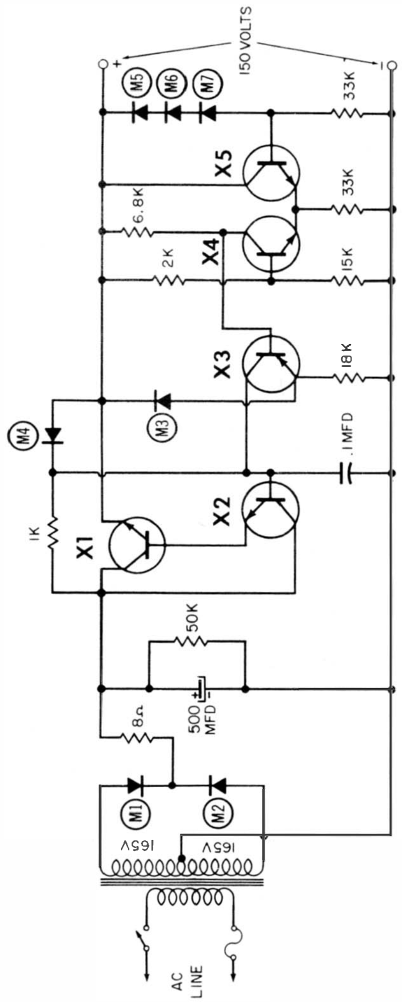


Fig. 5-5. A solid-state voltage-regulated supply that will deliver 150 volts at 500 ma.

regulators. Differential amplifier X5-X4 is powered from the regulated output. X3 is a voltage-amplifier stage whose operation is stabilized by zener diode M3 connected to its emitter circuit, and also by zener diode M4 in its collector circuit. M4 additionally prevents introduction of hum into the base of X2 through the 1K biasing resistor. X2 is an emitter-follower stage which provides a reasonable impedance match between the collector of voltage amplifier X3 and the base of series losser element X1. The presence of X2 enables X3 to develop a much higher voltage gain than would otherwise be the case.

Although 150 volts appears at the output terminals, the emitter-to-collector voltage of series losser X1 is in the

Table 5-4. Transistor and semiconductor diode data for the circuit of Fig. 5-5

Component	Description	Circuit Function	Type
M1, M2		Rectifier for converting incoming AC to DC	
M3	400-MW, 10V zener diode		1N758
M4	400-MW, 6.2V zener diode		1N753
M5, M6, M7	400-MW, 7.5V zener diode	Voltage reference for differential amplifier X5	1N755
X1		Series control element	2N389
X2		Emitter-follower current amplifier	2N1047
X3		DC voltage amplifier	2N497
X4, X5		Differential amplifier and comparator	2N338

neighborhood of one-tenth this value. In the event of a short circuit at the load, the full output voltage from the rectifier will then be impressed across X1 and destroy it. Hence, extreme precaution must be taken to prevent such an occurrence. Obviously, this supply is not suited for experimental work. Other transistor-regulated supplies, which operate at lower voltages and generally at higher currents, can survive momentary short circuits due to the thermal lag of the larger transistors used in the series control function. With such supplies, a fuse or a fast circuit breaker can provide additional protection. This power supply could be made short-circuitproof by additional electronic circuitry, which would instantaneously shunt the emitter of X1 to its collector. Providing cutoff bias to the base of X1 (a popular short-circuit protection in lower-voltage supplies) would not be

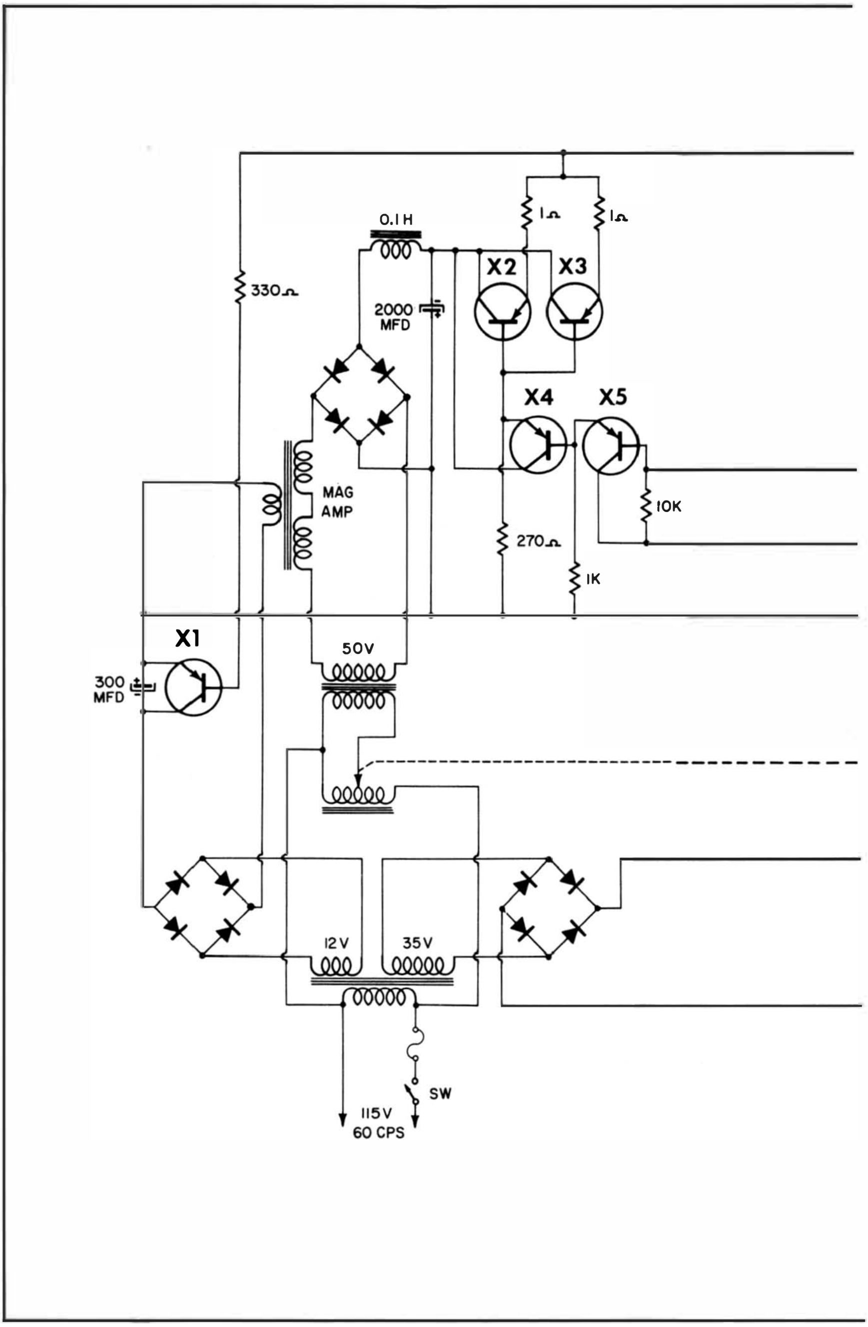
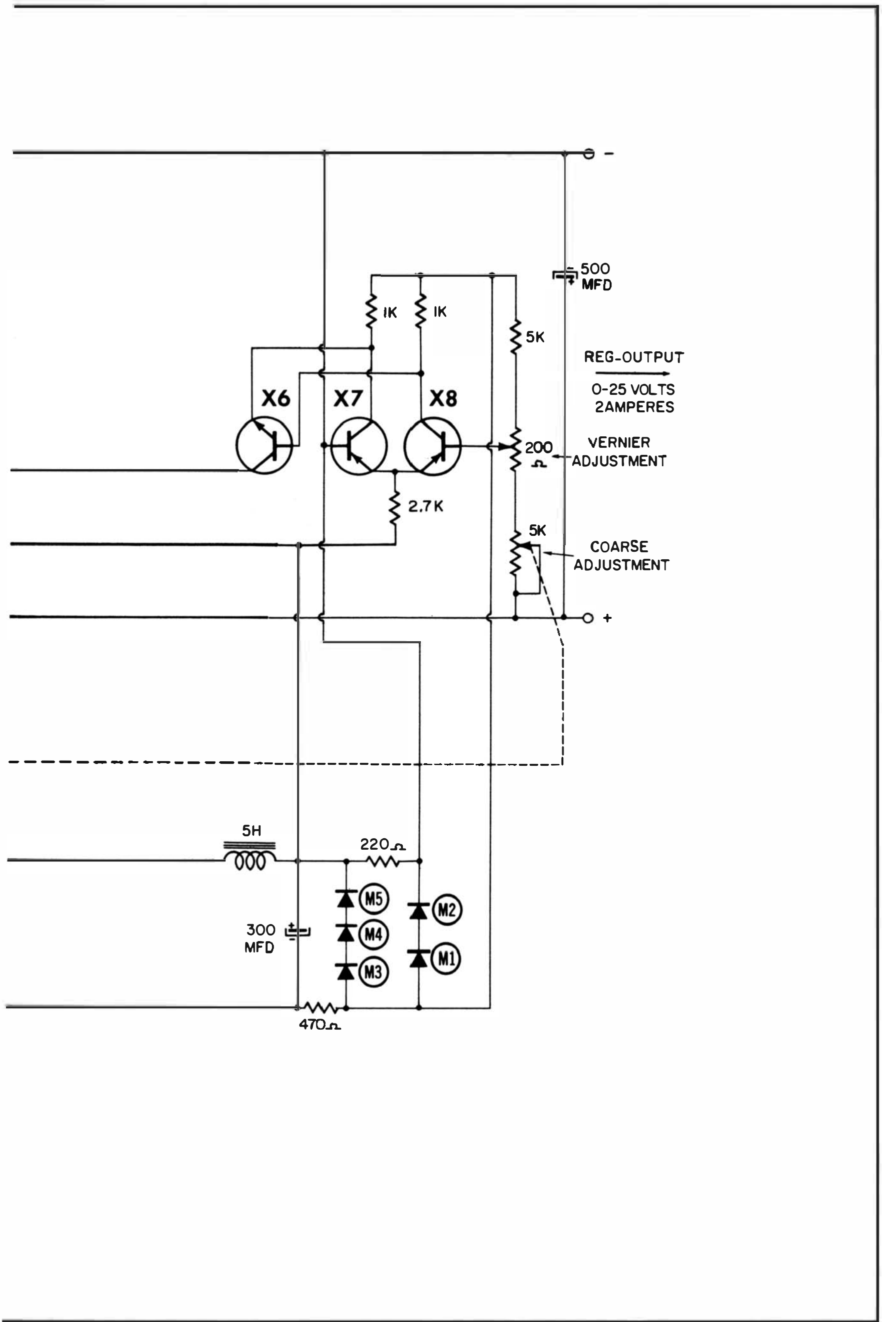


Fig. 5-6. A transistorized series regulator with



output adjustable from 0 to 25 volts at 2 amperes.

suitable here because X1 must be protected from excessive voltage as well as excessive current. (Refer to Table 5-4 for transistor and semiconductor diode data for this circuit.)

Transistor Voltage-Regulated Supply with Overcurrent Protection

The voltage-regulated supply shown in Fig. 5-6 is typical of many transistor regulator circuits, but it also involves circuitry which greatly enhances its practicability. Note that the AC windings of the magnetic amplifier are inserted in series with one AC line of the power-transformer secondary. The control winding of this magnetic amplifier receives DC current through transistor X1 from a separate 12-volt DC power supply. The base of X1 is impressed with conduction bias from the voltage appearing at the output terminals of the regulated supply. As a result, X1 is normally biased for high emitter-to-collector conduction, thus permitting a heavy current to flow into the control winding of the magnetic amplifier. The core of the magnetic amplifier is magnetically saturated, and the AC windings have very little inductance. Therefore, nearly all the AC voltage developed in the secondary of the power transformer is impressed across the main rectifier bridge, and normal power-supply operation occurs.

In the event of a short circuit at the output terminals, X1 will be deprived of its forward-conduction bias. The control winding of the magnetic amplifier will then receive little or no DC magnetization current. This, in turn, results in a high inductance in the AC winding, because the flux in the core is now considerably below the densities corresponding to magnetic saturation. The high inductance limits the voltage output of the power supply, thereby protecting X2 and X3 from excessive power dissipation. Inasmuch as X2 and X3 are large power transistors and are heat-sinked to aluminum or copper plates, the response lag inherent in magnetic amplifiers is of little practical consequence, and reliable protection is obtained.

Another feature of this regulator is the circuit provision for continuous adjustment of the regulated output voltage from 0 to 25 volts. This is done by the more or less conventional method of using a variable resistance at the input circuit of the differential amplifier. However, it is ordinarily difficult to reduce the output voltage beyond a nominal amount because the series losser transistor(s) then must operate with a relatively high collector-to-emitter voltage. It is evident that the losser transistor must absorb the difference between the voltage from the rectifier system and the voltage across the load. If the load voltage is lowered too much, the losser transistor will be subjected to a relatively high voltage. Although this transistor may have an adequate voltage rating, it will also have to dissipate inordinately *high power* when the load consumes the rated current. It becomes impractical and uneconomical to use a losser transistor having the power-handling ability to withstand such service. This problem can be circumvented by inserting a variable autotransformer in the primary circuit of the main power transformer and mechanically ganging it to the voltage control resistance (potentiometer). As a result, the voltage output of the main rectifier system will automatically

Table 5-5. Transistor and semiconductor diode data for the circuit of Fig. 5-6

Component	Description	Circuit Function	Type
M1, M2	750-MW zener diode	Reference voltage for differential amplifier	MZ 6.8T10
M3, M4, M5	750-MW zener diode	Regulated source for operation of M1 and M2	MZ 6.8T10
X1	PNP germanium	DC current control of mag-amp for short-circuit protection of power supply	2N441
X2, X3	PNP germanium	Variable resistance series control elements	2N442
X4	PNP germanium	Emitter-follower to drive control elements	2N307
X5	PNP germanium	Emitter-follower for impedance transformation	2N1375
X6	NPN silicon	2nd-stage voltage amplifier	2N338
X7, X8	PNP germanium	Differential amplifier-comparator and voltage amplifier.	2N1375

be reduced as the load voltage is lowered. Therefore, the voltage drop across losser transistors X2 and X3 will be maintained at a low, almost constant value.

In order that the load voltage may be adjusted down to zero, separate regulated power supplies are provided for amplifiers X8, X7, X6, and X5. Transistor X4, the driver for X2 and X3, operates from the unregulated supply line and is thereby ensured an operating voltage for any adjustment of the regulated output voltage. Table 5-5 gives the type number and function of the transistors and zener diodes used in Fig. 5-6.

A 1.5-Ampere Current-Regulated Power Supply

In the current-regulated power supply, the input amplifier compares a reference voltage with a sampled voltage. The latter is proportionate to the current delivered by the supply, rather than to the voltage across the output terminals (as in voltage-regulated supplies). Thus, in Fig. 5-7 the 10-ohm series resistance provides a voltage drop that is directly proportionate to the current flowing through it. Should the load current increase, X3 will be deprived of sufficient forward bias to lower the load current to its nominal regulated value. The converse action occurs if the load current should tend to decrease. An error signal of appropriate sign and amplitude appears at the base of X1 to restore the output current to its nominal regulated value whenever the line voltage or load resistance changes. The error signal then

Table 5-6. Transistor and semiconductor diode data for the circuit of Fig. 5-7

Component	Description	Circuit Function	Type
M1, M2, M3, M4	Silicon rectifiers	Bridge-rectifier arrangement	1N1084
M5	Zener diode	Voltage reference	1N752
X1	Germanium PNP transistor	1st-stage DC amplifier and comparator	2N1379
X2	Germanium PNP transistor	2nd-stage DC amplifier	2N1379
X3	Germanium power transistor	Series control element	2N174

restores the load current to its regulated value. This value may, however, be adjusted over a wide range by means of the 75-ohm potentiometer.

The 50-ohm resistance connected across losser transistor X3 protects the latter from the voltage surges produced when inductive loads are switched (not recommended). This resistance is not low enough to seriously impair the current regulation. A large capacitor in this circuit position is also helpful here. X2, which is commonly an emitter follower in voltage-regulated supplies, is connected in the common-emitter configuration. Although little voltage amplification is obtained from this stage, it does provide the necessary phase inversion so that the effective resistance of X3 will change in the proper direction and thereby regulate the load

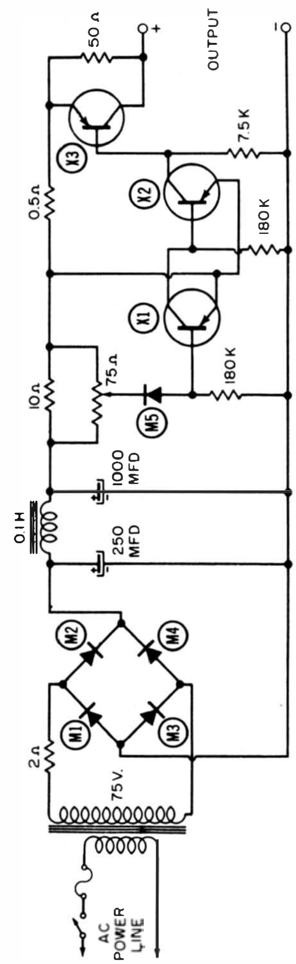


Fig. 5-7. A 1.5-ampere current-regulated power supply.

current. Table 5-6 gives the transistor and semiconductor diode data for Fig. 5-7.

REMOTE SENSING OF THE LOAD

No matter how closely a supply is voltage regulated, the wiring between the output terminals and load will always tend to increase the output resistance and dynamic impedance. This can be avoided by freeing the ends of the voltage-sampling network from their connections to the positive and negative output terminals. Separate connections are then made between the voltage-sampling network and load terminals. A typical power supply making use of such a pro-

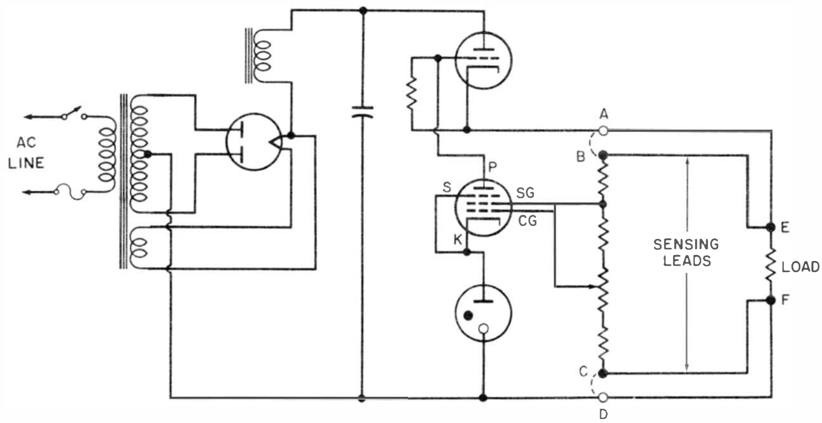


Fig. 5-8. Basic output-terminal arrangement for remote sensing.

vision is illustrated in Fig. 5-8. The positive and negative output terminals (A and D, respectively) are connected to the load—which, for the purpose at hand, is assumed to be some distance from the power supply. As a consequence, an appreciable potential drop occurs in wires AE and DF. If terminal A were connected to terminal B, and C to D (as indicated by the dashed lines in Fig. 5-8), the voltage regulation across load terminals EF would not be as good (that is, as low) as the regulation across terminals AD. On the other hand, suppose terminals B and C are connected by *extra wires* to load terminals E and F (rather than to A and D). Then the resistance of wires AE and DF will be regulated out.

Wires BE and CF can be of relatively small gauge, since they do not carry appreciable current. Wires AE and DF must have large enough cross sections to comply with safety requirements. However, they need not be selected primarily to maintain regulation, because their resistances will now be effectively reduced by the gain of the amplifiers in the power supply. Thus, although four wires are required instead of two, less copper is needed to interconnect the power supply and remote load.

POSITIVE REGULATION

It is possible to make the load voltage actually rise in response to a decrease in AC line voltage or an increase in load current. Although such overregulation is not frequently desired, the techniques for accomplishing it are sometimes

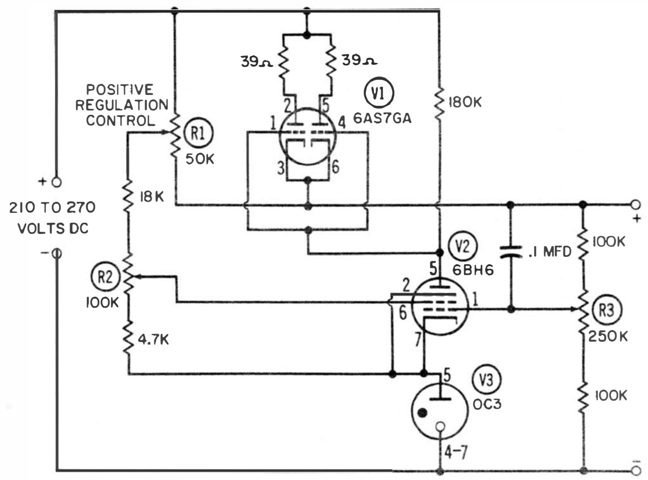


Fig. 5-9. A 180-volt, 150-milliampere voltage regulator that has provision for positive regulation.

useful as a means of more closely approaching zero regulation. For example, in Fig. 5-9 the screen grid of DC voltage amplifier V2 can be, to an adjustable degree, supplied from the unregulated DC side of series losser tube V1. The screen grid of the pentode acts as another control grid in the electron stream. Much lower voltage amplification is displayed at the plate, with respect to small variations of screen-grid voltage, than to like variations at the control grid. However, the excursions of unregulated DC voltage (resulting from changes in the AC line voltage) can be considerably greater than the variations in amplitude of the error signal at the control grid. Therefore, permitting the screen grid to sample the unregulated DC voltage can exert

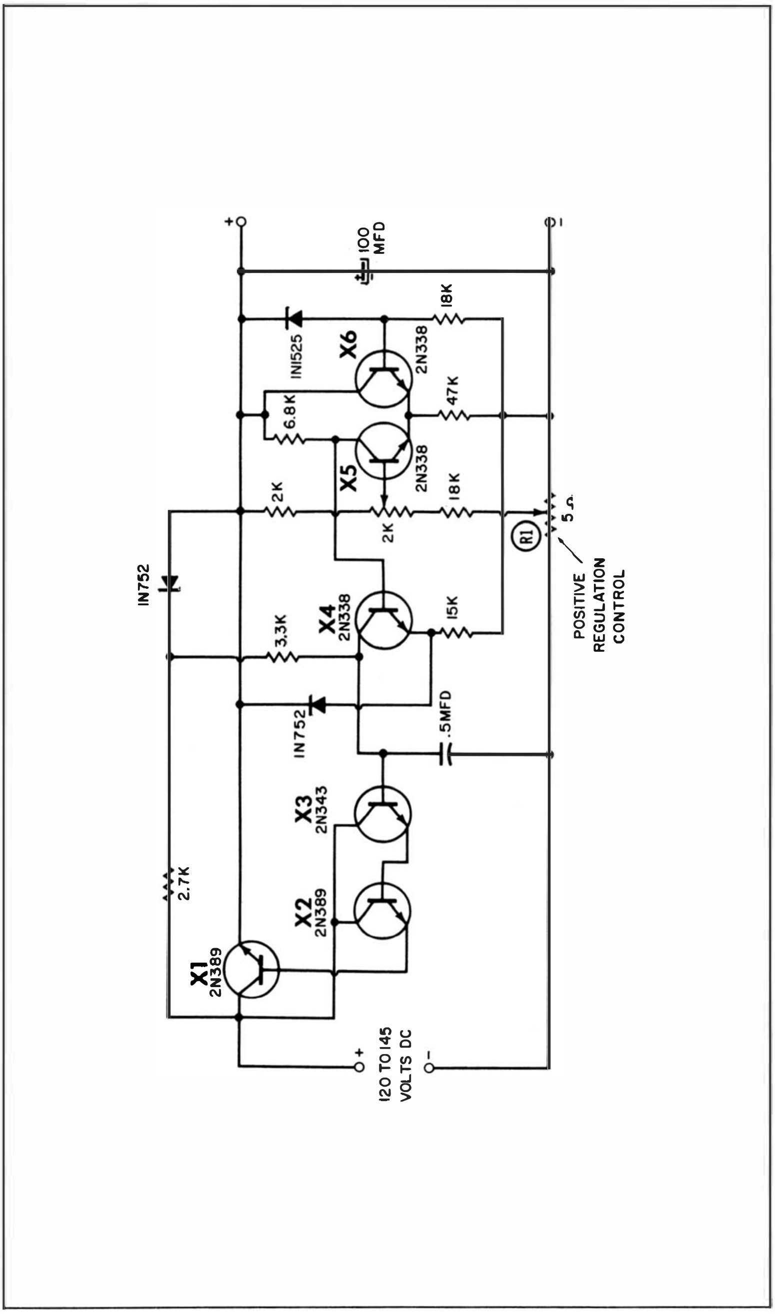


Fig. 5-10. A transistorized 100-volt, 0.5-ampere voltage regulator that has provision for positive regulation.

considerable effect on the regulating characteristics of the circuit.

In this circuit, R1 (the 50K potentiometer) is adjusted to provide the desired amount of overcompensation. The 100,000-ohm potentiometer (R2) permits compensating for the effect which the adjustment of R1 has on the screen-grid voltage; the intended function of the latter potentiometer is not to change the screen-grid voltage, but rather its *source*. These adjustments can be facilitated by connecting a voltmeter between the screen grid and cathode of V2. If R2 is then adjusted to maintain a constant screen-grid voltage as R1 is also being adjusted, the regulating characteristics of the circuit can be varied while the nominal output voltage is held at nearly a fixed value. Adjustment of the nominal output voltage is conventional, being provided by 250K potentiometer R3 in the control-grid circuit of V2.

Another method of providing overregulation is shown in the solid-state voltage regulator of Fig. 5-10. Differential amplifier X5 senses the load current as well as load voltage. The voltage drop developed across R1 (the 5-ohm potentiometer) or a portion of it constitutes the emitter-to-base bias for X5. This bias, through the amplifier chain, causes X1 to raise the DC output voltage as the load current increases. Thus, this circuit has positive current feedback, in addition to the conventional negative voltage feedback. This type of overregulation frequently occurs inadvertently in heavy-current supplies. In such instances, the current-sensing resistance exists in the connecting wire itself.

A VOLTAGE REGULATOR USING COMPLEMENTARY SYMMETRY

An interesting voltage regulated power supply is shown in Fig. 5-11. By means of *complementary symmetry*, an

Table 5-7. Transistor and semiconductor diode data for the circuit of Fig. 5-11

Component	Description	Circuit Function	Type
X1	PNP Transistor	Series losser	2N404, 2N1379, 2N1374, etc.
X2	NPN Transistor	Comparator and DC amplifier	2N697, 2N338, 2N834, etc.
D1, 2, 3, 4	Junction Diodes	Rectifier bridge	1N91, 1N537, 1N2613, etc.
D5	Zener Diode	Voltage reference	1N751

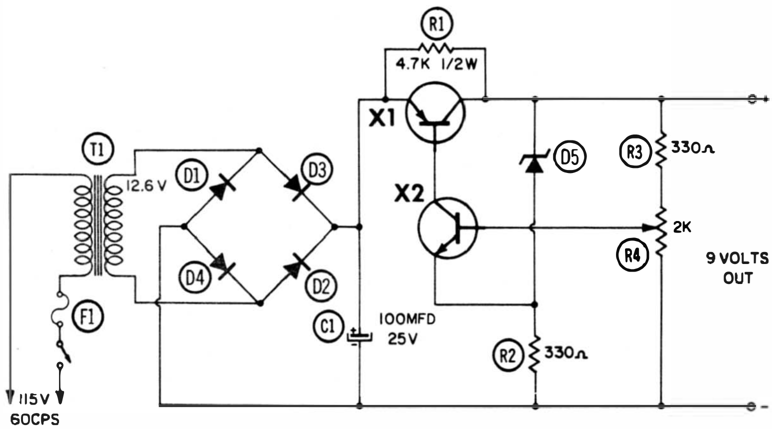


Fig. 5-11. Representative complementary symmetry voltage regulator power supply.

NPN transistor functions as a comparator/DC amplifier and a PNP transistor operates as a series losser. Note that the output is derived from the collector of the PNP transistor, rather than the emitter as is generally encountered in regulator circuits employing like-poled transistors throughout. The series losser transistor performs as an emitter follower stage. This can be readily seen by considering that the basic relationships pertaining to an emitter follower stage are not altered by interchanging the DC supply to the collector and the load. A unique feature of this circuit is that the performance is, for many applications, not downgraded by the use of a relatively inexpensive germanium PNP transistor and silicon NPN transistor. The circuit is very flexible, however, and will perform well with various combinations of different PNP and NPN transistors. This includes the use of an NPN series losser and a PNP voltage amplifier, if considerations other than the cited cost factor so dictate. (Of course, such an inversion requires appropriate changes in the polarities of the applied direct current and the zener diode.)

Fig. 5-11 together with table 5-7 specify components suitable for operation at 9 volts. This makes the power supply useful as a substitute DC source for the home operation of small transistor radios. When so used, a surprising improvement in quality of sound reproduction will generally be observed. This is because the so-called "transistor radio sound" is overcome. The distortion is not due primarily to the small speaker as is commonly supposed, but is a consequence of the poor dynamic regulation of transistor batteries. Most of these sets employ push-pull class B output stages. A poorly regulated power source is one of the most conducive factors for generating distortion in class B amplification. This seems to be even more the case with transistors than with tubes wherein the closer match permits better cancellation of even order distortion products.

This regulated supply can also be used to operate the transistor radio from the 12-volt automobile battery. Many of these sets are sufficiently sensitive and operate quite well from their own ferrite antenna.

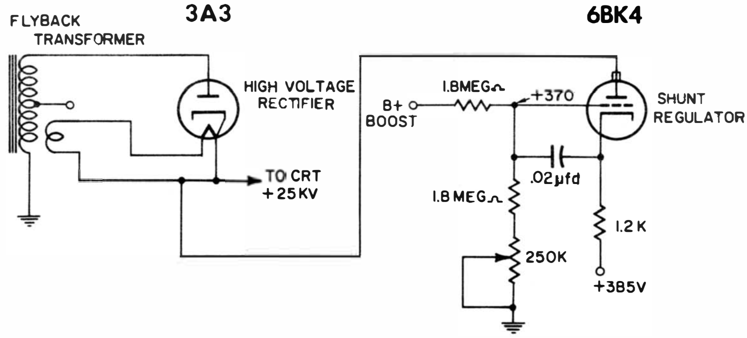


Fig. 5-12. Simplified schematic of a high-voltage regulator circuit in a color television circuit.

Fig. 5-12 is a simplified circuit of a shunt-voltage regulating technique found in a number of color television receivers. In this circuit the 6BK4 shunt regulator tube works in conjunction with the internal resistance of the 3A3 half wave rectifier. The objective is to maintain the nominal 25 kilovolts applied to the cathode ray tube at a near-constant value so that picture size, brightness and hue will not vary with AC line voltage and the setting of the various panel adjustments. There is no voltage reference used, but the regulating action is quite suitable for the intended purpose. The error signal is derived from the B+ boost voltage. Boost voltage is obtained from the damper circuit (not shown) and is a dependent function of the high voltage developed in the flyback transformer. A rise in the rectified output voltage, for any reason, produces a higher boost voltage. A portion of the higher boost voltage appears at the control grid of the shunt regulator making it more conductive. The high-voltage supply is accordingly loaded more heavily and its output voltage drops.

CHAPTER 6

Special Techniques in Closed-Loop Regulators

Conventional closed-loop regulator circuits employ losser elements controlled by amplified error signals. The error signals are derived from a portion of the output voltage or current that has been compared to a stable voltage reference. This approach is a good one, for regulation and dynamic output impedance can readily be improved by simply increasing the gain imparted to the error signal. Large amounts of power can be regulated, and only negligible power need be dissipated in the reference element. Despite this, impelling reasons exist for resorting to modified circuit techniques to attain the same objective. DC amplification, for example, is inherently a difficult process because of the drift from cascaded stages. Another shortcoming of the conventional approach to closed-loop voltage or current regulation is that the losser element must dissipate considerable power. This greatly reduces the over-all efficiency of the regulated power supply and necessitates some means of heat removal. Heat removal rapidly becomes a design limitation where large-capacity supplies must be confined within a reasonable cabinet space. Still another obstacle is the design of solid-state supplies capable of reliable performance above several dozen volts. The losser transistor then becomes increasingly susceptible to destruction from shorts and transients. For these and other reasons, we will now investigate regulated power supplies which, although of the closed-loop species, deviate significantly from the circuits presented previously. Final discussion will deal with auxiliary circuits for protecting series losser regulation from destructive overload effects.

CHOPPER STABILIZED SUPPLIES

The problem of obtaining high loop gain of the error signal, without the accompanying drift which characterizes DC amplifiers, can be solved by wholly or partially dispensing with DC amplification. In order to do this, the DC error signal must be converted to AC or pulses to which high gain can be imparted by stable AC amplifiers. After amplification, the relatively high-level error signal is rectified, filtered, and then applied to the control electrode of the conventional series losser element. Although this basically describes the technique of substituting AC for DC amplification, an important consideration remains—merely converting the DC error signal to AC permits the use of AC amplification, but at the same time destroys sensing of the phase or polarity of the original DC error signal. As a consequence, unless polarity sensing is somehow recovered or restored, the losser element will receive a control signal which can just as readily cause the output-terminal voltage to change in the opposite direction required for compensation. Hence, some form of synchronous rectification or DC restoration must be introduced at the output of the AC amplifier. This is done so that positive or negative DC control (whichever is needed) can be fed to the losser element to counteract the changes in line voltage or load current.

How this important function is accomplished can be grasped by studying the schematic diagram of Fig. 6-1. Here, transistors X1 and X2 produce AC amplification of the interrupted, or "chopped," error signal. Transistor X3 provides subsequent DC amplification. Transistor X4 is the familiar emitter follower which drives series losser transistor X5. Although there are three cascaded direct-coupled stages, the DC drift is negligible for two reasons: First, the DC amplification is much lower than the preceding AC amplification. Secondly, the relatively low DC amplification takes place *after* the error signal has been boosted by the AC amplifier. X3 is the only DC voltage-amplifier stage.

In addition to being the first AC amplifier stage, X1 also performs as a zero sensing element. In this function X1 is analogous to the circuitry position occupied by the first DC amplifier stage in conventional regulators—that is, those employing DC amplification throughout. Let us consider the situation at point Y, the base of transistor X1 (Fig. 6-1). When the chopper armature is *not* at position B, the 12-mfd capacitor connected to the base of X1 is charged to the voltage sampled from the DC output terminals of the supply. When the chopper armature *is* at B, however, the positive reference voltage derived from zener diode M13 is connected to the positive plate of the 12-mfd capacitor. If, then, the stored charge from the output terminals has produced a

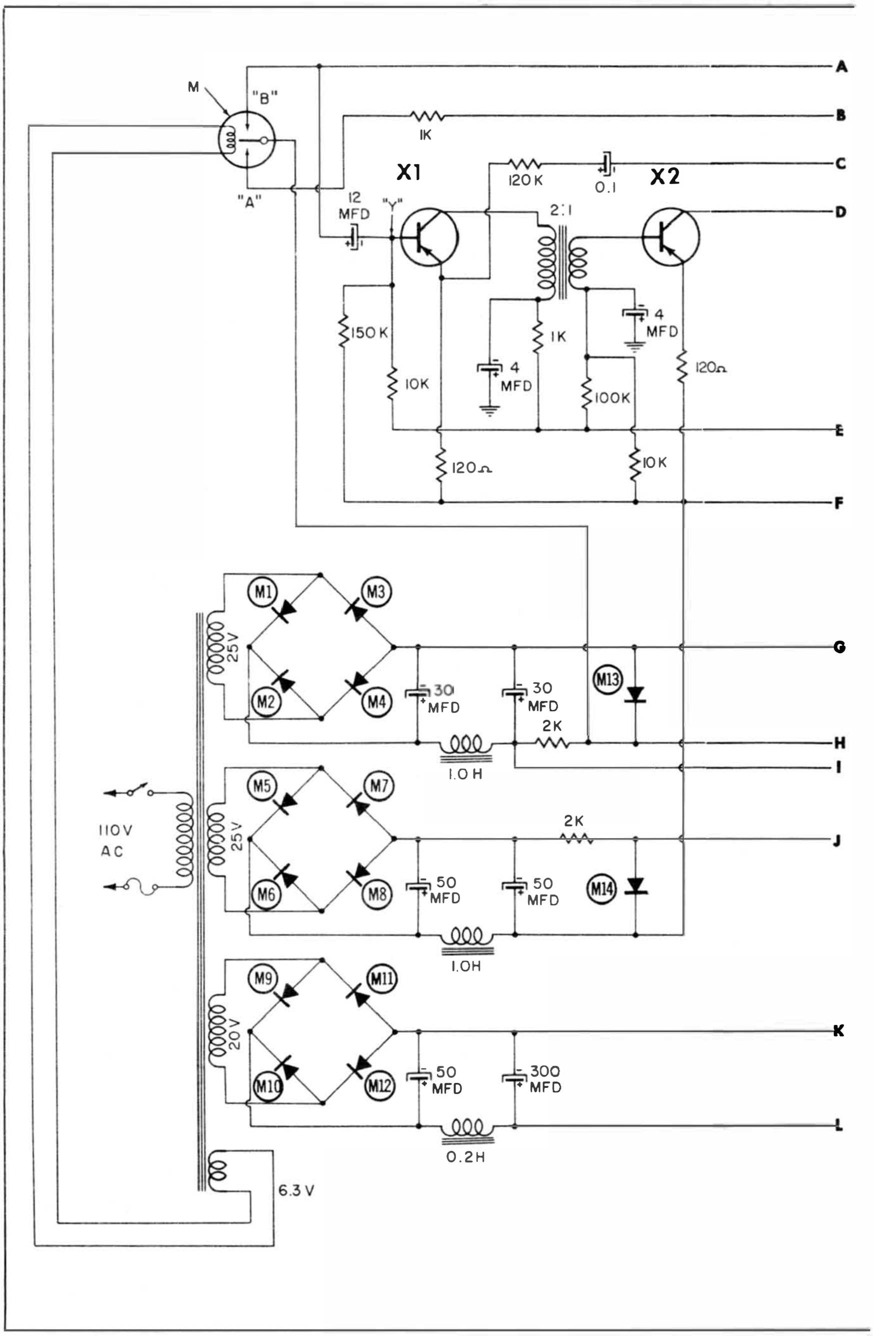
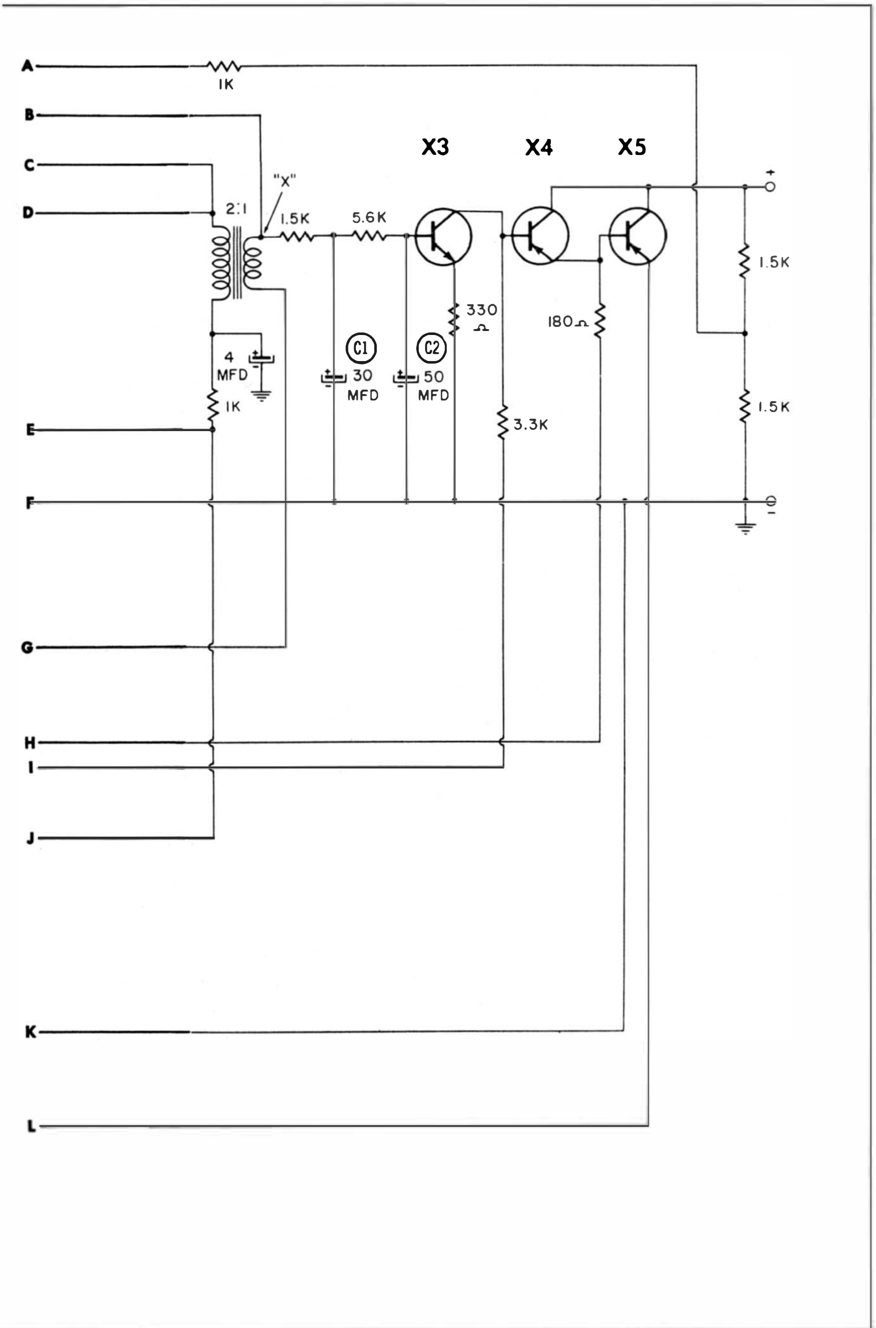


Fig. 6-1. A chopper stabilized voltage regulator



rated at 12 volts and 1 ampere.

higher voltage in this capacitor than the voltage from the reference source, a partial discharge must occur. This results in a *negative* pulse at point Y. If the voltage sampled from the output terminals is *lower* than the voltage sampled from the voltage reference source, this capacitor must receive an additional charge, thereby producing a *positive* pulse at point Y. When the two sampled voltages are the *same,* no change in charge condition is imposed on the 12-mfd capacitor and *no* error signal appears at point Y.

The amplified error signal, when it exists, appears as a relatively high positive or negative pulse (or pulses) at point S. More specifically, it appears at point X when the chopper armature is at position B. When the chopper armature is at position A, point X receives the positive reference voltage from zener diode M13. Capacitors C1 and C2 are the storage elements of a low-pass filter, the function of which is to present the base of X3 with a fairly low-ripple DC bias despite the cyclic fluctuations appearing at point X. Residual ripple then tends to be electronically filtered by the regulating action of the circuit. The DC bias at the base of X3 is varied, by the process just discussed, in such a direction that the terminal voltage change responsible for the error signal is counteracted. It can readily be appreciated that, without the position-A function of the chopper, X3 and consequently series losser X5 would not be informed which way to change conduction in order to correct the terminal voltage. Table 6-1 contains the component data for the chopper circuit in Fig. 6-1.

There are other ways to apply electromagnetic choppers for the basic purpose of modulating the error signal. Furthermore, the DC-to-AC conversion can be accomplished by either vacuum tubes or transistors, as shown in Fig. 6-2. However, this power supply is representative of the objectives attained by providing for AC amplification within the feedback loop. The inference should not be drawn that so-called "chopper-stabilized" amplifiers or power supplies derive their stability from the chopper—in the same sense that negative feedback confers stability, for example. The enhanced "stabilization" comes about because the chopper permits dispensing with at least a major portion of the DC amplification, which would otherwise be required to produce the same over-all loop gain. Theoretically, the chopper amplifier cannot respond more rapidly than the duration of a half-cycle of the chop frequency. In most practical applications, the response time is a small fraction of this time because of the low-pass filter network generally employed to

Table 6-1. Component data for circuit in Fig. 6-1

Component	Description	Circuit Function	Type
M1, M2, M3, M4	Silicon rectifier diodes	Bridge rectifier; DC power source	1N537
M5, M6, M7, M8	Silicon rectifier diodes	Bridge rectifier; DC power source	1N537
M9, M10 M11, M12	Silicon rectifier diodes	Bridge rectifier; main DC power source	
M13	Zener diode	Voltage reference	
X1	Germanium PNP transistor	1st-stage AC amplifier and AC comparator	2N1379
X2	Germanium PNP transistor	2nd-stage AC amplifier	2N1379
X3	Silicon NPN transistor	DC voltage gain stage and DC comparator	2N335
X4	Germanium PNP power transistor	Emitter-follower driver to X5	2N242
X5	Germanium PNP power transistor	Controllable series losser.	2N242
M	DTSP magnetic chopper	Modulates (by interruption) DC error signal; restores DC polarity after AC amplification	AIRPAX 300

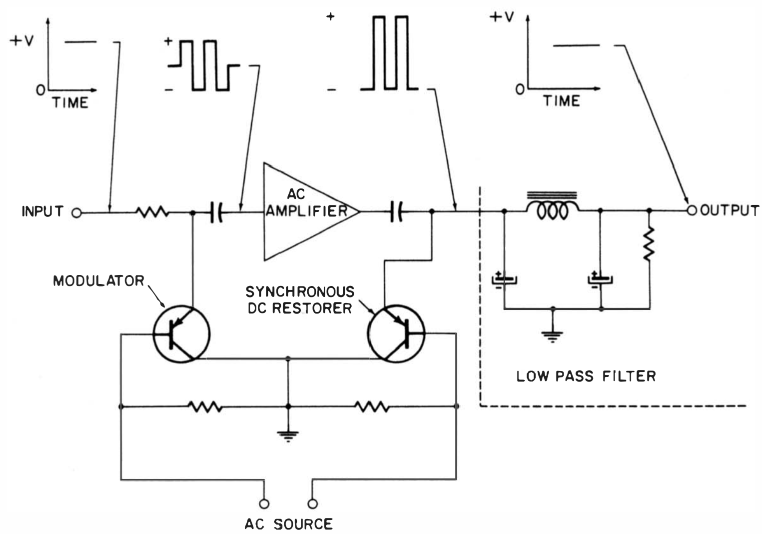


Fig. 6-2. Basic configuration of a transistor chopper.

attenuate the AC components from the demodulated or rectified output. Consequently, regulated power supplies with chopper-stabilized amplifiers tend to exhibit slow response times. Because of the high loop gain provided by AC amplification, however, they are characterized by an extremely low dynamic output impedance within their frequency response.

CONVERTER-TYPE VOLTAGE-REGULATED POWER SUPPLY

The voltage-regulated supply of Fig. 6-3 is designed around a DC-to-DC converter involving essentially power transistors X1 and X2 and saturating transformer T1. Although a 12-volt battery is depicted, a suitable rectifier-and-filter arrangement could be substituted to provide operation from the AC power line. Briefly, the transistor which happens to conduct first when the switch is closed thereby induces voltages in the tapped transformer winding. As a result, conduction simultaneously increases in that transistor and drives the alternate transistor into nonconduction. The current through the emitter-to-collector circuit of the conducting transistor rapidly attains a saturation value beyond which the emitter-to-collector resistance can decrease no further. The current through the portion of the transformer winding involved with the conducting transistor will, however, continue to increase until the transformer core becomes magnetically saturated. When this occurs, the prevailing regenerative cycle of events is abruptly halted, since the electromagnetic induction is insufficient to develop the voltages necessary to maintain forward-conduction bias to the conducting transistor and cutoff bias to the nonconducting transistor.

A rapid collapse now takes place, with the transistors alternating their conduction roles. The collapse is regenerative because once it has started, the *reversed* voltages now bias the respective transistors in the direction *favoring* the switching action. The alternate transistor is now switched *on,* remaining on until saturation of the core again initiates a new half-cycle. A square wave is generated, and the process is extremely efficient because of the low saturation resistance of the transistors and the rectangular hysteresis loop of the transformer core. Ordinary transformer action in the secondary winding then changes the generated voltage to some desired value.

The AC developed in the secondary of transformer T1 is converted to DC by the rectifying circuit comprising power

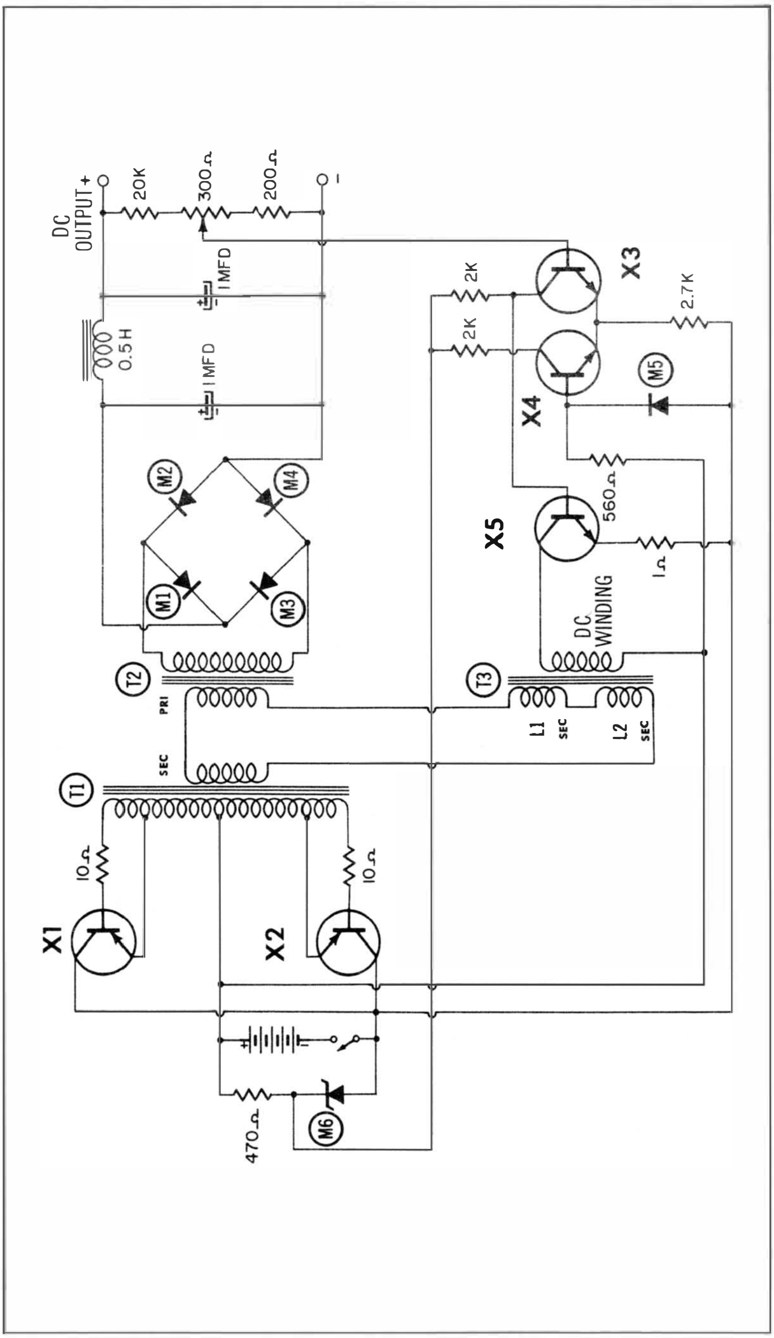
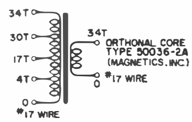
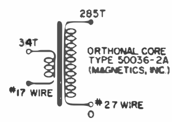
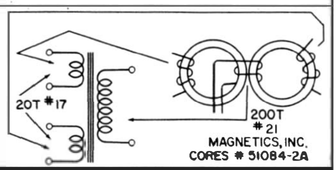


Fig. 6-3. A converter-type voltage-regulated power supply.

Table 6-2. Component data for circuit in Fig. 6-3

Component	Description	Circuit Function	Type
M1, M2, M3, M4	Silicon rectifiers	Elements of full-wave bridge rectifier	1N540
M5	Zener diode	Voltage reference for DC amplifier	1N751
M6	Zener diode	Voltage stabilizer for X3 and X4	1N752
X1, X2	Germanium PNP power transistors	Switching elements in saturating core oscillator	CTP-1509 or 2N442
X3, X4	NPN silicon transistors	DC differential amplifier and comparator	2N338
X5	NPN silicon power transistor	Current-controlling element for DC winding of magnetic amplifier T3	2N1292
T1	Switching transformer	Saturates at alternate half-cycles of oscillation frequency 	
T2	Step-up power transformer	Converts output of oscillator to high voltage 	
T3	Magnetic amplifier	Controls AC input to T2 	

transformer T2, the silicon diode bridge, and the associated filtering components. Notice, however, that the secondary windings of T3 are inserted *between* transformers T1 and T2. T3 functions as a magnetic amplifier, or saturable choke, in which the magnetizing current supplied to the DC winding is controlled by transistor X5. X5 is actuated by the error signal produced by the differential amplifier (X3 and X4), in response to a portion of the DC output voltage and to the reference voltage developed by zener diode, M5.

Regulating action is such that transistor X5 is driven harder into conduction when the DC output voltage tends to fall, thereby bringing the core of T3 more into its region of magnetic saturation. This is accompanied by *decreased inductance* in AC windings L1 and L2, permitting more AC voltage to be impressed across the primary winding of **T2**. In this way, the DC output voltage is restored to its nominal regulated value. If the DC output voltage tends to *rise*, the *converse* sequence occurs.

Although optimum design for T3 generally involves a measure of experimental work, the above method of regulation has the outstanding feature that the transistors are isolated from high voltage. Also, an accidental short across the output of this supply will kill oscillation but should not damage any components. (Table 6-2 gives the parts data for this circuit.)

VOLTAGE REGULATION WITH SILICON CONTROLLED RECTIFIERS

The silicon controlled rectifier is the solid-state counterpart of the thyratron. Like the thyratron, it permits the varying of a current through it by changing the phase (time of occurrence) of the signal applied to a control electrode. In thyratron fashion, once conduction is initiated the control electrode loses control; conduction is extinguished as the AC voltage passes through zero. Fig. 6-4 depicts a simple circuit employing a silicon controlled rectifier to provide half-wave

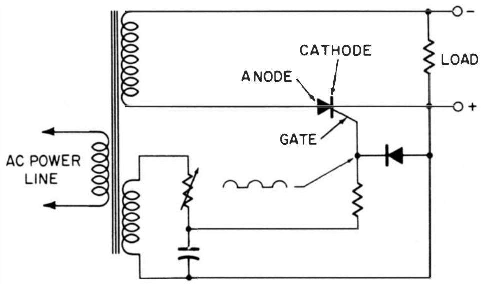


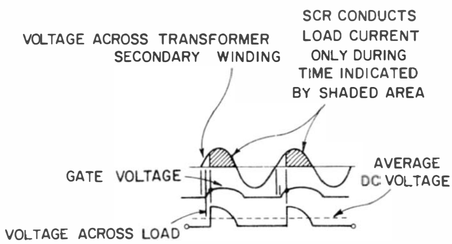
Fig. 6-4. Half-wave silicon controlled rectifier (SCR) supply with phase-shift adjustable gate signal.

rectification. The control electrode, or gate, is supplied with a small voltage from a phase-shifting network. No conduction takes place through the rectifier until the gate potential reaches a certain value. The time required to reach this value is a function of the phase difference between the gate-cathode and anode-cathode voltages. If the phasing is such that conduction is established early in the positive excursion of the AC voltage impressed between anode and cathode (Fig. 6-5A), the duty cycle of the rectified pulses will be relatively high and so will the average DC output voltage.

The converse is true in Fig. 6-5B, where the phasing of the gate signal holds off conduction until the positive excursion of the anode-cathode voltage is nearly completed. (In this simple arrangement, the positions of the variable resistance and the capacitor would have to be interchanged to obtain the amount of phase shift indicated in the waveform diagrams. However, this does not invalidate the basic principle under investigation.)

We see at once that control of conduction time is much more efficient than using variable losser elements to vary the output current and/or voltage. While the silicon controlled rectifier is conducting—whether for a tiny or a major

(A) Moderate delay.



(B) More than moderate delay.

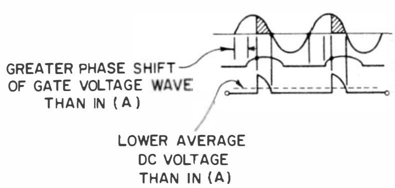
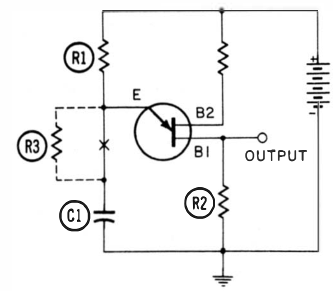


Fig. 6-5. Effect of gate firing-voltage delay on average DC output of circuit in Fig. 6-4.

portion of the impressed half-cycle—the power wasted in the device itself is negligible because saturation exists, regardless of the duration of conduction. Conduction is always at or near maximum, or not at all; consequently, the I^2R losses in a silicon controlled rectifier are very low. This greatly relieves the thermal problem, a major design and operational headache when power transistors are used as losser elements in conventional voltage-regulator circuits.

To impose automatic voltage regulation on a silicon controlled rectifier supply, it is necessary to insert a circuit or device into the feedback path to shift the phase (vary the time of occurrence) of the gate signals in response to an error signal. A relaxation oscillator, involving a unijunction transistor, provides this requisite with a minimum of components. Fig. 6-6A shows a unijunction relaxation oscillator and Fig. 6-6B a somewhat analogous circuit employing the

(A) Unijunction transistor circuit.



(B) Neon-bulb circuit.

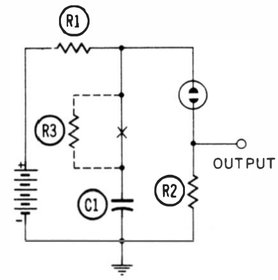


Fig. 6-6. Relaxation oscillators.

more familiar neon bulb. In both instances, a capacitor (C1) is connected across the electrodes, which abruptly change from an open to a closed circuit when the voltage across the capacitor has charged to a certain value. In the neon bulb, this event is brought about by the onset of ionization, which changes the gas from an insulator to a conductor. In the unijunction transistor, the switch from the nonconductive to the conductive state is the result of a change from reverse to forward bias across a PN junction. Fig. 6-7 shows the

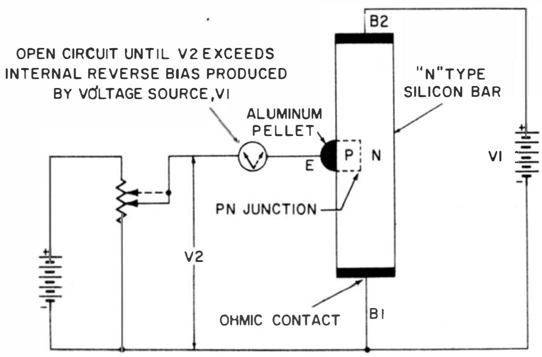
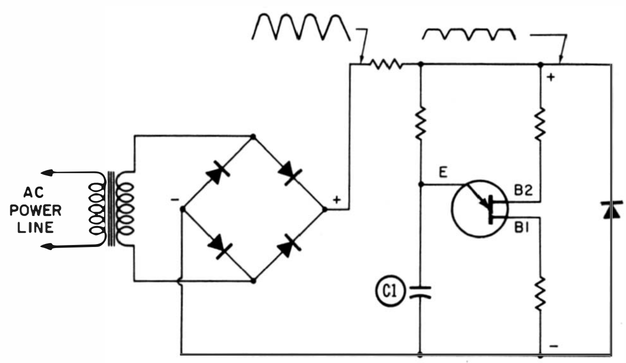
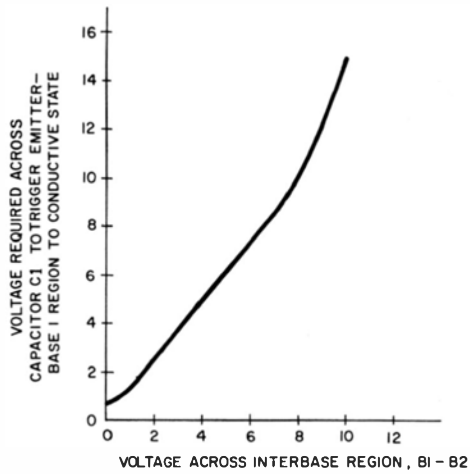


Fig. 6-7. Circuit for demonstrating switching of conductive states in unijunction transistors.

unijunction transistor. It consists of an N-type silicon bar with ohmic contacts established at the ends, at B1 and B2. (This device used to be called a double-base diode.) P-type semiconductor material is alloyed or diffused to the central portion of the bar to form a PN junction. If a current is passed between B1 and B2 by voltage source V1, the voltage distribution within the silicon bar will be such that the PN diode between E and B1 will be reverse-biased. As a result, no current can flow through the diode unless voltage source V2 is high enough to overcome the reverse bias.



(A) Circuit.



(B) Characteristic curve.

Fig. 6-8. Triggering characteristics of the unijunction transistor.

It is important to realize that the presence of a resistance such as R3, in either the neon bulb or the unijunction oscillator of Fig. 6-6, will slow down the repetition rate of the pulses. The reason is that R3 forms a voltage divider (in conjunction with R1) and thereby reduces the voltage applied to capacitor C1. Under this condition, C1 requires a longer charging time to attain the firing potential for its associated device. This property is readily exploited to vary the timing of the gate signal of silicon controlled rectifiers. Another useful characteristic of the unijunction relaxation oscillator is revealed by Fig. 6-8. Note that the capacitor voltage required to trigger the device is a function of the voltage applied across contacts B1 and B2 (Fig. 6-8A). In particular, if the latter voltage is removed, a fraction of a volt across capacitor C1 will suffice to establish forward conduction in diode section E-B1 (Fig. 6-8B). As a corollary of this relationship, it follows that removal of the B1-B2 voltage will cause forward conduction in the E-B1 diode whenever the capacitor voltage is more than a fraction of a volt. Thus, in the circuit of Fig. 6-8A the relaxation oscillator, regardless of its free-running repetition rate, is synchronized to the 60-cps power line in such a way that the generation of pulses must begin anew every 120th of a second.

A complete circuit of a controlled-rectifier voltage-regulated supply is shown in Fig. 6-9. A synchronized unijunction oscillator provides the gate signals. Transistor X1 is actuated from an error signal derived from the difference between the reference voltage across zener diode M6 and a sampled portion of the power-supply output voltage. The error signal controls the effective base-to-collector resistance of X1 (which, for practical purposes, is shunted across the 0.25-mfd charging capacitor). Should the DC output voltage tend to rise, the emitter of X1 will be biased further negative with respect to its base. Since X1 is an NPN transistor in a common-base configuration, it conducts more heavily—that is, the effective resistance represented by the collector-to-base diode is decreased. The 0.25-mfd capacitor then takes longer to attain the potential it needs to fire unijunction transistor X2. This, in turn, delays firing of that silicon controlled rectifier, which is properly polarized from power transformer T1. In alternate half-cycles of the line frequency, the firing of the opposite silicon controlled rectifier is similarly delayed. Because the conduction current in the two silicon controlled rectifiers has been reduced to smaller portions of a half-cycle, a lower average DC current flows to the filter and load (Fig. 6-10). This is manifested as a reduction in DC output voltage, which counteracts the voltage rise responsible for the corrective error signal. The converse sequence of events occurs whenever the output voltage tends to fall (see Fig. 6-11). The power amplification realized in this type of error-sensing circuit can be extremely high. Component data for the circuit of Fig. 6-9 are given in Table 6-3.

DC SWITCHING-TYPE VOLTAGE REGULATOR

Duty-cycle modulated pulsing of the output of a DC power supply is a circuit technique which holds considerable promise for more efficient operation than can be attained from a conventional losser regulator, wherein the resistance of the losser element is varied to compensate for any change in out-

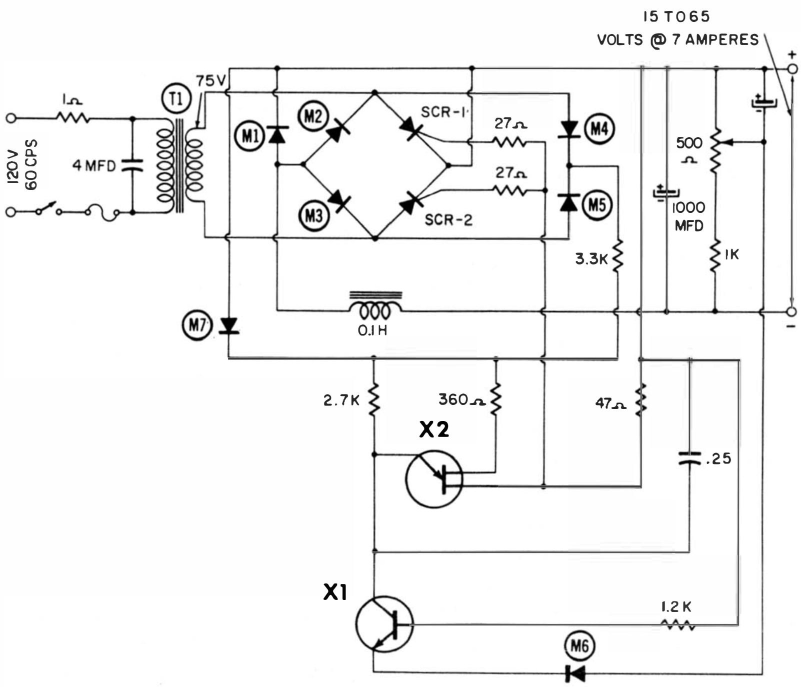


Fig. 6-9. Silicon controlled rectifier (SCR) voltage-regulated supply using variable-conduction time technique.

put voltage. A block diagram of a voltage regulator employing the pulsing or DC switching principle is shown in Fig. 6-12, and the schematic appears in Fig. 6-13. The parts data for this circuit are given in Table 6-4. The significant feature of this supply is that the variable-resistance characteristic of the series losser is replaced by a type of switching action whereby the load current is either fully on or fully off. The

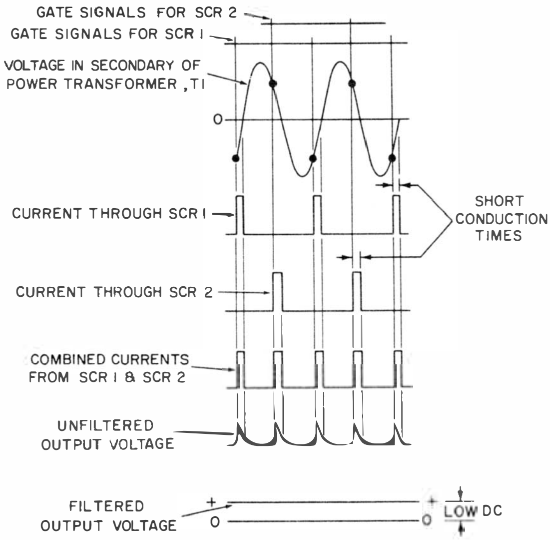


Fig. 6-10. Conditions in circuit of Fig. 6-9 when phasing of gate signal allows short conduction time.

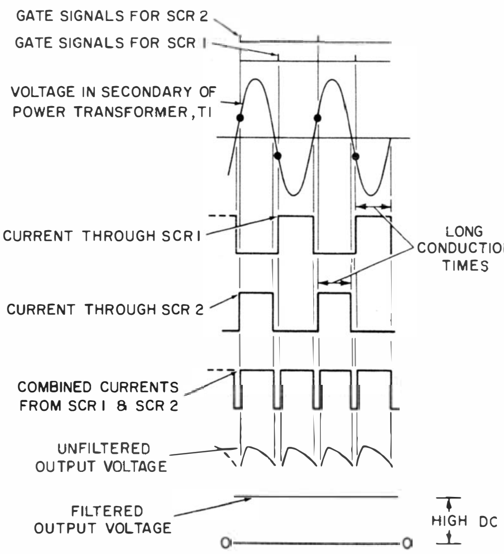
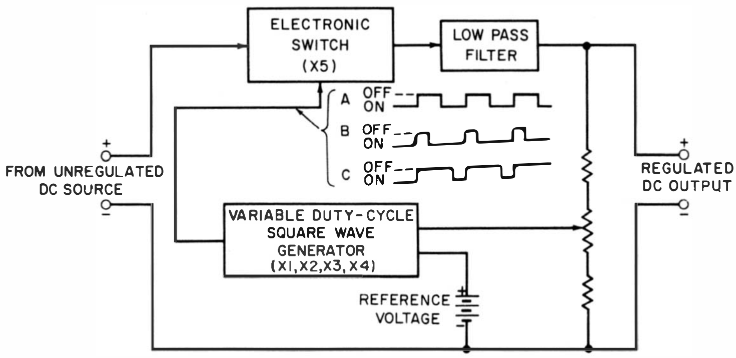


Fig. 6-11. Conditions in circuit of Fig. 6-9 when phasing of gate signal allows long conduction time.

Table 6-3. Component data for circuit in Fig. 6-9

Component	Description	Circuit Function	Type
M1	Silicon diode	Provides current-flow path for energy stored in filter choke	MR 315
M2, M3	Silicon diodes	Rectifiers in bridge comprising M2, M3, SCR-1, SCR-2	MR 315
M4, M5	Silicon diodes	Rectifiers in bridge comprising M2, M3, M4, M5, power supply for gate signal generator X2	1N539
M6	Zener diode	Voltage reference	1N751
M7	Silicon diode	Clips unidirectional pulses supplied to unijunction transistor X2 in order to synchronize gate signals to line frequency	1N1527
X1	Silicon NPN transistor	Error-signal–actuated variable resistance for control of charging time of 0.25-mfd capacitor	2N338
X2	Unijunction transistor	Line-synchronized relaxation oscillator for supplying gate signals to silicon controlled rectifiers SCR-1, SCR-2	2N489
SCR-1 & SCR-2	Silicon controlled rectifiers	Part of main rectifier bridge M2, M3, SCR-1, SCR-2. Controls output voltage	C10B



(A) CONTROL SIGNAL SUPPLIED TO ELECTRONIC SWITCH WHEN OUTPUT VOLTAGE IS AT NOMINALLY REGULATED VALUE.

(B) CONTROL SIGNAL SUPPLIED TO ELECTRONIC SWITCH WHEN OUTPUT VOLTAGE FALLS.

(C) CONTROL SIGNAL SUPPLIED TO ELECTRONIC SWITCH WHEN OUTPUT VOLTAGE RISES.

Fig. 6-12. Block diagram of DC switching regulator.

average load current—therefore, the value of the filtered output voltage—depends on the ratio of time on to time off. It is this ratio which is varied by the error signal. If the output voltage rises, less "on" time is produced—to the extent that the voltage is nearly restored to its nominal regulated value. The converse events occur if the output voltage falls. When the switch is on, its resistance is ideally zero and, practically, quite low. Therefore, the I^2R loss in the

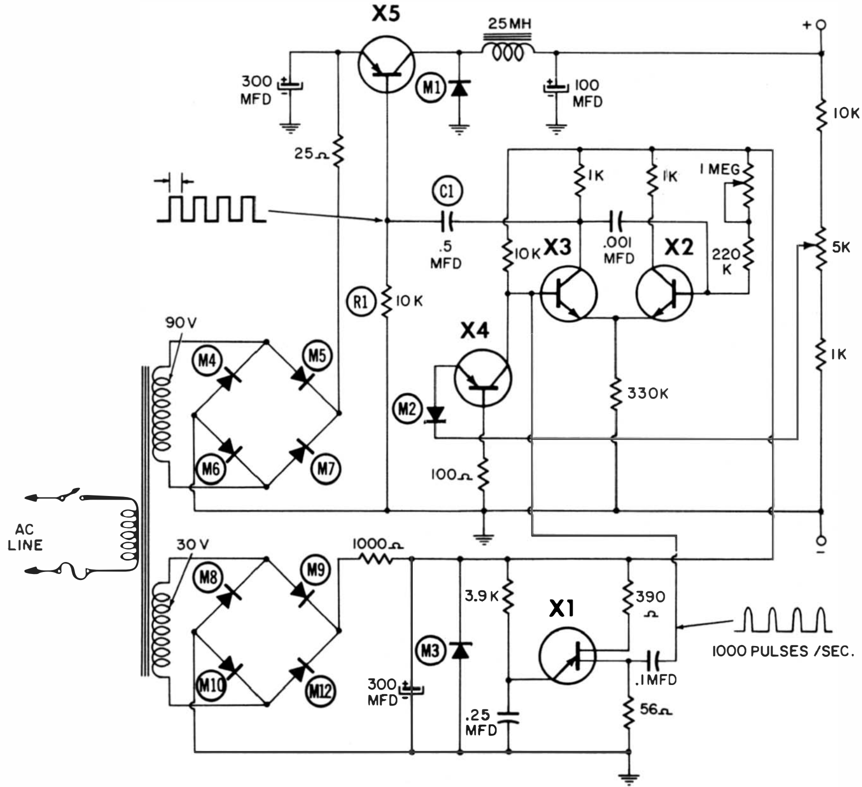


Fig. 6-13. Schematic of typical DC switching-type voltage regulator with an 80-volt, 500-milliampere output rating.

switching element is relatively small. The thermal problem is considerably relaxed with respect to a conventional losser regulator of like capacity. This type of supply has another advantage—high-gain DC amplification is not required for the error signal.

In the schematic diagram of Fig. 6-13, transistors X2 and X3 are connected as a single-shot multivibrator which, not

Table 6-4. Component data for circuit in Fig. 6-13

Component	Description	Circuit Function	Type
M1	Silicon diode	Provides current path for energy stored in filter choke during "off" time of X5	1N539
M2	Zener diode	Voltage reference for error signal	651-C5
M3	Zener diode	Voltage regulator for X1, X2, X3 power supply	1N1527
M4, M5, M6, M7	Silicon diodes	Bridge-rectifier elements for X1, X2, X3 power supply	1N537
M8, M9, M10, M11	Silicon diodes	Bridge-rectifier elements for main power supply	1N539
X1	Unijunction transistor	Supplies trigger pulses to monostable multivibrator	GE 2N489
X2, X3	NPN transistors	Monostable multivibrator	2N338
X4	PNP transistor	Variable resistance for controlling duty cycle of multivibrator	2N217
X5	PNP transistor	Electronic switch	2N389

being free-running, is repeatedly being triggered through its cycle by sharp pulses which have been derived from the unijunction relaxation oscillator X1. The duration of the multivibrator output pulses is then controlled by transistor X4, which acts as a variable resistance in the base-to-emitter circuit of X3. The total effective resistance of the collector-to-base circuit of X4 is governed by the error signal at the emitter of X4.

The output pulses generating from the multivibrator are transferred, through capacitor C1, to the base of switching element X5. Transistor X5 is forward biased by resistor R1 in its base lead, but is turned off while positive pulses are being supplied from the multivibrator. Consequently, the ratio of "on" to "off" time of transistor X5 is controlled by the duty cycle of the multivibrator. Inasmuch as the switching rate of this network is relatively high (approximately 1 kc in this circuit), the response time is not limited by the power-line frequency, as it is in the AC switching regulator such as the phase-control SCR supply in Fig. 6-9. Diode M1 provides a current path through which the stored energy in the filter choke can safely dissipate itself when the switching element is turned off. This produces a flywheel effect which contributes to filtering action. Without M1, the emf developed across the choke, when X5 is switched off, could destroy the transistor.

OVERLOAD PROTECTION FOR SERIES LOSSER VOLTAGE REGULATORS

Some words are now in order with regard to the protection of solid state regulated supplies from the effects of short circuits. The transistor used as the series losser is often far from being an inexpensive item. Yet, this circuit element is extremely vulnerable to damage from overcurrent (and over-voltage in some cases). In some of the higher voltage supplies, the situation is aggravated by the fact that the output terminal voltage exceeds the safe collector-emitter voltage. This operational mode is perfectly satisfactory when a normal load is being powered; the voltage dropped across the series losser need be only a relatively small fraction of the voltage actually available from the rectifier-filter circuit. Such a voltage drop can be safely within the allowable ratings of the losser transistor. However, the occurrance of a short circuit at the output terminals of the regulator immediately places the rectifier output voltage across this transistor. Obviously, some investment of time and money in protective techniques becomes worthwhile in those instances where short circuits are probable. This certainly pertains to laboratory supplies used for experimental and developmental purposes.

The deployment of the control transistor as a shunt, rather than a series losser element neatly circumvents the short circuit problem. A short circuited shunt regulator is deprived of its voltage and thus automatically removes the shunt control transistor from danger. (A continued overcurrent situation can, of course damage the series resistance or components in the rectifier circuit). However, the very low efficiency of the shunt regulator when delivering low-load current tends to lessen its popularity. Considerable protection can be conferred upon the generally more efficient series type regulator by making use of appropriate design philosophies and circuit techniques.

Where feasible, the use of larger than necessary germanium power transistors sometimes constitutes a big step in the right direction. Such transistors tend to be inexpensive and, when suitably heat-sinked, display relatively long thermal time constants. This decreases the likelihood that a momentary short circuit will damage the transistor. Moreover, it is only with such large transistors that fuses can provide a measure of protection. Actually, the basic problem is centered about the difficulty of preventing transistor damage during the interval in which the fuse is responding to a condition of overload. It is, in most cases, best to incorporate some means of protection which asserts itself more rapidly than the many tens of milliseconds involved in blowing a fuse with excessive current.

Such protection can be obtained by incorporating semiconductors switching devices which have completed an evolutionary phase not unlike that of the first decade following the introduction of the junction transistor. The products now available are characterized by a high order of reliability, refinement in characteristics, and wide variety of ratings. These devices find useful application in protective circuits associated with series losser regulators. They have also been successfully employed in regulators based on switching or phase-shift control techniques. Four types appear to have met widespread approval from circuit designers. These are: the silicon-controlled rectifier (SCR), the silicon controlled switch, the gate-turn off switch, and the controlled AC switch.

The general device characteristics of the silicon-controlled rectifier and the silicon-controlled switch are similar. Both lose gate control of forward conduction current when once turned on in DC circuits. Turn-off must be accomplished in both devices by momentarily bringing about a drastic decrease or an interruption in forward conduction current. (This is automatic in AC circuits due to the passage of the AC wave through zero.) Turn-on in both devices is accomplished by a tiny amount of power relative to the power passed to the load. The silicon-controlled switch is fabricated of two transistor-like structures, whereas the silicon-controlled rectifier utilizes a four-layer, three-junction construction. The application-oriented difference between the two devices involves the fact that the silicon controlled rectifier is made in larger current carrying capacities than is the silicon-controlled switch. The former switching device is generally used in circuits where current flow ranges from about one-quarter ampere to hundreds of amperes. The latter performs similar switching functions where the current flow ranges from tens to hundreds of milliamperes.

Typical General Electric silicon-controlled switches are listed in Table A-12 in the Appendix. Table A-13 (also in Appendix) shows a portion of a large number of silicon-controlled rectifiers made by the same manufacturer.

Unlike the two aforementioned switching devices, the silicon gate turn-off switch *retains* gate control of its forward conduction current in *both* DC and AC circuits. Its use has merit in high efficiency regulated supplies based on series-switching of one DC line. Output voltage or current control of such supplies is achieved by changing either the duty cycle or the switching rate in response to an error signal generated by a change in output. A drawback of these devices is that the turn-off gate power is much greater than the turn-on gate power. For this reason, some designers have felt that ordinary power transistors are at least competitive for this regular application. A family of silicon gate turn-off switches is shown in Table A-14 in the Appendix.

The controlled AC switch is in essence a back-to-back arrangement of two SCR's with a common gate terminal. This device permits full wave control of AC with very high efficiency. It will prove exceedingly useful as a means of pre-regulating a supply which employs some other method for securing fine regulation of output voltage or current. When so used, the voltage and power dissipation requirements of series losser regulating transistors can be considerably relaxed and the regulating range of the power supply can be extended. Its response time is much better than can be obtained with magnetic amplifiers, or with saturable reactors. Table A-15 (in Appendix) lists several controlled AC switches ("triacs") made by the General Electric Company.

Most commonly, the previously described solid state switching devices are phase controlled. This is accomplished by varying the time of occurrance of the firing pulse applied to the gate with respect to the AC being controlled. Generation and phase variation of the firing pulse is conveniently brought about by means of another solid state switching device, the unijunction transistor. A single unijunction transistor can, by operation in appropriate circuits, produce either sharp spikes or square waves similar to those derived from a multivibrator. Its repetition rate is readily synchronized or "slaved" to another frequency. The unijunction transistor can generally be employed to provide greater reliability and better economy than other relaxation oscillators suitable for controlling the duty cycle of silicon-controlled switches, silicon-controlled rectifiers, controlled AC switches, and gate turn-off switches. Table A-16 (in Appendix) lists a number of unijunction transistors.

An excellent current overload protection curcuit is shown in Fig. 6-14. Paradoxically, the operation of this circuit is dependent on fuse blow-out. However, the fuse is caused to interrupt the power supply current very much more rap-

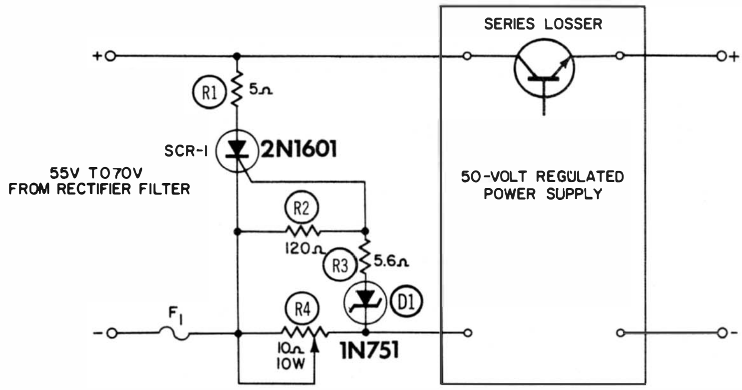


Fig. 6-14. Schematic of protective device in which deliberate blowing of a fuse removes over-current condition.

idly than would be the case with an "unaided" fuse. Under ordinary load conditions, insufficient reverse voltage appears across zener diode D1 to cause it to breakdown. That being the case, the gate of the SCR does not receive firing potential and remains in its non-conducting state. Thus, the only effect of the protective circuitry on normal operation of the voltage regulator is the slight degradation of the regulation of the rectifier-filter system. This generally is of little practical consequence. The rectifier circuit usually has quite poor regulation. Moreover, it is usually not difficult to attain sufficient DC voltage gain in the feedback path of the regulating circuit to largely overcome the effects of regulation in the rectifier-filter system.

Suppose, now, that either a heavy load, or a short circuit is placed across the output terminals of the voltage regulator. The voltage developed across R4 appears in large measure across the junction of D1. The resultant breakdown of diode D1 causes the SCR gate to receive firing potential from the current flowing through R2 and R3. In turn, the SCR is triggered to its conductive state. The rectifier-filter system is thereby loaded more heavily than results from even a short circuit placed at the output terminals of the regulator. Significantly, when this occurs, the series losser transistor in the voltage regulator is relieved of nearly all of its current. The loading of the rectifier-filter system is limited only by R1, the negligible resistance of the SCR, and the fuse. Such loading causes the fuse to blow out more quickly than would be the case without the protective circuitry. Even more important is the fact that the series losser transistor is deprived of dangerous voltage and current, whatever the requisite time for the fuse element to melt. The SCR can fire in about a microsecond from the time an overload appears at the output terminals of the voltage regulator. Once fired, it remains conducting until its anode-cathode current is interrupted. Such interruption would here be brought about by the opening of the fuse. After the overload is cleared and the fuse has been replaced, the SCR again assumes its non-conducting state and the entire system is again ready for normal operation. It can be seen that the overload current at which the zener diode, and therefore the SCR, will conduct is adjustable by means of the rheostat R4.

The basic protection is brought about by the very quick removal of voltage and current from the series losser transistor, this being essentially due to the near-short circuit placed across the rectifier-filter system by the SCR. However, it is also important to minimize the thermal delay of the fuse. Otherwise, rectifier-filter components will be endangered by the heavy current flow. A slow-blow fuse, or a fuse with too high a current-carrying capacity would be contrary to the philosophy of this protective scheme. It is suggested that a fuse no larger than necessary be used to carry rated load current under worst conditions of high AC line voltage. The semiconductors and resistors employed can be scaled to optimum size and cost with respect to the voltage and currents involved. This method allows some degree of over-voltage protection also. However, it is primarily intended to protect the series losser transistor from the effects of over-current.

The protective schematic shown in the representative circuit of Fig. 6-15 possesses the operational feature of instant recovery to normal performance when the excessive load is removed. In this circuit current limiting is imposed by the action of zener diode D1 in conjunction with series dropping resistance R1. Resistor R1 is adjusted so that the potential developed across zener diode D1 is insufficient for breakdown under normal load conditions. However, with excessive load current the increased voltage drop developed across resistance R1 is impressed across D1, causing breakdown of this diode at its rated avalanche voltage. When this occurs, the base of transistor X2 and the collector of transistor X1 become clamped at a voltage considerably negative with respect to the voltage necessary to sustain the condition of current overload. Inasmuch as these are **NPN** transistors, such action limits current flow in the Darlington pair, X1 and X2. Thus, even a short cir-

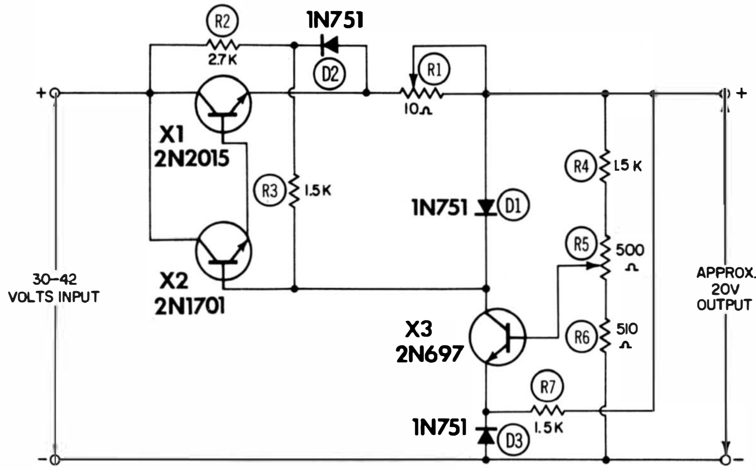


Fig. 6-15. Representative schematic of protective device with instant recovery feature.

cuit across the output terminals cannot result in excessive current flow through transistor X1. Some degradation of regulation takes place as a consequence of resistor R1, but this need not be appreciable if it has a low breakdown voltage. This method provides protection against thermal destruction of the series transistor X1. For complete protection of this transistor, it is also necessary to spare it from greater-than-rated collector-emitter voltage. This can be accomplished by appropriate selection of X1 or by also incorporating the protective technique illustrated in Fig. 6-16.

In the event economic, or other considerations preclude the use of a series losser transistor with a collector-emitter voltage rating above the voltage provided by the rectifier-filter system, the simple over-voltage protective technique

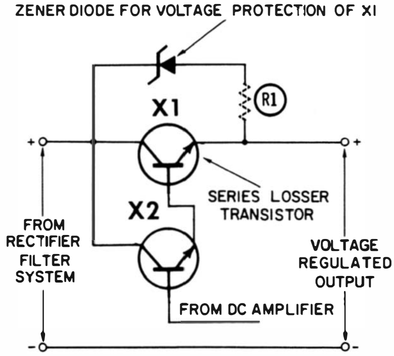


Fig. 6-16. Technique for over-voltage protection of the series losser transistor.

of Fig. 6-16 is recommended. When the regulator is supplying normal load current, the voltage drop across the series losser transistor is insufficient to breakdown the zener diode. Excessive load current will, however, develop breakdown voltage across this diode. When this occurs, the series losser transistor is protected from voltage higher than that corresponding to the zener breakdown voltage. The zener breakdown voltage should be higher than it would receive during normal operation, but lower than the maximum rated collector-to-emitter voltage of the losser transistor. This technique is most readily applied to low- and medium-power systems. The wattage rating of the zener diode may, in some cases, be sufficient to safely withstand the effects of momentary short circuits. It is more desirable, of course, that this protective diode be able to carry the continuous current of a sustained short circuited output. It is here that difficulties may arise on higher power systems, for a zener diode of very large power handling capacity may be required. (If R1 is included, it must obviously be a compromise value. Too low a value will not protect the zener diode. Too high a value will prevent protection of transistor X1.)

This voltage protection method may not give complete protection to the rectifier-filter components. However, in many cases, these may then be protected by a fuse or circuit breaker. This technique can be used in conjunction with other methods of protection. For example, in a high voltage system, the current limiting circuit of Fig. 6-15 should also incorporate a zener diode across the collector-emitter terminals of transistor X1.

The protection of series regulators handling high power or high currents often presents special problems. Energies involved in surges from abrupt changes in current can be considerably more dangerous to components despite the scaling up of ratings over those used at lower power levels. Additionally, fuses and circuit breakers for high currents generally involve inordinately long actuating times. These matters lead to the design of protective circuitry such as depicted in Fig. 6-17. Here, not only is the connection between rectifier-filter system and the series regulator interrupted, but a moderately heavy artificial load is shunted across the rectifier output. Such a load greatly dampens oscillatory surges. The energy in such surges could endanger components in the rectifier filter system if the protective circuit only interrupted current flow in response to an overload or a short at the regulator output terminals.

In Fig. 6-17, the following is true during normal operation: zener diode D1 does not conduct. SCR1 does not conduct. SCR2 does conduct. SW1 is closed. SW2 is open. (Conduction of SCR2 has been brought about by an initial momentary closure of SW2.)

Suppose now, that loading conditions at the output terminals of the regulator cause excessive current consump-

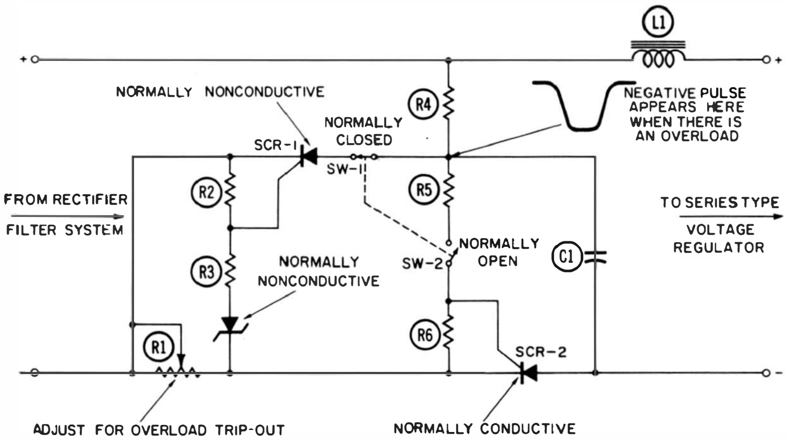


Fig. 6-17. SCR protective circuit diagram.

tion. The higher voltage drop thereby developed across resistance R1 causes breakdown of zener diode, D1. This, by virtue of the connection to the junction of R2 and R3, makes the gate of SCR1 positive with respect to its cathode. As a consequence, SCR1 is turned on and places artificial load R4 across the output of the rectifier-filter system. This, in turn, sends a negative-going pulse through capacitor C1 to the anode of SCR2. Because of this pulse the current flow through SCR2 is momentarily depressed below the value required to maintain conduction. As a result, SCR-2 becomes nonconductive, thereby freeing the series losser transistor from the effects of excessive power dissipation. L1, a small inductance generally in the neighborhood of several tens of millihenrys, aids in the generation of a negative pulse with sufficient energy to reliably turn off SCR2.

Resistor R4 should consume about a third of the maximum rated output current of the system when in normal operation. This allows SCR1 to be a smaller unit than SCR2. Resistor R1 is adjusted to develop sufficient voltage across zener diode D1 to cause breakdown at some desired overload current. Switches SW1 and SW2 should be springloaded to maintain the two switch sections in the contact positions shown except when normally actuated for reset. Actually, this system is not completely foolproof because its operation is inhibited in the event that this switch is *held* in its reset position while the output terminals of the regulator are still shorted. A mechanical arrangement whereby SW2 can be momentarily, but not continuously closed would provide considerable safeguard against such a situation. Conversely, this could be accomplished electrically by inserting a small capacitor in the gate lead of SCR2. A discharge path would have to be provided for the capacitor so that the reset function would not be found inoperative due to trapped charge in this capacitor. A resistance in the several megohm range should suffice. (In any event, the resistance must be high enough so that closure of SW2 cannot trigger SCR2 when the small capacitor is removed from the circuit.) Also, the greater the inductance of L1 the more inherent protection is provided against inordinately long closures of SW2 while the regulator feeds a short circuit. This is because L1 slows down the attainment of maximum current flow. Unfortunately, the degree of such protection provided by L1 is limited in practice by the physical size necessary for protection beyond, say several seconds.

For many purposes this general method can be considered to provide very satisfactory protection. This follows from the fact that a condition of overload would normally be analyzed and remedied *prior* to reset of the power supply.

APPENDIX

Manufacturers Parameters

Table A-1. General Electric one-watt zener diodes

Type Number	V_{ZT} @ I_{ZT} ZENER VOLTAGE Volts @ 25°C ±10% Nominal	Test Current I_{ZT} mA	Maximum Dynamic Impedance Z_{ZT} @ I_{ZT} @ 25°C Ohms	Maximum Leakage Current I_R @ V_R @ 25°C		Typical Temp. Coeffi-cient %/°C	Typical Voltage Regu-lation ΔV_Z Volts	MAXIMUM ZENER CURRENT I_{ZM} — mA	
				V_R Volts	I_R mA			TA = 50°C Max.	TA = 100°C Max.
1N1765	5.6	100	1.2	4.5	5.00	0.021	0.14	150	88
1N1766	6.2	100	1.5	5.0	1.70	0.030	0.16	130	79
1N1767	6.8	100	1.7	5.4	0.96	0.037	0.20	120	72
1N1768	7.5	100	2.1	6.0	0.75	0.044	0.24	109	65
1N1769	8.2	100	2.4	6.6	0.64	0.050	0.28	100	60
1N1770	9.1	50	3.0	7.3	0.52	0.056	0.34	90	54
1N1771	10.0	50	3.5	8.0	0.44	0.062	0.41	82	49
1N1772	11.0	50	4.2	8.8	0.38	0.067	0.48	74	45
1N1773	12.0	50	5.0	9.6	0.34	0.071	0.57	68	41
1N1774	13.0	50	5.8	10.4	0.31	0.074	0.66	63	38
1N1775	15.0	50	7.6	12.0	0.26	0.080	0.86	54	33
1N1776	16.0	50	8.6	12.8	0.24	0.082	0.97	51	31

Table A-2. Honeywell high-current germanium power transistors

PNP Germanium	2N2730 2N2733 2N2736	2N2731 2N2734 2N2737	2N2732 2N2735 2N2738
V_{CE}(sat) @ I_C = —65A	—0.45 V max.	—0.45 V max.	—0.45 V max.
V_{BE}(sat) @ I_C = —65A	—2.0 V max.	—2.0 V max.	—2.0 V max.
h_{FE} @ I_C = —65A	15 min.	15 min.	15 min.
V_{CBO}	—80 V min.	—60 V min.	—40 V min.
V_{CEO}	—60 V min.	—45 V min.	—30 V min.

Table A-3. Motorola power zener diodes

10 WATT Cathode to Case = 1N3993 series Anode to Case = 1N2970 series	50 WATT Anode to Case		Nominal Zener Voltage
1N3993	1N4557&R	1N4549&R	3.9
1N3994	1N4558&R	1N4550&R	4.3
1N3995	1N4559&R	1N4551&R	4.7
1N3996	1N4560&R	1N4552&R	5.1
1N3997	1N4561&R	1N4553&R	5.6
1N3998	1N4562&R	1N4554&R	6.2
1N3999	1N4563&R	1N4555&R	6.8
1N2970&R	1N2804&R	1N3305&R	
1N4000	1N4564&R	1N4556&R	7.5
1N2971&R	1N2805&R	1N3306&R	
1N2972&R	1N2806&R	1N3307&R	8.2
1N2973&R	1N2807&R	1N3308&R	9.1
1N2974&R	1N2808&R	1N3309&R	10
1N2975&R	1N2809&R	1N3310&R	11
1N2976&R	1N2810&R	1N3311&R	12
1N2977&R	1N2811&R	1N3312&R	13
1N2979&R	1N2813&R	1N3314&R	15
1N2980&R	1N2814&R	1N3315&R	16
1N2982&R	1N2816&R	1N3317&R	18
1N2984&R	1N2818&R	1N3319&R	20
1N2985&R	1N2819&R	1N3320&R	22
1N2986&R	1N2820&R	1N3321&R	24
1N2988&R	1N2822&R	1N3323&R	27
1N2989&R	1N2823&R	1N3324&R	30
1N2990&R	1N2824&R	1N3325&R	33
1N2991&R	1N2825&R	1N3326&R	36
1N2992&R	1N2826&R	1N3327&R	39
1N2993&R	1N2827&R	1N3328&R	43
1N2995&R	1N2829&R	1N3330&R	47
1N2997&R	1N2831&R	1N3332&R	51
1N2999&R	1N2832&R	1N3334&R	56
1N3000&R	1N2833&R	1N3335&R	62
1N3001&R	1N2834&R	1N3336&R	68
1N3002&R	1N2835&R	1N3337&R	75
1N3003&R	1N2836&R	1N3338&R	82
1N3004&R	1N2837&R	1N3339&R	91
1N3005&R	1N2838&R	1N3340&R	100
1N3007&R	1N2840&R	1N3342&R	110
1N3008&R	1N2841&R	1N3343&R	120
1N3009&R	1N2842&R	1N3344&R	130
1N3011&R	1N2843&R	1N3346&R	150
1N3012&R	1N2844&R	1N3347&R	160
1N3014&R	1N2845&R	1N3349&R	180
1N3015&R	1N2846&R	1N3350&R	200

Table A-4. Fairchild epoxy-encapsulated silicon NPN transistor

Type	Description	JEDEC Outline		f_T Typical	P_d@25°C Ambient	Beta Ranges h_{FE} @ Conditions				Maximum Voltages			V_{BE} SAT Conditions			V_{CE} SAT Conditions			I_{CBO} (5) Maximum	
						Min.	Max.	I_C	V_{CE}	V_{CER} (3)	V_{CBO}	V_{EBO}	Max.	I_C	I_B	Max.	I_C	I_B	@25°C	@150°C
		TO-5	TO-18	mC	Watts			mA	Volts	Volts	Volts	Volts	Volts	mA	mA	Volts	mA	mA		
2N497	For medium power, fast switching applications	X		50	0.8	12	36	200	10	60	60	8	—	—	—	5	200	40	—	10μA
2N498	For medium power, fast switching applications	X		50	0.8	12	36	200	10	100	100	8	—	—	—	5	200	40	—	10μA
2N656	For medium power, fast switching applications	X		70	0.8	30	90	200	10	60	60	8	1.1	200	40	5	200	40	10μA (9)	—
2N657	For medium power, fast switching applications	X		70	0.8	30	90	200	10	100	100	8	1.1	200	40	5	200	40	10μA (9)	—
2N696	For RF and DC switching applications	X		60	0.6	20 40	60 120	150 150	10 10	40	60	5	1.3	150	15	1.5	150	15	1μA	100μA
2N697	For RF and DC switching applications	X		80	0.6	20 40	60 120	150 150	10 10	40	60	5	1.3	150	15	1.5	150	15	1μA	100μA
2N698	High voltage type, particularly suited for video amplifiers and oscillators	X		70	0.8	20	60	150	10	80	120	7	0.9	50	5	1.2	50	5	0.005μA	15μA
2N699	High voltage type, particularly suited for video amplifiers and oscillators	X		80	0.6	40	120	150	10	80	120	5	1.3	150	15	5	150	15	2μA	200μA
2N699B	High voltage type, particularly suited for video amplifiers and oscillators	X		70	0.87	40 35 20 20	120 — — —	150 10 10 0.1	10 10 10 10	100	120	7	1.0	50	5	1.2	50	5	0.01μA	15μA
2N703	High speed, high current switch		X	70	0.3	40	100	10	5.0	—	25	5.0	—	—	—	0.5	10	0.5	0.5μA	50μA
2N717	For RF and DC switching applications over a wide current range		X	60	0.4	20	60	150	10	40	60	5	1.3	150	15	1.5	150	15	1μA	100μA
2N718	For RF and DC switching applications over a wide current range		X	80	0.4	40	120	150	10	40	60	5	1.3	150	15	1.5	150	15	1μA	100μA

Table A-5. Fairchild epoxy-encapsulated silicon PNP transistors

Type	Description	JEDEC Outline		f_T Typical	P_d@25°C Ambient	Beta Ranges h_{FE} @ Conditions				Maximum Voltages			V_{BE} SAT Conditions			V_{CE} SAT Conditions			I_{CBO}(5) Maximum	
						Min.	Max.	I_C	V_{CE}	V_{CER}(3)	V_{CBO}	V_{EBO}	Max.	I_C	I_B	Max.	I_C	I_B	@25°C	@150°C
		TO-5	TO-18	mC	Watts			mA	Volts	Volts	Volts	Volts	Volts	mA	mA	Volts	mA	mA		
2N721	PNP complement to 2N717		X	70	0.4	20	45	150	—10	—50	—50	—5	—1.3	150	15	—1.5	150	15	1μA	100μA
2N722	PNP complement to 2N718		X	90	0.4	30	90	150	—50	—50	—50	—5	—1.3	150	15	—1.5	150	15	1μA	1μA
2N1131	PNP complements 2N696 suitable for identical uses in opposite polarity	X		70	0.6	20	45	150	—10	—50	—50	—5	—1.3	150	15	—1.5	150	15	1μA	100μA
2N1132	PNP complements 2N697 suitable for identical uses in opposite polarity	X		90	0.6	30	90	150	—10	—50	—50	—5	—1.3	150	15	—1.5	150	15	1μA	100μA
2N2303	Medium frequency amplifier	X		90	0.6	75	200	150	—10	—50	—50	—5	1.3	150	15	—1.5	150	15	1μA	100μA

Table A-6. Fairchild singularly packaged differential amplifiers

DIFFERENTIAL AMPLIFIERS

TO-5	TO-18	Similar Type	NPN/ PNP	h_{FE} Ratio Range	V_{BE1}-V_{BE2} (mV) Max.	$\Delta(V_{BE1}$-$V_{BE2})$ (μV/°C) Max.
2N2060	2N2980	2N1893	NPN	.90-1.00	5.0	10
2N2223	2N2981	2N1973	NPN	.80-1.00	15	25
2N2223A		2N1973	NPN	.90-1.00	5.0	25
	2N2982	2N1972	NPN	.90-1.00	5.0	15
2N2915	2N2974	2N2484	NPN	.90-1.00	3.0	10
2N2916 (Low Noise)	2N2975	2N2484	NPN	.90-1.00	3.0	10
2N2917	2N2976	2N2484	NPN	.80-1.00	5.0	20
2N2918 (Low Noise)	2N2977	2N2484	NPN	.80-1.00	5.0	20
2N3423		2N918	NPN	.80-1.00	10	40
2N3424		2N918	NPN	.90-1.00	5.0	20
2N3726		2N3502	PNP	.90-1.00	5.0	20
2N3727		2N3502	PNP	.90-1.00	2.5	10
2N3728		2N3302	NPN	.80-1.00	5.0	20
2N2729		2N3302	NPN	.90-1.00	3.0	10

Table A-7. Fairchild singularly packaged Darlington amplifiers

DARLINGTON AMPLIFIERS

TO-18	NPN/ PNP	BV_{CBO} (Volts)	h_{FE}@I_C = 100μA Min.	h_{FE}@I_C = 1.0mA Min.	h_{FE}@I_C = 10mA Min.	Max.	h_{FE}@I_C = 100mA Min.	Max.
2N997	NPN	75	1,000	—	4,000	—	7,000	70,000
2N998	NPN	100	—	800	1,600	8,000	2,000	—
2N999	NPN	60	1,000	—	4,000	—	7,000	70,000

Table A-8. General Electric epoxy-encapsulated silicon NPN transistors

Type	h_{FE} (V_{CE} = 10 v; I_C = 2 ma) min.	typ.	max.	BV_{CEO}	f_t typ. mc	P_T @ 25°C (mw)
2N3402	75		225	25	120	900
2N3403	180		540	25	120	900
2N3404	75		225	50	120	900
2N3405	180		540	50	120	900
2N3414	75		225	25	120	360
2N3415	180		540	25	120	360
2N3416	75		225	50	120	360
2N3417	180		540	50	120	360
2N2713	30		90	18	120	200
2N2714	75		225	18	120	200

Table A-9. Fairchild epoxy-encapsulated silicon NPN transistors

Type	h_{FE} min.	h_{FE} max.	@ I_C ma	V_{CEO} volts
2N3563	20	200	8	12
2N3564	20	120	15	15
2N3565	150	600	1	25
2N3566	150	600	10	30
2N3567	40	120	150	40
2N3568	40	120	150	60
2N3569	100	300	150	40
2N3641	40	120	150	30
2N3642	40	120	150	45

Table A-10. Fairchild epoxy-encapsulated silicon PNP transistors

Type	h_{FE} min.	h_{FE} max.	@ I_C ma.	V_{CEO} volts
2N3638	30		50	25
2N3638A	30	180	50	45
2N3639	30	120	10	6
2N3640	30	120	10	12

Table A-11. Fairchild temperature compensated reference diodes

TEMPERATURE-COMPENSATED REFERENCE DIODES

Type	Test Current (mA)	Zener Impedance I_{DC} = 1 0 mA (Ohms)I_{AC} = 1.0 mA	Zener V @ I_{DC} = 10 mA (Volts)	TC (%/°C)
FA8001	10	20	8.0 (Min.)	0.0005
FA8002	10	20	8.0 (Min.)	0.001
FA8003	10	20	8.0 (Min.)	0.002
FA8004	10	20	8.8 (Max.)	0.005
FA8005	10	25	10.8 (Min.)	0.0005
FA8006	10	25	10.8 (Min.)	0.001
FA8007	10	25	10.8 (Min.)	0.002
FA8008	10	25	12 (Max.)	0.005
FA8009	10	30	14.3 (Min.)	0.0005
FA8010	10	30	14.3 (Min.)	0.001
FA8011	10	30	14.3 (Min.)	0.002
FA8012	10	30	15.7 (max.)	0.005

Table A-12. General Electric Silicon-controlled switches

SILICON CONTROLLED SWITCHES (SCS)

	Anode Blocking Voltage	Continuous DC Forward Current	Peak Recurrent Forward Current 100 μsec	Peak Cathode Gate Current	Dissipation	CUTOFF CHARACTERISTICS		CONDUCTING CHARACTERISTICS			MAXIMUM GATE RATINGS		GATE TRIGGERING CHARACTERISTICS			
						I_B		V_F		I_H	V_{GC}	V_{GA}	I_{GTC}	V_{GTC}	I_{GTA}	V_{GTC}
						@V_{AC}	$R_{GC} = 10K$, 150°C	@I_F	Maximum	$R_{GC} = 10K$	$I_{GC} = 20\mu a$	$I_{GA} = 1\mu a$	$V_{AC} = 40v$, $R_L = 800\Omega$, $R_{GA} = \infty$		$V_{AC} = 40v$, $R_L = 800\Omega$, $R_{GC} = 10K$	
Type	Volts	ma	amp	ma	mw	Volts	μa	ma	Volts	ma	Volts	Volts	μa	Volts	ma	Volts
3N81	65	200	1.0	500	400	65	20	200	2.0	1.5	5	65	1.0	.4 to .65	1.5	—.4 to —.8
3N82	100	200	1.0	500	400	100	20	200	2.0	1.5	5	100	1.0	.4 to .65	1.5	—.4 to —.8
3N83	70	50	0.1	50	200	70	20*	50	1.4	4.0†	5	70	150†	.4 to .80	—	—
3N84	40	175	0.5	100	320	40	20*	175	1.9	2.0	5	40	10	.4 to .65	—	—
3N85	100	175	0.5	100	320	100	20*	175	1.9	2.0	5	100	10	.4 to .65	—	—
3N86	65	200	1.0	500	400	65	20	200	2.0	0.2	5	65	1.0	.4 to .65	0.1	—.4 to —.8

Table A-13. General Electric silicon controlled rectifiers

Type Number	JEDEC Number	PRV and V_{BO}	Max. I_{DC} @ Temp. °C	Max. Temp. °C Oper.	Max. Temp. °C Stor.	Max. Req'd Gate Signal @ 25°C T_J
——	2N877	30	.5 A @ 65°C case	150°	150°	0.8V, 200 μa
——	2N878	60	.5 A @ 65°C case	150°	150°	0.8V, 200 μa
——	2N879	100	.5 A @ 65°C case	150°	150°	0.8V, 200 μa
——	2N880	150	.5 A @ 65°C case	150°	150°	0.8V, 200 μa
——	2N881	200	.5 A @ 65°C case	150°	150°	0.8V, 200 μa
C6U	——	25	1.6 A @ 87°C case	125°	150°	0.8V, 1 ma
C6F	——	50	1.6 A @ 87°C case	125°	150°	0.8V, 1 ma
C6A	——	100	1.6 A @ 87°C case	125°	150°	0.8V, 1 ma
C6G	——	150	1.6 A @ 87°C case	125°	150°	0.8V, 1 ma
C6B	——	200	1.6 A @ 87°C case	125°	150°	0.8V, 1 ma
C10U	2N1770A	25	7 A @ 115°C stud	150°	150°	2 V, 15 ma
C10F	2N1771A	50	7 A @ 115°C stud	150°	150°	2 V, 15 ma
C10A	2N1772A	100	7 A @ 115°C stud	150°	150°	2 V, 15 ma
C10G	2N1773A	150	7 A @ 115°C stud	150°	150°	2 V, 15 ma
C10B	2N1774A	200	7 A @ 115°C stud	150°	150°	2 V, 15 ma
C10H	2N1775A	250	7 A @ 115°C stud	150°	150°	2 V, 15 ma
C10C	2N1776A	300	7 A @ 115°C stud	150°	150°	2 V, 15 ma
C10D	2N1777A	400	7 A @ 115°C stud	150°	150°	2 V, 15 ma
C35U	2N681	25	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35F	2N682	50	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35A	2N683	100	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35G	2N684	150	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35B	2N685	200	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35H	2N686	250	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35C	2N687	300	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35D	2N688	400	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35E	2N689	500	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35M	2N690	600	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35S	2N691	700	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C35N	2N692	800	25 A @ 57°C stud	125°	150°	3 V, 40 ma
C50U	2N1909	25	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50F	2N1910	50	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50A	2N1911	100	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50G	2N1912	150	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50B	2N1913	200	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50H	2N1914	250	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50C	2N1915	300	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50D	2N1916	400	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50E	——	500	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50M	——	600	110 A	125°	150°	3 V, 40 ma @ 125°C T_J

Table A-13. General Electric silicon controlled rectifiers (Cont'd)

Type Number	JEDEC Number	PRV and V_{BO}	Max. I_{DC} @ Temp. °C	Max. Temp. °C Oper.	Max. Temp. °C Stor.	Max. Req'd Gate Signal @ 25°C T_J
C50S	——	700	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50N	——	800	110 A	125°	150°	3 V, 40 ma @ 125°C T_J
C50T	——	900	110 A	125°	150°	3 V, 40 ma @ 125°C T_J

Table A-14. General Electric silicon gate turn-off switches

SILICON GATE TURN-OFF SWITCHES

The G5/G6 is a three-terminal PNPN semiconductor switch. While similar in many ways to conventional Silicon Controlled Rectifiers, the G5/G6 can be turned "off" as well as "on" from its gate input terminal. It is suited for low power, high speed DC switching applications.

Type Number	V_{BO}	PRV	Peak Controllable Anode Current (I_P)	Turn-off Gain (Min.)	I_H (Max.)
				I_P = .2 to 1A	
G5U, G511U*	25	25	1 A	5	40 ma
G5F, G511F*	50	25	1 A	5	40 ma
G5A, G511A*	100	25	1 A	5	40 ma
G5B, G511B*	200	25	1 A	5	40 ma
G5C, G511C*	300	25	1 A	5	40 ma
G5D, G511D*	400	25	1 A	5	40 ma
				I_P = .5 to 2A	
G6U, G611U*	25	25	2 A	5	60 ma
G6F, G611F*	50	25	2 A	5	60 ma
G6A, G611A*	100	25	2 A	5	60 ma
G6B, G611B*	200	25	2 A	5	60 ma
G6C, G611C*	300	25	2 A	5	60 ma
G6D, G611D*	400	25	2 A	5	60 ma

Table A-15. General Electric controlled AC switches

Triac—Bi-directional Triodes for gate-control of AC power

Type Number	Breakover Voltage at 100°C T_J Volts	Peak Gate Power Dissipation Watts	RMS Current at 75°C T_C Amps	Gate Triggering Requirements, Typical	Operating Temperature T_J	Outline Drwg.
SC40B, SC41B*	±200	5.0	6	I_G = ±50 ma V_G = ±3 V T_J = 25°C	—40°C to 100°C	105 106
SC40D SC41D	±400					105 106
SC45B, SC46B†	±200	5.0	10	I_G = ±50 ma V_G = ±3 V T_J = 25°C	—40°C to 100°C	105 106
SC45D SC46D	±400					105 106

Table A-16. General Electric unijunction transistors

TYPES			R_{BB}	η	I_V	I_P	I_{EO}		V_{OB1}
Cube Structure	Bar Structure	Bar Structure	Interbase Resistance $V_{BB}=3v$ $I_E=0$ Kohms	Intrinsic Standoff Ratio $V_{BB}=10v$	Min. Valley Current ma	Max. Peak Point Emitter Current μa	Max. Emitter Reverse Current μa	$T_J=25°C$ @V_{B2E}	Min. Base One Peak Pulse Voltage Volts
—	2N2417 2N2417A 2N2417B	2N489 2N489A 2N489B	4.7-6.8	.51-.62	8	12 12 6	2 2 0.2	60 60 30	— 3 3
—	2N2418 2N2418A 2N2418B	2N490 2N490A 2N490B	6.2-9.1	.51-.62	8	12 12 6	2 2 0.2	60 60 30	— 3 3
—	2N2419 2N2419A 2N2419B	2N491 2N491A 2N491B	4.7-6.8	.56-.68	8	12 12 6	2 2 0.2	60 60 30	— 3 3
—	2N2420 2N2420A 2N2420B	2N492 2N492A 2N492B	6.2-9.1	.56-.68	8	12 12 6	2 2 0.2	60 60 30	— 3 3
—	2N2421 2N2421A 2N2421B	2N493 2N493A 2N493B	4.7-6.8	.62-.75	8	12 12 6	2 2 0.2	60 60 30	— 3 3
	2N2422 2N2422A 2N2422B —	2N494 2N494A 2N494B 2N494C	6.2-9.1	.62-.75	8	12 12 6 2	2 2 0.2 .02	60 60 30 30	— 3 3 3
—	5G514 5G515 5G516	2N1671 2N1671A 2N1671B	4.7-9.1	.47-.62	8	25 25 6	12 12 0.2	30 30 30	— 3 3
—	—	2N2160	4.0-12.0	.47-.80	8	25	12	30	3
2N2646	—	—	4.7-9.1	.56-.75	4	5	12	30	3
2N2647	—	—	4.7-9.1	.68-.82	8	2	0.2	30	6
5E35	—	—	4.7-9.1	.62-.82	Frequency 360-440		12	30	Duty Cycle 44-55
5E36	—	—	4.7-9.1	.62-.82	380-420		1.0	30	47-53

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DESIGN AND OPERATION OF REGULATED POWER SUPPLIES

by Irving M. Gottlieb, P.E.

ABOUT THE AUTHOR

Mr. Gottlieb has been writing articles and books on electronics for more than a decade, and his authoritative works have been read and used by hundreds of thousands of technicians, engineers, radio amateurs, hobbyists, and experimenters. He is a California-registered engineer, and holds a B.S.E.E. degree from the University of Southern California. Most of his endeavors have involved electronic circuit designs, primarily in the development of power supplies. In this book he combines his own first-hand knowledge of regulated power supplies with that developed by others in the industry to provide an authoritative guide for anyone interested in the subject. *Frequency Changers* is another popular SAMS book by Mr. Gottlieb.

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