Circuit Concepts

Cathode-Ray Tube

Horizontal
Amplifier

Vertical
Amplifier

Sweep
Generator

Trigger

Power
Supply

Power Supply Circuits

POWER
SUPPLY
CIRCUITS

BY

KENNETH ARTHUR

Significant Contributions
by
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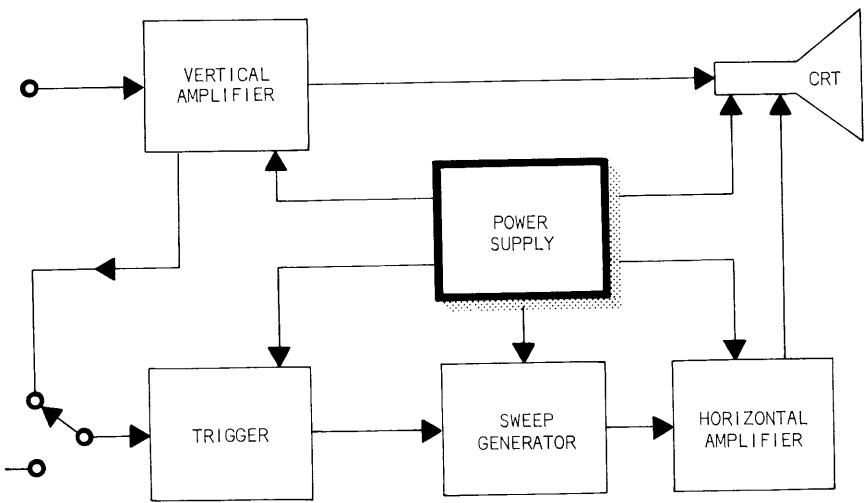
CIRCUIT CONCEPTS

CONTENTS

VERTICAL AMPLIFIER

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INTRODUCTION

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

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

Because of its central role in the operation of
electronic instruments, the power supply offers a
logical point of entry to the study of Tektronix
circuit concepts. The objective of this book is to
set forth the basic principles and practical
considerations governing the operation and design of
Tektronix power supplies. The reader is assumed to
have only a limited background in electronic theory.
Therefore, an operational, rather than mathematical
approach to circuit descriptions, is employed. A
concept-to-instrument cross-index is included as an
aid to the study of specific instruments.
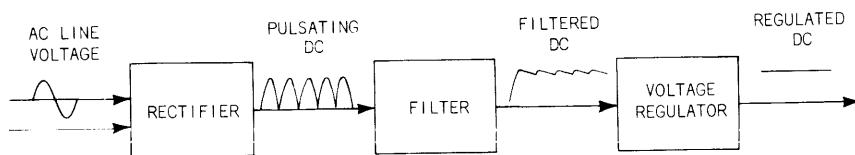
Fig. 1-1.   Power-supply block diagram.
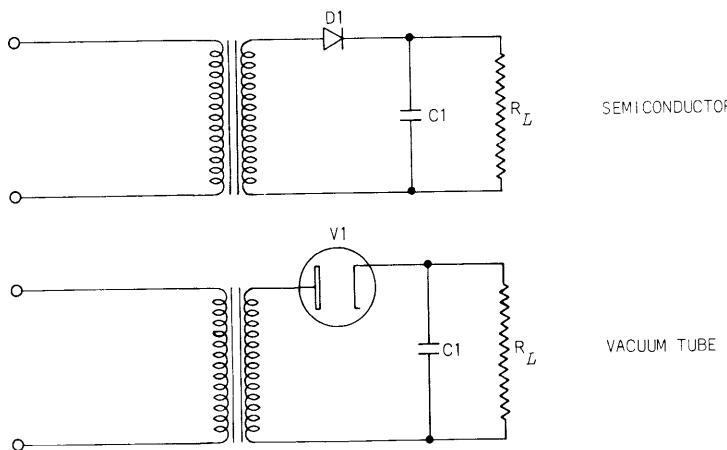
Fig. 1-2.   Half-wave rectifiers.

1

RECTIFIERS AND FILTERS

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

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

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

4

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

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

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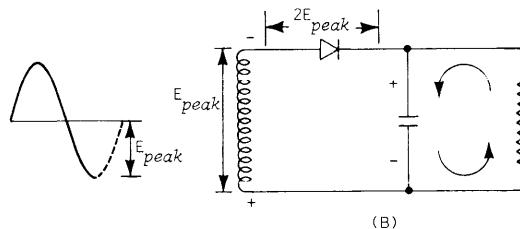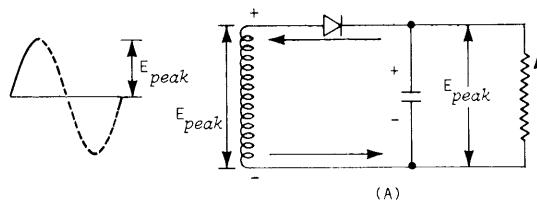
Fig. 1-3. Current and voltage in half-wave
rectifier.

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

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

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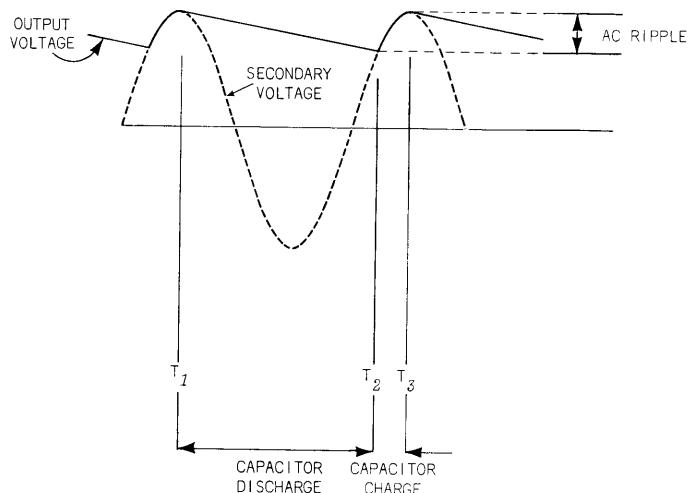
Fig. 1-4. Filtered half-wave rectifier output.

ripple
current

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

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

diode
ratings

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


Fig. 1-5. Current-limiting resistor.

50A

SURGE CURRENT

35A

DIODE CHARACTERISTIC

I

AVERAGE OPERATING
LEVEL

400µs   1ms

T ——⟶
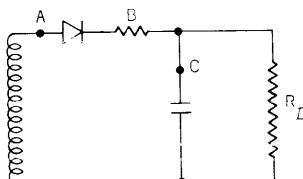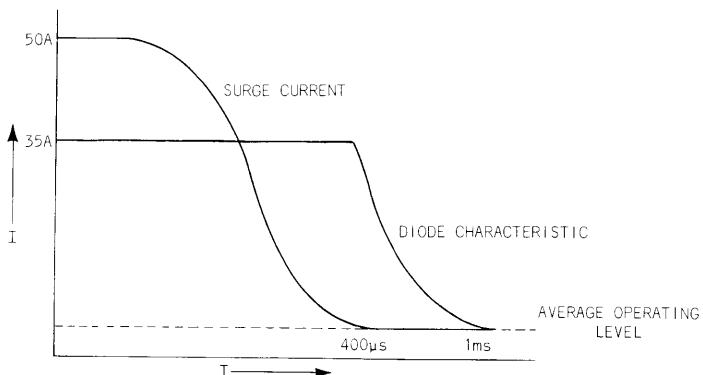
Fig. 1-6.   Typical silicon-diode surge-current
            tolerance.

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

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

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

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

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

full-wave
rectifier

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

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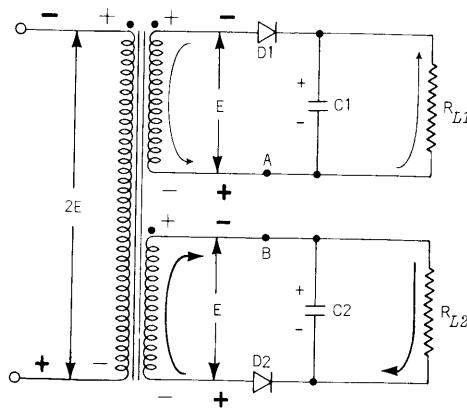
Fig. 1-7.   Evolution of full-wave rectifier -- I.

(A)

(B)

(C)
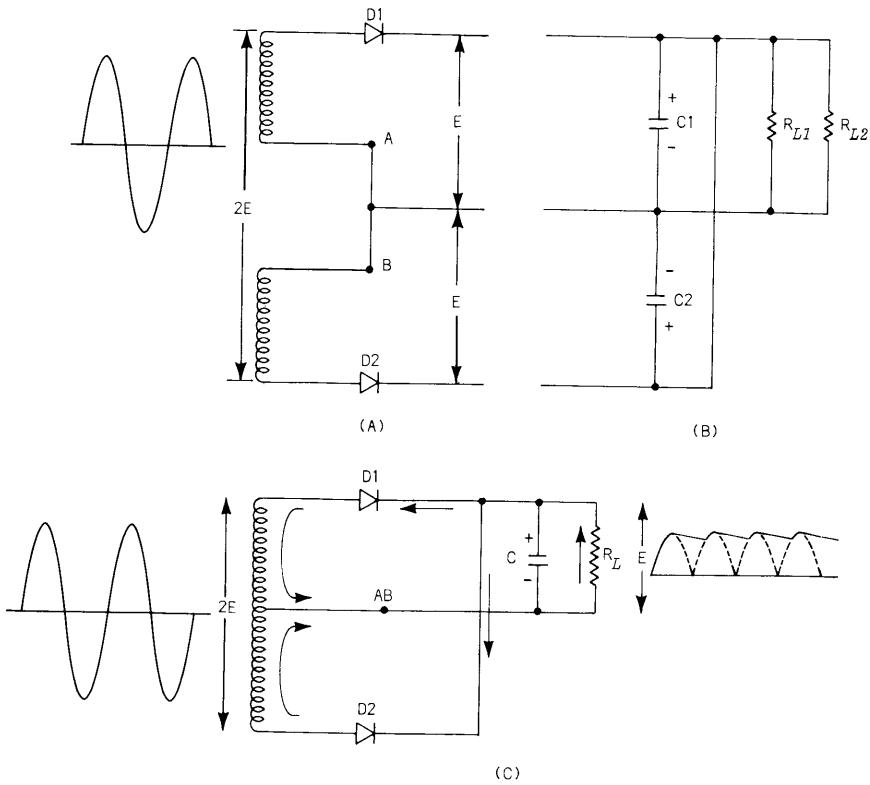
Fig. 1-8.   Evolution of full-wave rectifier -- II.

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

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

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

bridge
rectifier

The appearance of solid-state diodes posed new problems for the power-supply designer. Their small size and weight, relatively low cost, and lack of heater-voltage requirements made them highly attractive as rectifier components. However, the current and voltage limitations of solid-state diodes are in direct contrast to those of the vacuum tube. High current ratings in semiconductor diodes are rather inexpensively attained, while high PIV ratings are quite difficult to attain and therefore expensive. A silicon diode, for instance, rated at one or two amperes costs very little more than one rated at 500 mA. The same type of diode, rated at one or two kV, however, costs a great deal more than one rated at a few hundred volts. Therefore, the aim in designing a rectifier using semiconductor diodes was to reduce the PIV on the diodes, rather than to increase the rectifier's current output. Also, because semiconductor diodes require no heater voltages and are relatively inexpensive, they could be used more generously than would be practical in a vacuum-tube circuit.
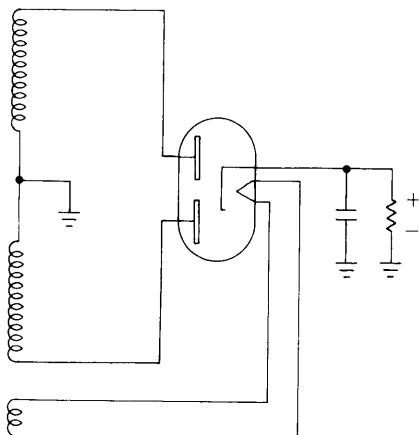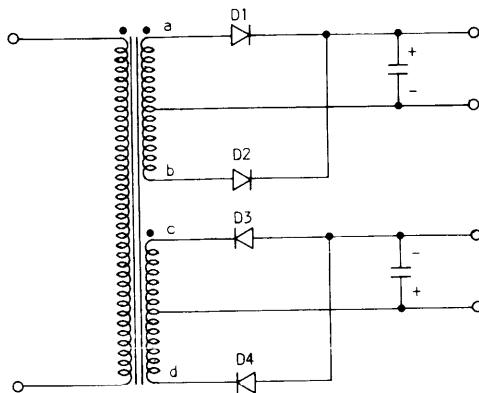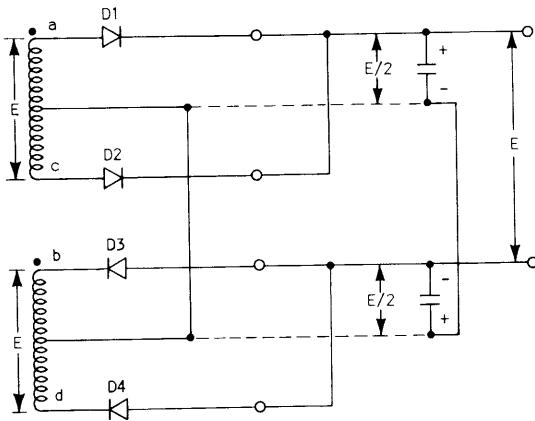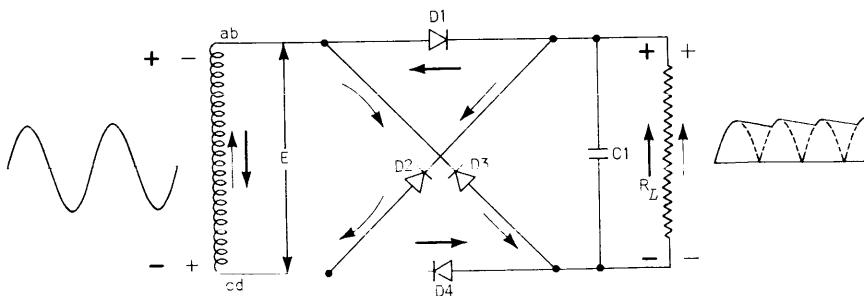
Fig. 1-9. Typical full-wave rectifier.

(A)

(B)

(C)

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

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

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

To find the peak inverse voltage on the nonconducting
diodes, it is only necessary to consider the conducting
diodes as short circuits. The total voltage across
the nonconducting diodes is then the sum of the
secondary and filter capacitor voltages, or 2E.
This voltage is divided across the two diodes so
that each has an impressed PIV of only 1E.
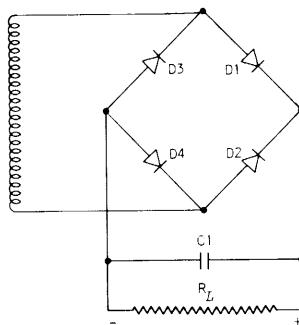
Fig. 1-11. Conventional bridge-rectifier
            schematic.

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

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

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

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

center-
tapped
bridge
rectifiers

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

As the top of the secondary winding becomes positive, diodes D2 and D3 are forward-biased (A). Electrons flow through R_{L1}, D3, the secondary winding, and back through D2 to ground. Diodes D1 and D4 are reverse-biased at this time. One-half the secondary winding acts as a source for R_{L2} (B). On the same alternation as above, electrons flow through R_{L2}, through the secondary winding and back through D2 to ground.
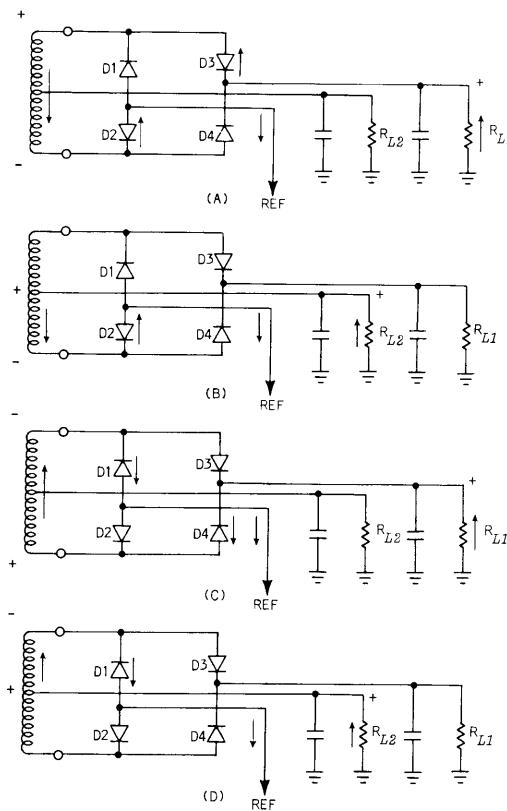
Fig. 1-12. Double-duty rectifier.

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

2

VOLTAGE REGULATORS

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

A voltage-regulated power supply must function so
that its effective internal impedance over a specified
range of load-current and input voltage fluctuations
remains extremely small as compared to the load
impedance.
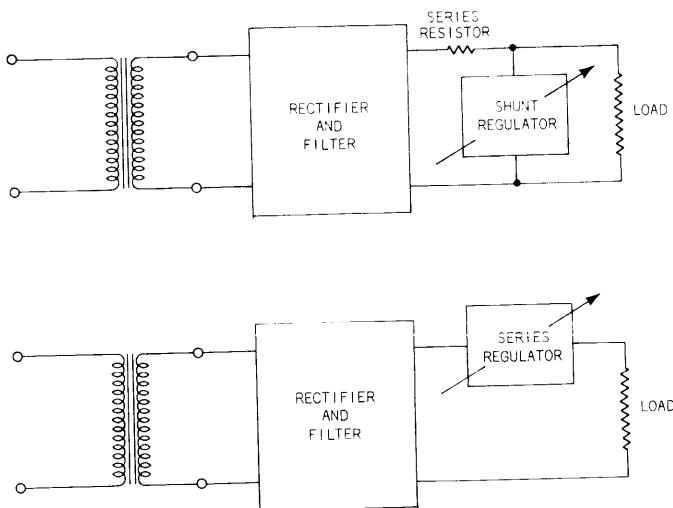
Fig. 2-1. Voltage regulators.

Two classes of voltage regulators are found in
conventional AC power supplies. These are the
shunt regulator and the series regulator. Block
diagrams of these types are shown in Fig. 2-1.

variable impedances

The shunt regulator parallels the load, while the
series regulator is in series with the load. Both
regulators act as variable impedances, reacting
automatically to changes in input voltage or load
current demand in such a way that the voltage
across the load remains essentially constant. The
power source, in most cases is a rectifier similar
to one of those discussed in the previous chapter.

shunt regulator

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

series
regulator

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

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

monitoring
circuit

so some type of monitoring circuit must be included
to communicate error information to the regulating
device.

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

It has already been mentioned that a shunt regulator
can theoretically be constructed with a single
device possessing the proper characteristics. Two
such devices, in fact, are available -- the first of
these is the voltage-regulating (VR) tube. This

VR tube

tube, also known as a glow-discharge tube, is a cold-
cathode diode filled with one of the inert gases
such as neon, argon, etc. When a voltage is impressed
on this tube, the gas begins to ionize. As the
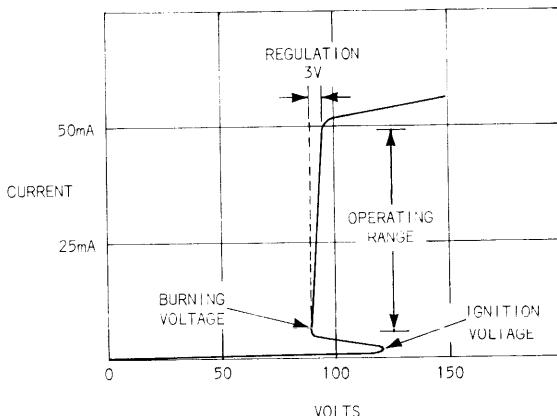
Fig. 2-2. Typical VR-tube characteristic.

ignition
voltage

burning
voltage

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

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

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

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

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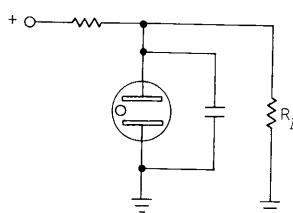

Fig. 2-3. VR-tube noise elimination.
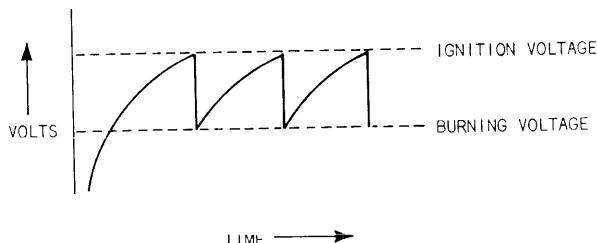
Fig. 2-4. Relaxation oscillation waveform.

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

Still another problem presented by VR tubes is that
the tube's burning voltage, a function of electrode
configuration and gas pressure, cannot be established
at any desired level in manufacture. Typically, gas-
filled diodes are only designable with certain
discrete burning voltages, usually about 75 V, 100 V,
and 150 V for VR tubes. Burning voltages less than
60 V are not obtainable, and the only way to
achieve regulated voltages above 150 V is to
connect two or more VR tubes in series. This can
be done, providing certain precautions are observed.
Fig. 2-5 illustrates the problem and its solution.
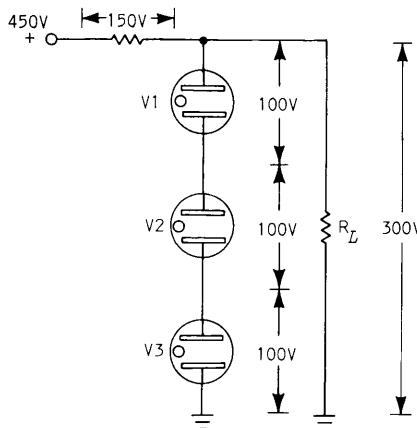
Fig. 2.5.   VR tubes in series.

series
VR's

When two or more VR tubes are connected in series,
the total voltage required to cause them to ignite
(assuming that they have identical characteristics
and zero leakage current) is the sum of their
ignition voltages. If, for instance, V1, V2, and V3
are 100-V VR tubes with 150-V ignition voltages,
the power-supply output voltage must be at least
450 V. Once the tubes have ignited, their burning
voltages will add to about 300 volts, so about
150 volts must be dropped across the series resistor.
This means that at least one-third (in practice,
nearer one-half) the source power is dissipated in
the series resistor -- not a very useful arrangement.
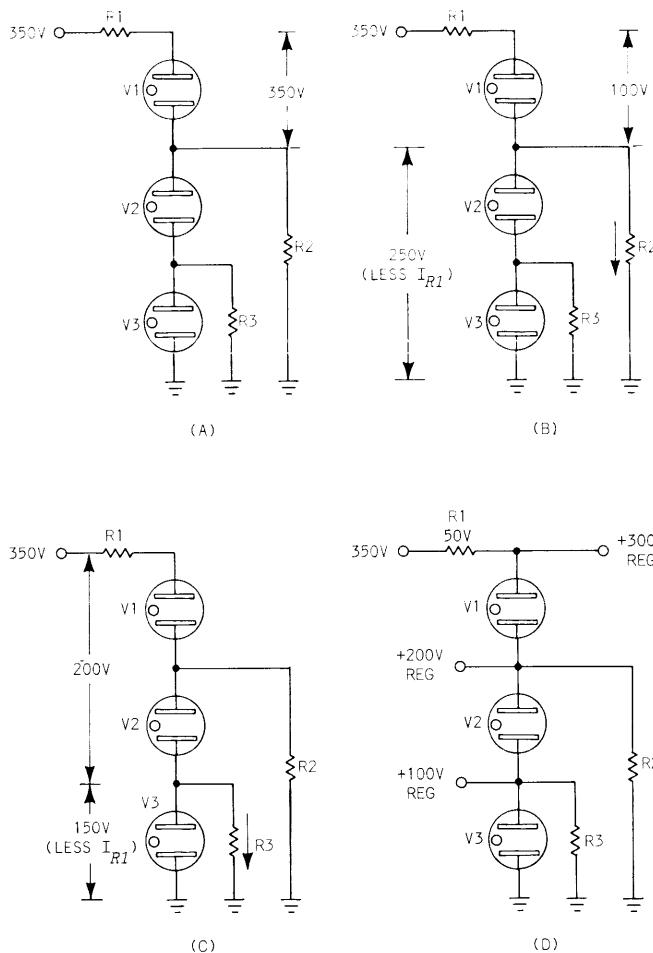
Fig. 2-6. Operation of practical series
VR-tube regulator.
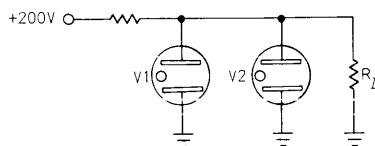
Fig. 2-7. Parallel VR tubes.

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

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

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

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

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

stable
reference
voltage

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

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

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

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

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

Two distinct mechanisms are responsible for the characteristics of the zener diode. The first, "zener" breakdown, occurs between about 2 and 7 volts, and resembles the phenomenon known as "quantum tunneling" exhibited by the tunnel diode. The other mechanism, which occurs between about 5 and 200 volts, resembles the phenomenon known as "avalanche" breakdown, which occurs in avalanche transistors. Diodes with reverse breakdown characteristics in this region might more accurately be called avalanche diodes, but for convenience are included in the zener-diode category.
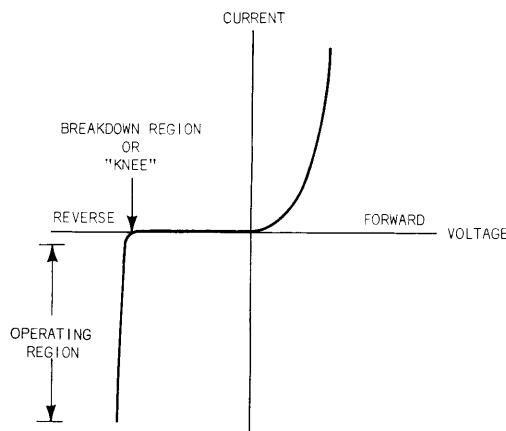
Fig. 2-8. Typical zener-diode characteristic.

temperature
coefficient

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

temperature
compensation

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

Zener diodes can handle large currents, for practical purposes limited only by their ability to dissipate power in the form of heat. Heat sinks, which conduct much of the heat to the chassis where it is radiated to the surrounding air, increase the power-handling capacity of zener diodes to more than 100 watts. Thus the diode can be used as a shunt regulator for high load currents, especially at lower voltages. Load currents of several amperes are not unusual.
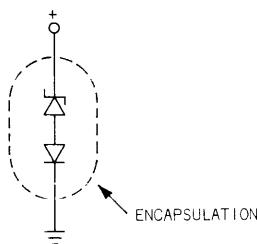
Fig. 2-9. Zener-diode temperature compensation.
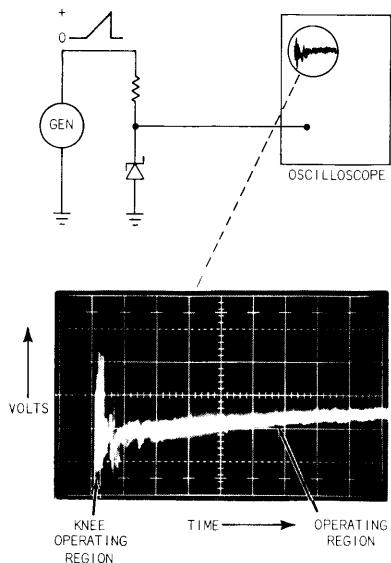
Fig. 2-10.   Zener-diode noise.

zener
noise

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

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

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

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

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

Most of the shunt regulators employed in Tektronix
instruments are single device types similar to those
discussed above. Their application is limited to
those infrequent situations where economy or extreme
accuracy are the ruling factors.
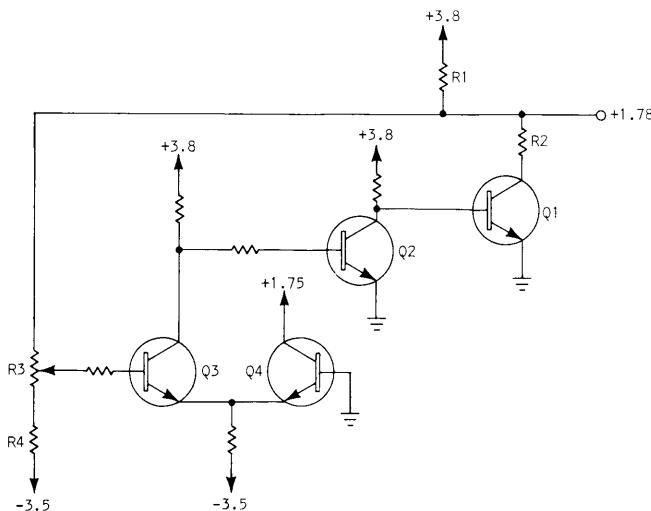
Fig. 2-11. Shunt voltage regulator.

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

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

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

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

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

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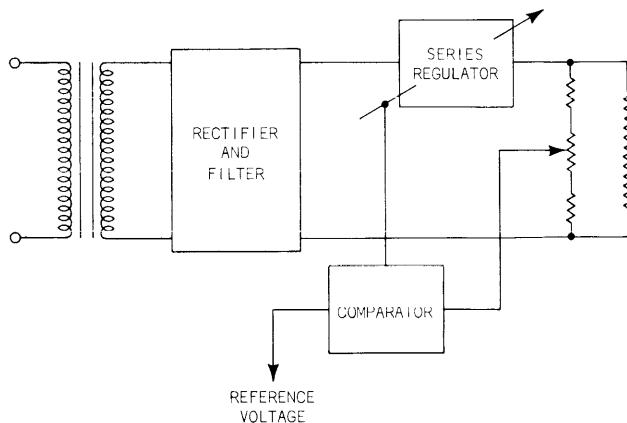

Fig. 2-12. Block diagram of series voltage
regulator.

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

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

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

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

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

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

The grid of the regulator tube, V1, must now be
controlled by the output of the comparator. To
determine the required polarity of the corrective
voltage it is only necessary to assume some change
in the load current. For instance, if load current
suddenly increases, the output voltage would tend
to drop, due to the increased voltage drop across V1.
To offset this tendency, V1 must allow more current
to flow, which in turn requires a positive signal
on its grid.
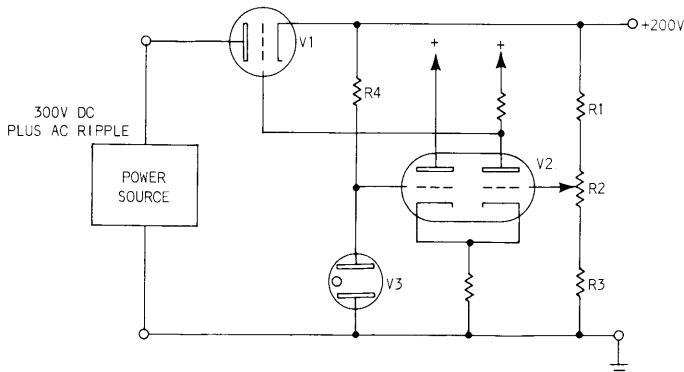
Fig. 2-13. Basic circuit of series regulator.

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

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

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

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

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

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

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

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

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

power-
supply
impedance

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

Obviously, considerable improvement is necessary before this regulator can be considered a practical circuit. Nevertheless, it serves to illustrate the basic mechanisms at work in all series regulators. Before going on to the task of improvement, however,

line-
voltage
variations

it will be advantageous to examine the behavior of this relatively crude regulator as it responds to changes in source voltage.

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

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

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

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

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

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

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

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

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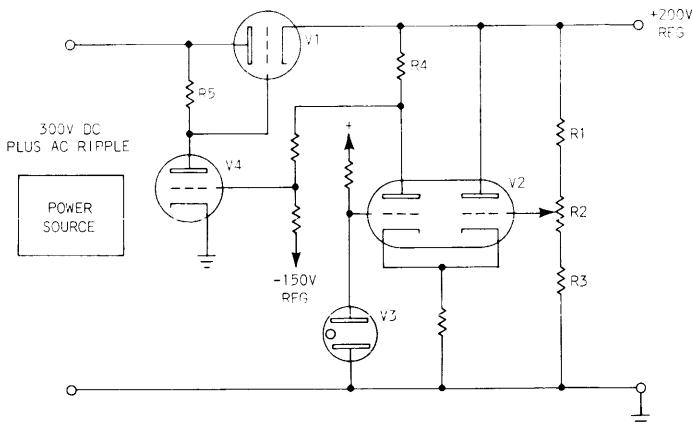
Fig. 2-14. Series regulator with triode error
            amplifier.

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

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

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

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

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

In order that V4 act as a pentode, its screen grid
must be connected to a constant voltage source; that
is, the screen grid-to-cathode potential must not be
allowed to fluctuate in sympathy with the error signal.
The regulated output voltage provides a convenient
source, and is brought to the desired level by a
voltage divider (R7-R8) connected to signal ground.
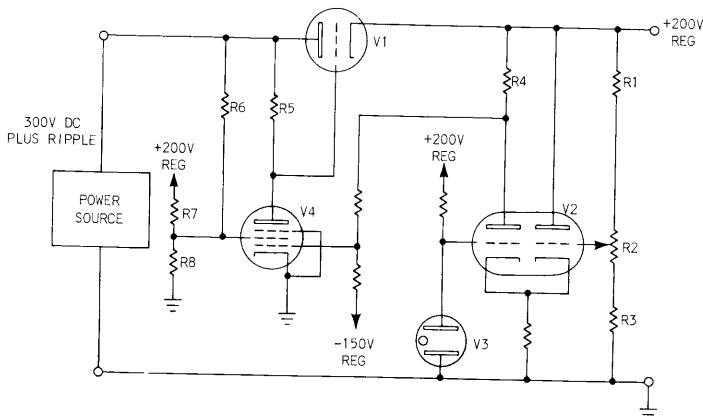
Fig. 2-15. Series regulator with pentode error
amplifier.

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

With the parameters of the pentode established,
consider the signal applied to the screen grid. In
this case it is the unregulated supply voltage, or
more accurately, *changes*, in the supply. The screen
grid is usually supplied from a three-resistor
network. Two of the resistors, R7 and R8 constitute
the voltage divider which establishes the DC level
of the screen-grid bias voltage. The other resistor,
R6, usually has a rather high value and is connected
to the unregulated supply, feeding a sample of any
change in supply voltage to the screen grid.
(Because both R7 and R8, are, in effect, connected
to signal ground, only R6 communicates changes in
the supply voltage to the screen grid.) The sample,
of course, is only a small fraction of the actual
supply voltage change, due to the attenuation
$\dfrac{\frac{R7 \cdot R8}{R7+R8}}{R6}$. If this attenuation is represented by K,
the amplified signal at the plate of V4 will be
$E_{reg}KA$, where A represents the amplification factor.

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

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

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

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

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

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

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


BUFFER CAPACITORS

SPEEDUP CAPACITORS
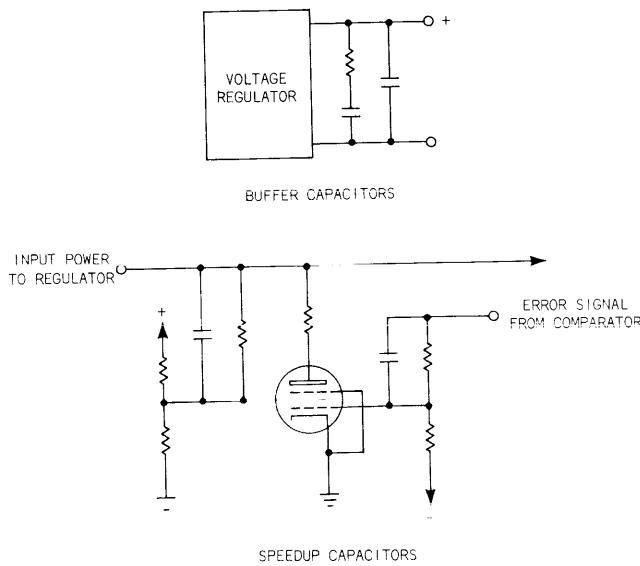
Fig. 2-16. Improving regulator frequency
response.

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

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

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

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

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

Because gain in the feedback loop is such an
important consideration, it is quite natural to
suppose that a pentode, used as a series regulator,
would further improve the regulator's performance.
The essential difference between a pentode and a
triode as a series regulator is that the pentode's
plate voltage can vary quite widely for a given
value of negative bias, without affecting plate
current to any significant extent. This means, of
course, that a pentode will provide regulation
against changes in line voltage much superior to
that of the triode. For this reason, pentodes are
pentode   used as series regulators in two situations:
series   (1) When optimum regulation against line voltages
regulators   is desired and the load current demand is fairly
constant, and (2) When the range of line voltages
from which the power supply must operate is very
wide. This situation occurs primarily in bench-
type power supplies, where a specified regulated
output current must be maintained over a line-
voltage range of several hundred volts. Assuming,
for example, an input-voltage range of 400 volts,
a triode-series regulator would require approximately
a 200 volt change on its grid to keep the output
current constant. This is obviously impossible for
the triode; however, a pentode would require only a
few volts of change in grid voltage to handle the
same conditions.

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

only to a negligible extent, a factor of electrode
structure. Above a certain envelope temperature,
there is a likelihood that occluded gas will be
released from the glass and produce a "gassy" tube.

heat
dissipation

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

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

plate
dissipation

current rating. It is then possible to determine the
plate-dissipation limitations of the tube.

As an example, consider a +200-volt power supply
with an output current of 100 mA and a typical input
voltage-regulation range of $\pm10\%$. The lowest
voltage which can be permitted across the series
tubes to maintain the desired output current will be
approximately 50 volts. The rectifier output voltage
therefore must go no lower than 250 volts if regulation
is to be maintained. This represents the -10%
condition. If 250 volts is 90% of the nominal
voltage, under high-line conditions this voltage will
rise by a factor of 2/9 to 305 volts. At this time
the plate-to-cathode voltage will be 105 volts and
the plate dissipation will be 10.5 watts.

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

shunt resistor

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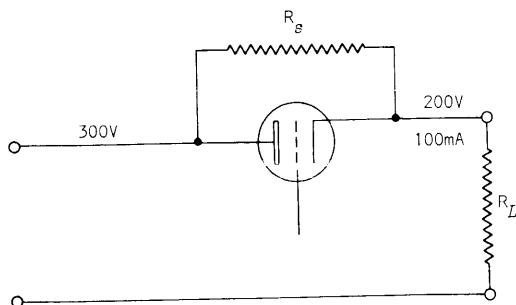
Fig. 2-17. Series tube shunt resistor.

Consider first the effects of load-current changes.
It has already been shown that as long as load
current remains within the regulating capability of
the circuit, the cathode voltage will vary only a
few millivolts with load-current changes. The plate
voltage will also vary a few volts, due to the finite
impedance of the transformer and rectifier; however,
as a first approximation it can be said that the
voltage across the series tube remains essentially
constant. Since this same voltage appears across
the shunt resistor, the current through the shunt
resistor will also be constant. Therefore, as far
as load-current changes are concerned, the shunt
resistor plays no part in the circuit; changes in
load current are matched by equal changes in the
cathode current of the series tube, just as if the
shunt resistor were not there. If, as a further
simplification, the bias characteristics of the
series tube are assumed to be linear, the change in
bias voltage will be equal to the change in load
current, divided by the g_m of the series tube, or
$\Delta E_g = \frac{\Delta I_l}{g_m}$. Thus with a 20-mA change in load current
and a series tube whose g_m is 5 mA/volt, a 4-volt
change on the grid will be required. The change in
output voltage will therefore be 4 volts divided by
the combined gain of the comparator and error
amplifier.

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

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

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

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

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

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

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

excessive
load-current
changes

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

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

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

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

excessive
supply-
voltage
changes

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

ripple
peaks

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

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

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

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

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

SWEEP SPEED: 5ms/div
VERTICAL LENS: 0.1V/div
NOMINAL LINE VOLTAGE: 120V AC
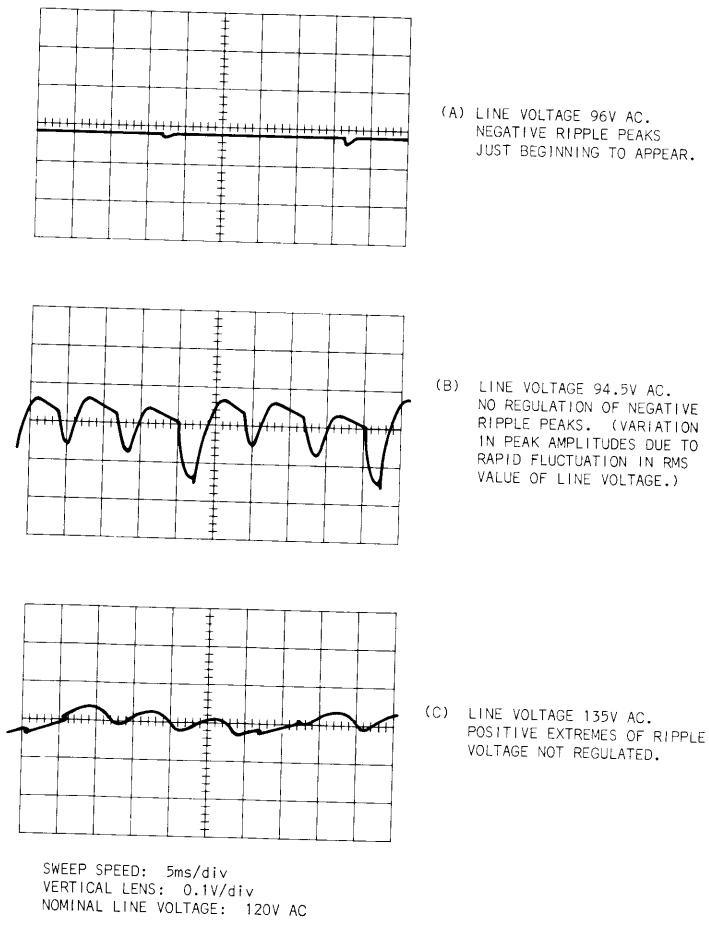
Fig. 2-18. Effect of regulation loss on CRT
display.

negative-
output
voltage
regulators

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

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

This configuration dictates a number of polarity
changes in the error-signal feedback loop. An
increase in load current or decrease in supply voltage
causes the monitor to detect a less negative, or
positive signal. The grid of the series-regulator
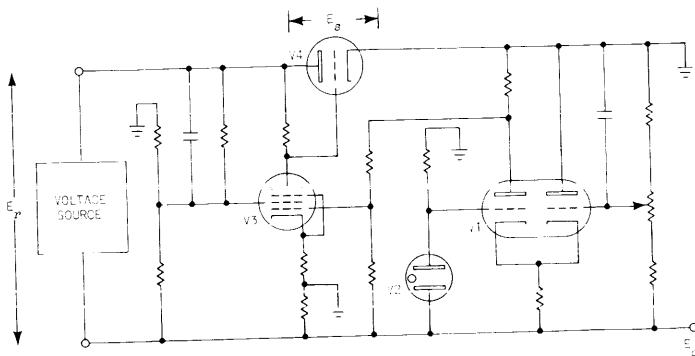
Fig. 2-19. Series-regulator negative output.

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

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

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

It should be noted that, contrary to the views expressed in many textbooks, no advantage other than convenience in DC-biasing is gained by this configuration. A quick examination of the circuit, using arbitrarily chosen values of voltage and resistance, will prove that the net difference between grids of the comparator is the same whether the VR tube is connected between grid and ground or grid and output.
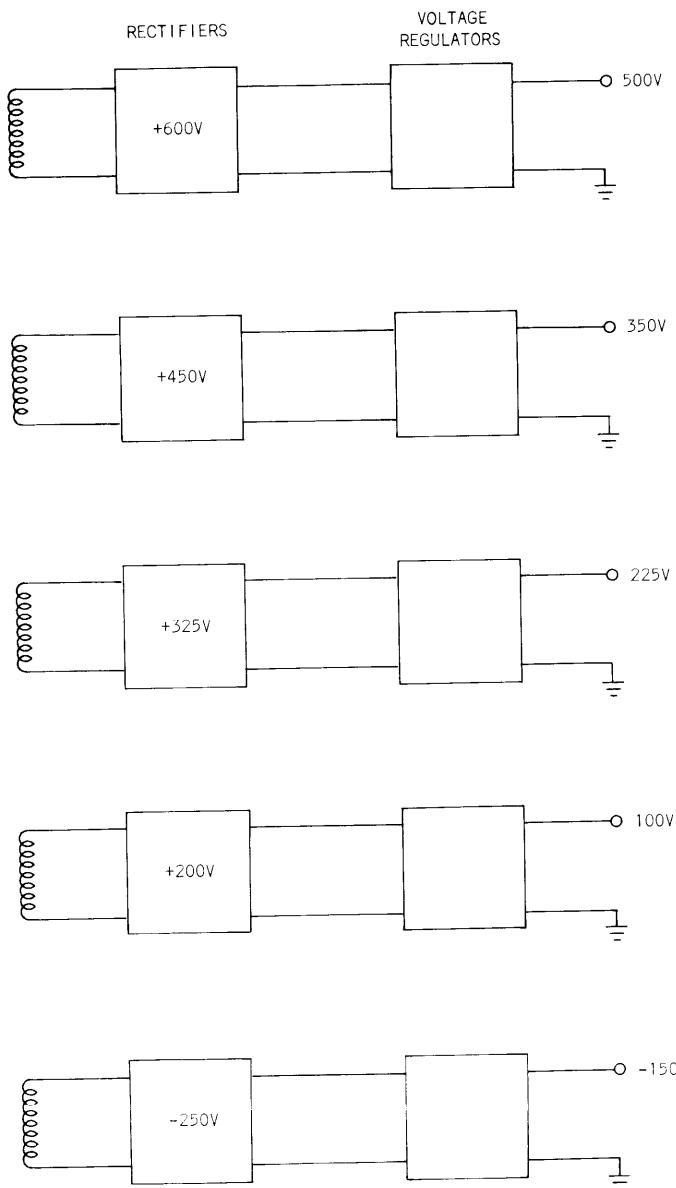
Fig. 2-20. Unstacked power supplies.

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

It might well be asked why the unregulated, rather
than the regulated outputs are stacked. The chief
disadvantage of the latter system is that all the
load current in the more positive supplies must pass
through the series regulators of the less positive
supplies. Thus all the load current drawn by the
instrument must eventually pass through the series
regulator of the +100-volt supply. This would
require a very expensive series tube in a vacuum-tube
regulator; in any case, considerable power would be
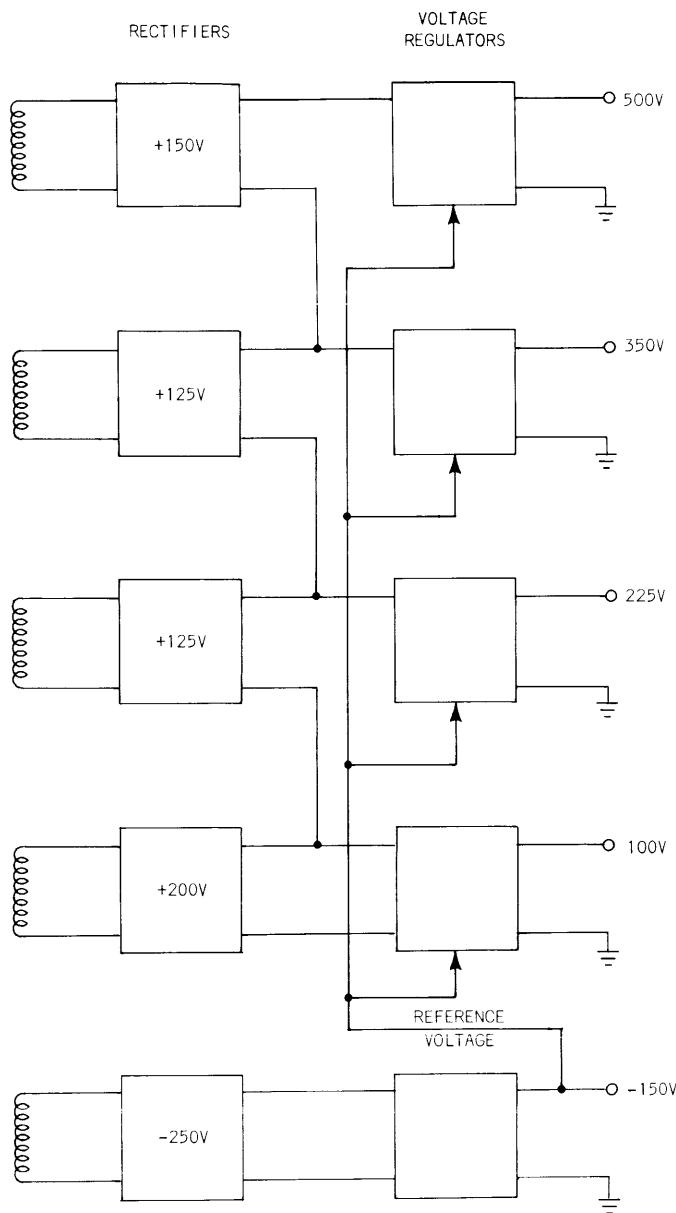
Fig. 2-21. Stacked power supplies.

dissipated in the series regulating device whether vacuum tube or semiconductor. Occasionally, however, it *is* advantageous to stack one power supply on the regulated output of another. First, compare the unregulated output of the isolated 500-volt power supply in Fig. 2-20 with that of the corresponding stacked power supply shown in Fig. 2-21. To simplify the problem, assume that all the unregulated outputs under discussion exhibit 10% ripple. The isolated power supply, with an unregulated output of 600 volts will have 60 volts of ripple. The stacked power supply also has 60 volts of ripple, since all the ripple is in phase and therefore *adds*. ($\frac{200}{10} + \frac{125}{10} + \frac{125}{10} + \frac{150}{10} = \frac{600}{10} = 60$.) Obviously, no reduction in AC ripple is effected by stacking unregulated outputs. However, if, as in Fig. 2-22, a 250-volt unregulated supply is stacked on the 350-volt regulated supply, the unregulated 600-volt output will have only 25 volts of ripple (10% of 250 volts). This technique permits a much simpler design in the series regulator, and can be employed whenever the load current of the higher voltage supply is low enough to prevent overloading the series regulator in the lower supply.
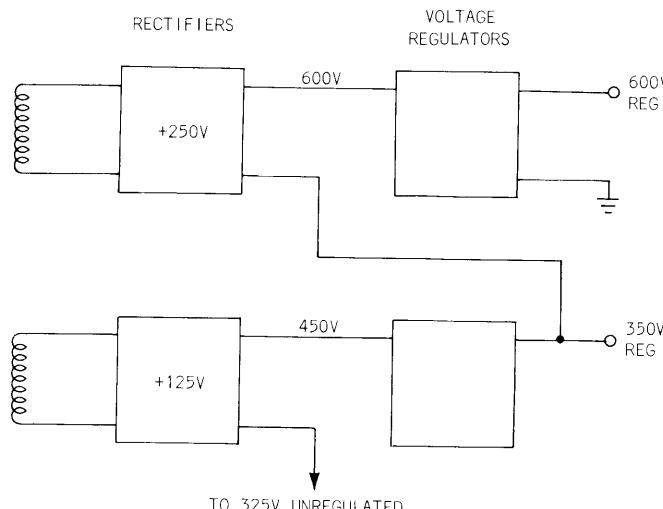
Fig. 2-22. Stacking on regulated output.

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

voltage-
reference
requirements

With a number of power supplies in the same instrument,
there is no point in using a VR tube or zener diode
as a voltage-reference source in each supply. It is
sufficient that one voltage-reference component be
used to establish the reference for one supply; the
regulated output of this supply can then be used as
the voltage reference for all the other supplies
(Fig. 2-21). In Tektronix instruments the voltage-
reference component is usually found in the most
negative supply. The regulated negative voltage
then provides the reference for the other supplies.
This permits some sweeping simplifications in the
circuits of the positive supplies. For example,
examine the typical comparator circuit in Fig. 2-23.
Note that the VR tube has been removed and the left-
hand grid of the comparator has been connected to
ground. The monitoring voltage divider, however, is
now connected between the regulated positive output
and the regulated negative voltage. Obviously there
exists some point on the divider at which the
potential is zero. When this point is connected to
the right-hand grid of the comparator, an error signal
will be generated only when the monitor voltage
departs from zero. Since one side of the voltage
divider is connected to the regulated negative output,
only changes in the positive supply's output will
produce an error signal.
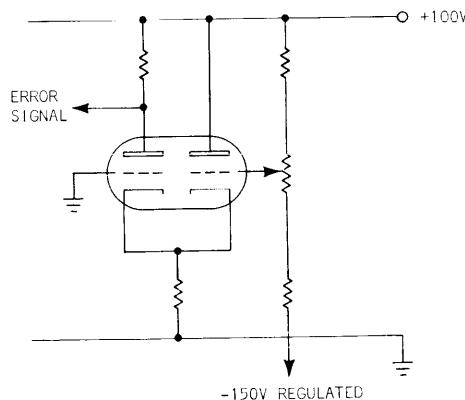
Fig. 2-23. Simplified comparator circuit.
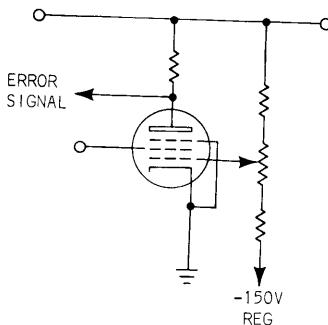
Fig. 2-24. Simplified control circuit.

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

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

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

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

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

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

It is now possible to predict the behavior of the
original circuit under a variety of conditions. Of
particular interest are those situations in which
(1) R_{th} is much larger than the load resistance, and
(2) R_{th} is much smaller than the load resistance.

In the first instance, as the value of R_{th} increases
with respect to that of the load resistance, small
changes in the load resistance have less and less
effect on the current flowing out of the circuit.
As the value of R_{th} increases, the circuit acts more

and more like a "constant-current" source. While a
true constant-current source can exist only in
theory, the concept is quite useful in analyzing the
operation of certain circuits.

In the second case, as the value of R_{th} decreases with
respect to that of the load resistance, small
variations in the load resistance have a decreasing
effect on the output voltage of the circuit. If the
value of R_{th} could be reduced to zero, the circuit

would be a "constant-voltage" source. Again, although
such a circuit can exist only in theory, the concept
is useful in explaining the behavior of certain
circuits.
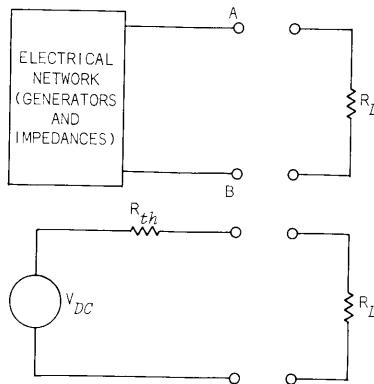
Fig. 2-25.   Thevenin equivalent circuit.
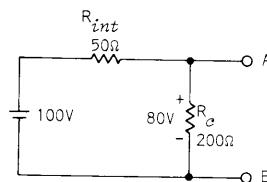

Fig. 2-26.   Thevenizing a simple circuit --
             step 1.
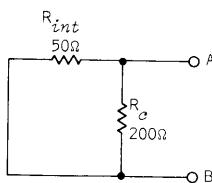
Fig. 2-27.   Thevenizing a simple circuit --
             step 2.
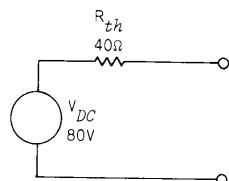

Fig. 2-28.   Thevenizing a simple circuit --
             step 3.
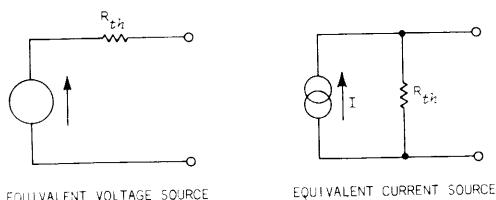
Fig. 2-29.   Equivalent sources.
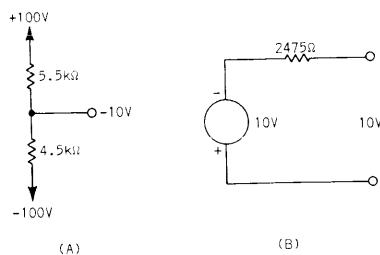
Fig. 2-30.   Thevenized voltage divider.
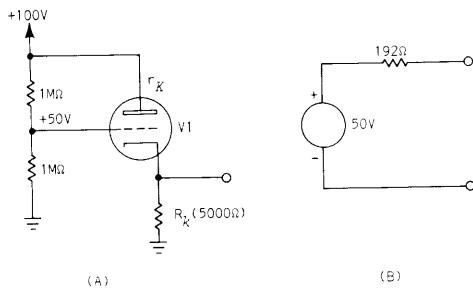
Fig. 2-31.   Thevenized vacuum-tube circuit.
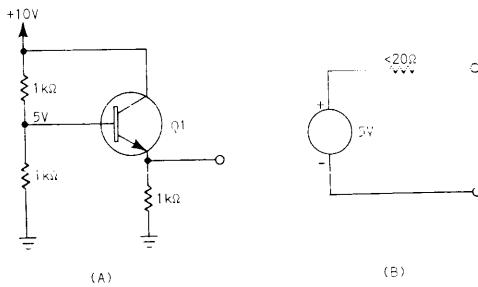
Fig. 2-32.   Thevenized transistor circuit.

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

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

In the example above, a lower output impedance can be obtained by lowering the value of the voltage-divider resistors. The disadvantage of this procedure, however, is that large amounts of current would inevitably be wasted in the form of dissipated heat. A more satisfactory technique is illustrated in Fig. 2-31. A very high-impedance voltage divider is used to develop the grid bias for cathode-follower V1. Assuming a typical g_m of 5000 μmho for V1, the equivalent circuit becomes that shown in Fig. 2-31B. Looking back into the circuit from the output terminals (V1's cathode and ground), the circuit's impedance is equal to the parallel combination of V1's internal cathode impedance r_K and its external cathode impedance R_k. The open-circuit voltage of the equivalent circuit will be about 50 volts. Since $r_K = \dfrac{1}{g_m}$, the internal cathode impedance becomes 200 ohms. The impedance of the equivalent circuit is therefore $\dfrac{200 \cdot 5000}{200 + 5000} = 192$ ohms.

emitter
follower

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

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

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

operational
amplifier

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

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

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

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

Before calculating E_{out2} it is necessary to calculate
the Thevenin equivalent of the second input circuit.
The result (assuming that the pickoff of potentiometer
is set midway between its extremes) is +12.5 V in
series with 6.925 megohms of resistance. After once
more subtracting the -3 volts at the + input from
E_{in2} and multiplying the results by the R_f/R_{in} ratio,
an output voltage (E_{out2}) of -0.22 V is found. The
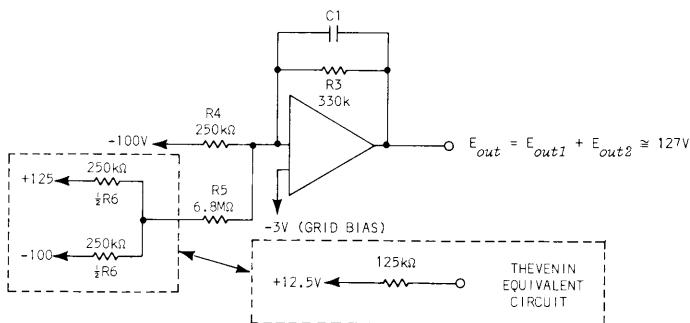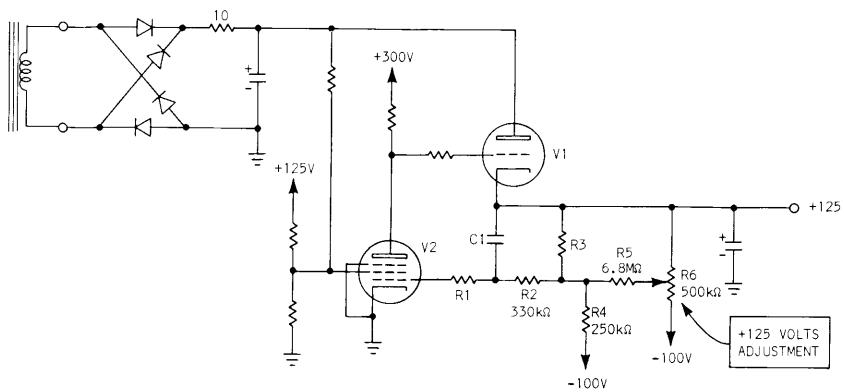
Fig. 2-33. Series regulator -- operational-
amplifier configuration.
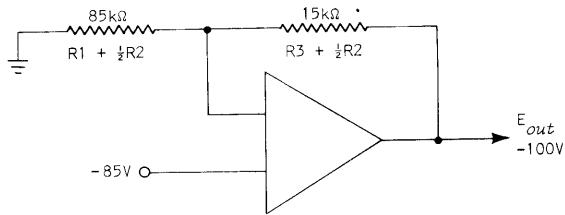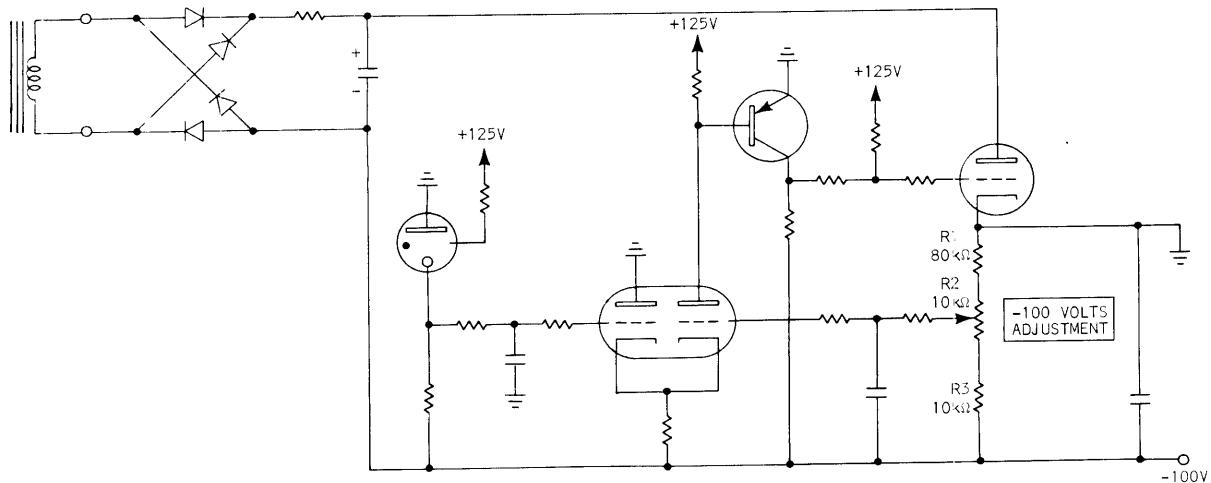
Fig. 2-34. Negative-output operational-amplifier configuration.

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

The operational-amplifier configuration for a typical
negative-output series regulator is shown in Fig. 2-34.
Feedback resistance R_f is the sum of R1 plus 1/2 R2
when potentiometer R2 is centered, while R_{in} is the
sum of R3 plus 1/2 R2. The input voltage is -85,
developed across the VR tube. The operational
amplifier is thus a noninverting type, with a gain
greater than one. The output voltage is $E_{in}(\frac{R_{in}+R_f}{R_i})$
or $-85 \ (\frac{85 \ \text{k}\Omega \ + \ 15 \ \text{k}\Omega}{85}) = -100$ V.

transistor-ized voltage regulators

Although the operating principles of transistorized
voltage regulators are substantially the same as those
of the vacuum-tube type, a much wider variety of
configurations is possible. This is due principally
to the existence of both PNP and NPN transistors.
In essence, the transistor becomes a bidirectional
device. It is thus possible to construct a negative
power supply with the series-regulating device in
the output side of the regulator -- a configuration
already shown to be impossible with vacuum tubes.
Because of this flexibility, an attempt to describe
all the possible variations in transistorized-power-
supply design would obviously be impractical. However,
it is possible to point out the essential differences
between vacuum-tube and transistor-circuit designs.
These differences are dictated solely by the
characteristics of the transistors used in the circuit.
The discussion will be largely confined to NPN
circuits, mainly to emphasize the analogy between
transistors and vacuum tubes.
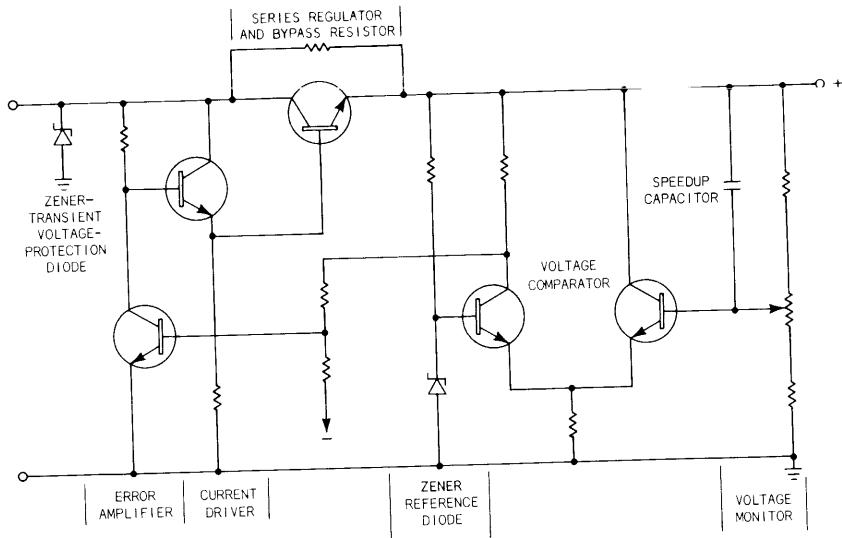
Fig. 2-35. Transistorized series regulator.

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

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

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

The same current-driving problem is encountered in
the voltage-comparator circuit. In a vacuum-tube
regulator, the voltage-divider monitor has a
resistance on the order of several hundred kilohms.
In a transistorized regulator, a much lower impedance,
say 5 to 10 kilohms, must be used so that the base-
driving current for the comparator represents only
a small fraction of the total current in the voltage
divider. The same consideration applies to the value
of the speedup capacitance. To be effective at all,
the capacitor impedance must be small compared to
the divider impedance at frequencies at or above the
ripple frequencies. Since the resistance of the
voltage divider is 30 to 50 times smaller than those
in vacuum-tube regulators, it follows that the values
of the speedup capacitors must be 30 to 50 times
larger. Thus, where a 0.1-μF capacitor would be
encountered in a typical vacuum-tube regulator, a
3 to 5-μF capacitor is required in a transistorized
supply. Only electrolytic capacitors can provide
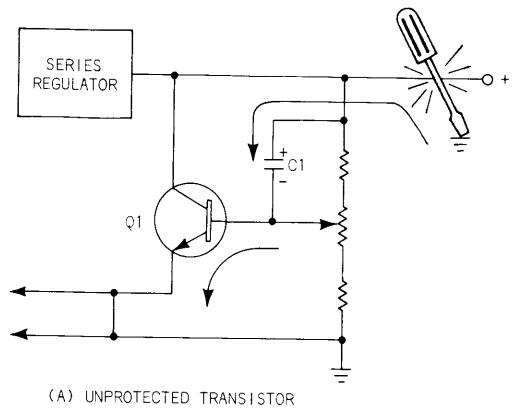
these high values without becoming excessively
bulky. To minimize the leakage problems associated
with the electrolytic type, tantalum capacitors are
usually employed in this application. (In trouble-
shooting a power supply it is often wise to disconnect
the speedup capacitor. If the output voltage
changes, the capacitor is leaky and should be replaced.)
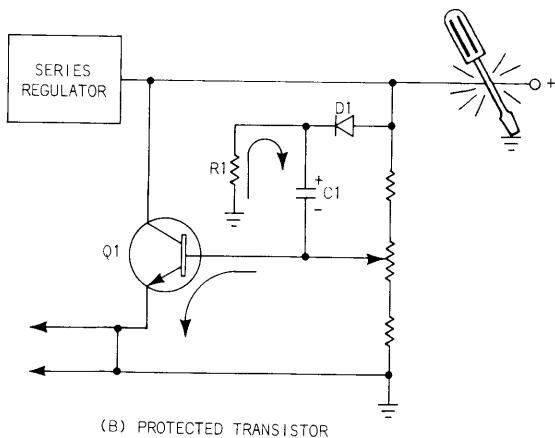
The presence of the speedup capacitor introduces
another problem not encountered in vacuum-tube
regulators. (Fig. 2-36A.) If the output of the
regulator is accidentally short circuited, the voltage
at the top of capacitor C1 instantaneously drops to
ground potential. The voltage across the capacitor
cannot change instantaneously, however, so the
potential at the base of Q1 (one half of the
comparator) is driven negative by an amount
approximately equal to the amplitude of the original
supply voltage. This usually exceeds the base-to-
emitter reverse-breakdown tolerance of any transistor
likely to be used in the comparator circuit. The
transistor will therefore suffer irreversible damage,
unless it is deliberately protected from this
eventuality.

protective
devices

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

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

(A) UNPROTECTED TRANSISTOR

(B) PROTECTED TRANSISTOR

Fig. 2-36. Short-circuit protection.
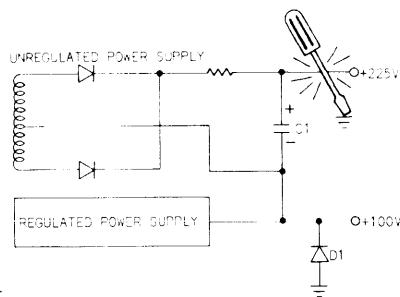
Fig. 2-37.   Short-circuit protection.
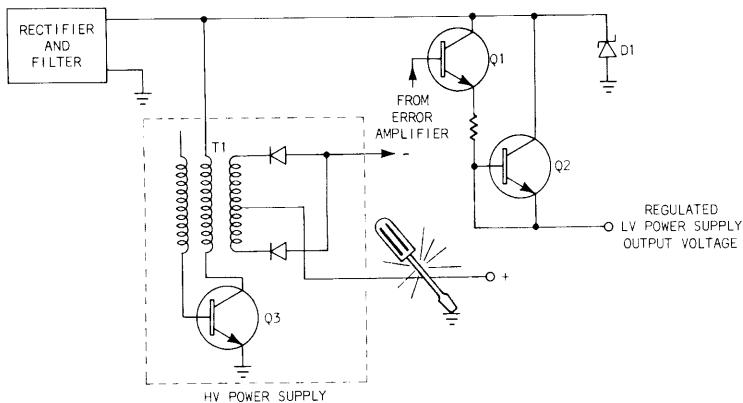

Fig. 2-38.   Short-circuit protection.
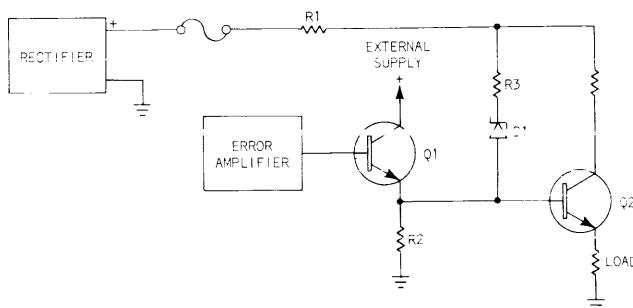
Fig. 2-39.   "High-line" protection by zener diode.

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

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

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

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

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

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

temperature
compensation
Although temperature compensation is not a protective
technique in the strictest sense, it does give rise
to a number of odd configurations which are found
almost exclusively in transistorized power supplies.
One example of this technique has already been
encountered in connection with temperature-compensated
voltage-reference diodes.
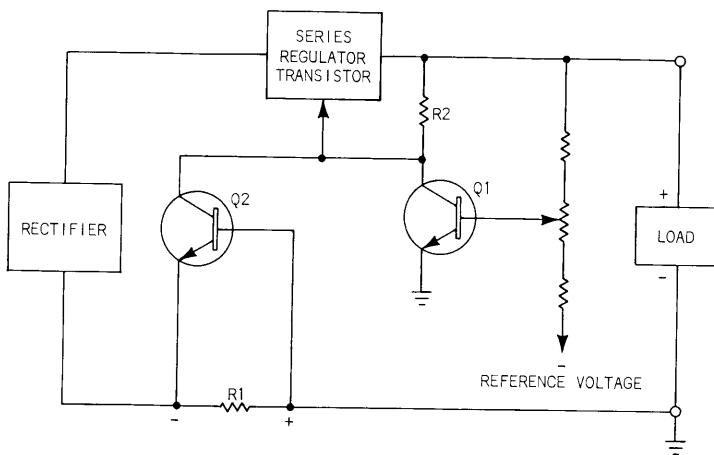
Fig. 2-40. Current-overload protection.
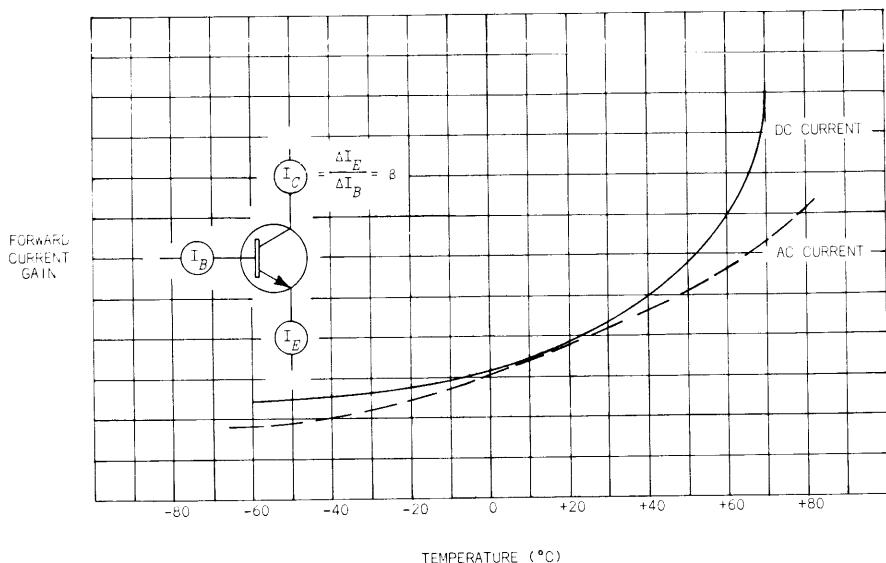
Fig. 2-41. AC-DC current gain versus temperature.

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

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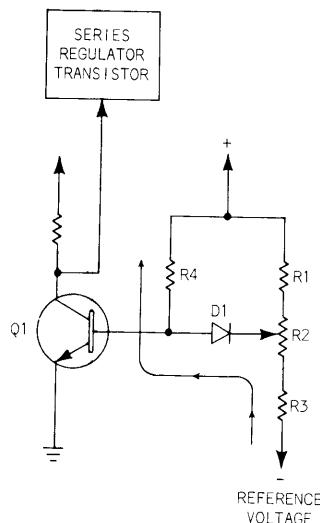
Fig. 2-42. Temperature-compensated transistor.

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

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

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

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

NOTES
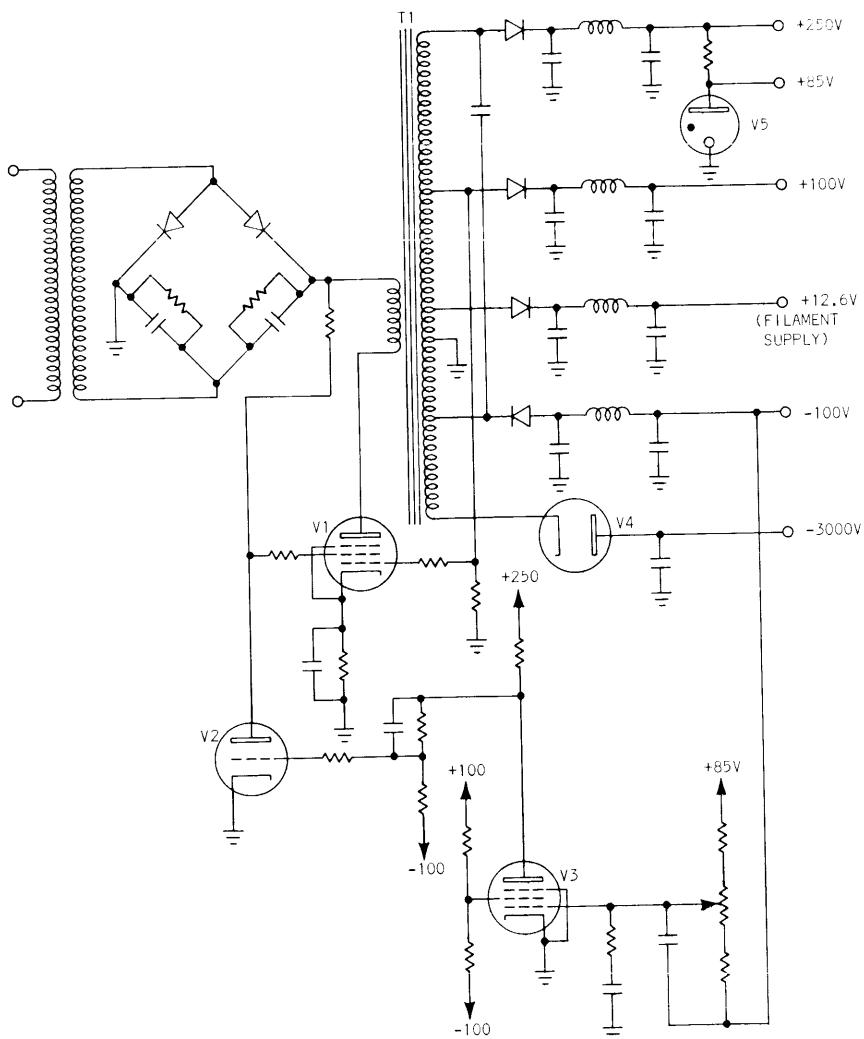
Fig. 3-1. PRLV power supply.

3

THE PRLV POWER SUPPLY

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

primary
power

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

regulation

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

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

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

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

4

AC—DC POWER SUPPLIES

portable-
oscilloscope
character-
istics

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

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

With the advent of solid-state technology came the
first real opportunity to create a truly versatile,
high-performance, portable oscilloscope. Before such
an instrument could be realized, however, a new type
of power supply had to be developed; for while
conventional power supplies could be reduced in size

power-
supply
efficiency

and weight to a striking degree they had one
disadvantage in common: Poor *efficiency*. That is,
much of the power delivered to the instrument was
wasted in the voltage-regulating circuits themselves.
Although not an important consideration in laboratory
instruments, the *efficiency* of a power supply (its
ratio of input power to delivered power) determines
how long a battery-powered instrument containing that
supply can operate between recharging periods. Thus
the design of battery-powered instruments called for
an entirely new approach.
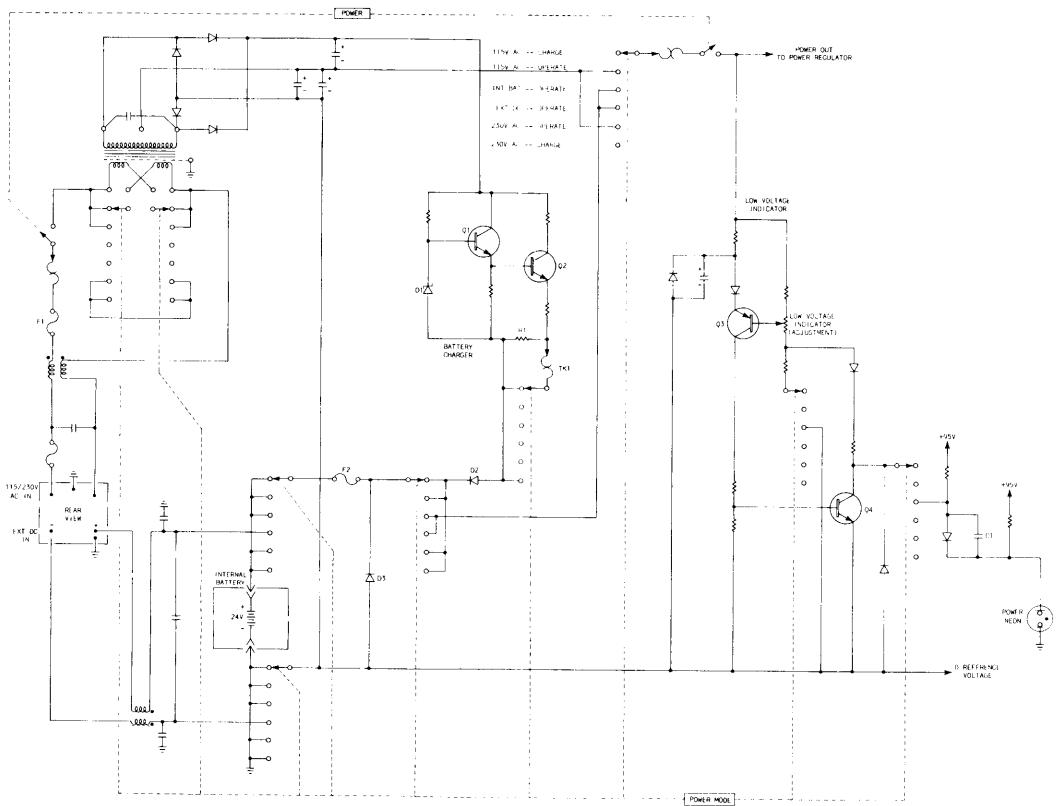
Fig. 4-1. AC-DC selector.

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

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

AC-DC
selector

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

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

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

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

DC-DC converter

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

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

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

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

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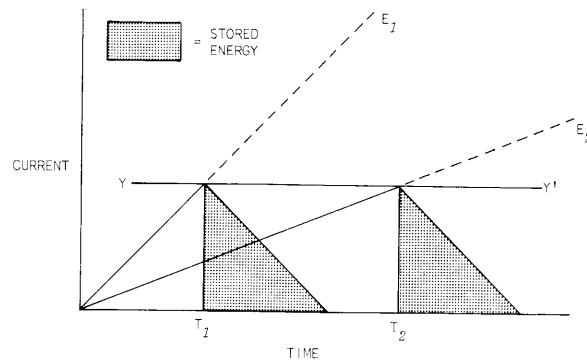

Fig. 4-2. Energy-storage principle.

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

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

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

Fig. 4-4 is a block diagram of the complete DC-DC
regulator. Before examining the details of the
individual circuits it will be advantageous to
establish the function of each circuit as it relates
to the overall operation of the regulator.
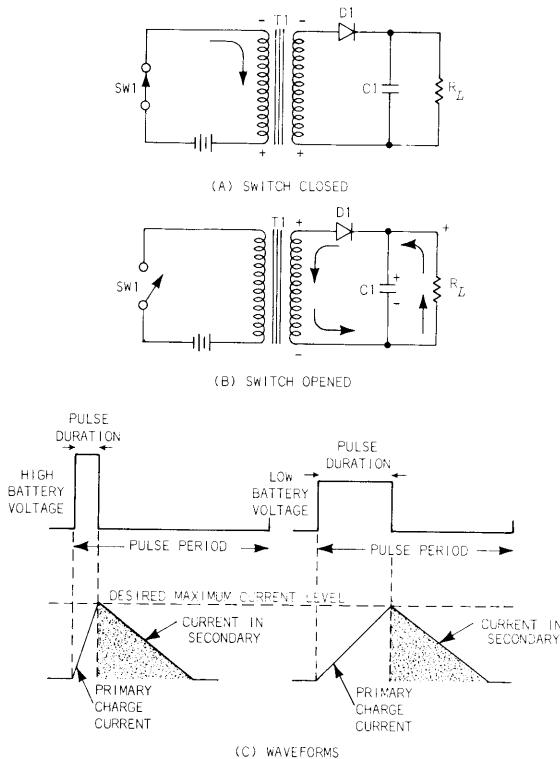
Fig. 4-3. Energy-storage switch -- principles of operation.
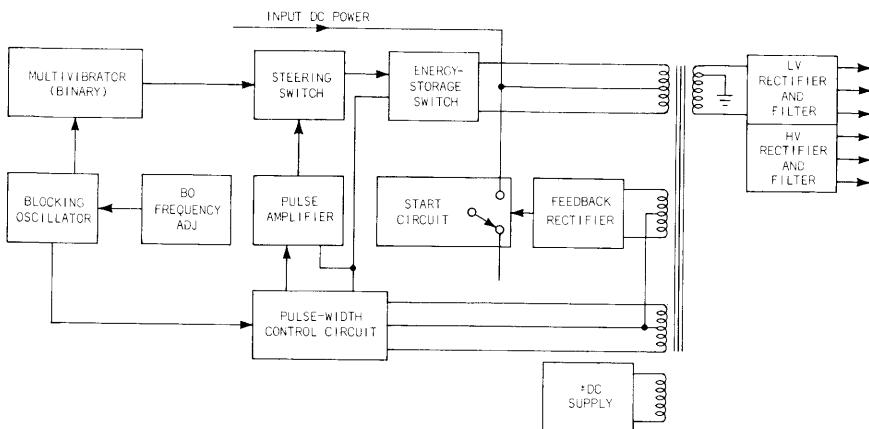
Fig. 4-4. DC-DC regulator -- block diagram.

(A) BO TRIGGER PULSES TO MULTIVIBRATOR

(B) SAWTOOTH PULSES TO PULSE-WIDTH CONTROL CIRCUITS
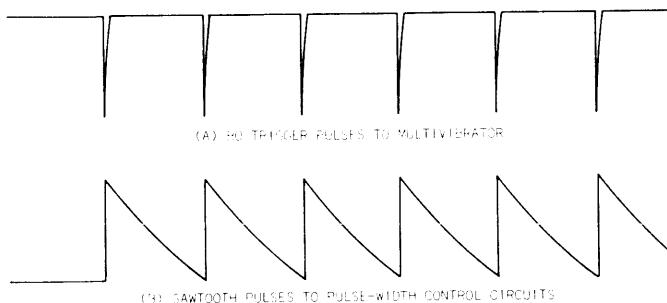
Fig. 4-5. Blocking-oscillator output waveforms.

MAXIMUM BO FREQUENCY = $\dfrac{1}{\text{MINIMUM ALLOWABLE PULSE PERIOD}}$
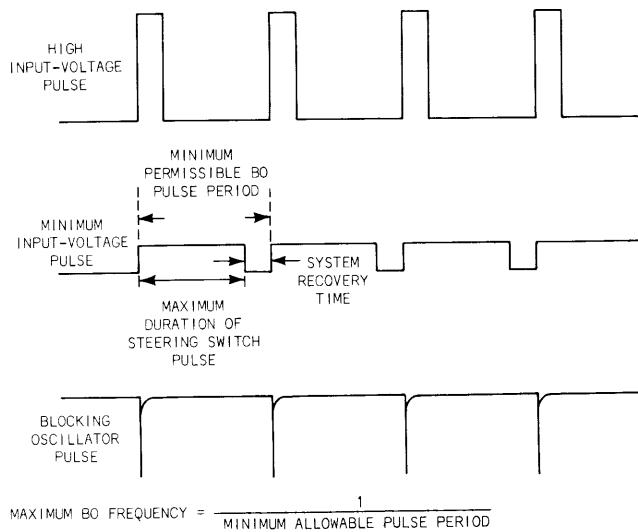
Fig. 4-6. Determination of maximum BO frequency.

(A) BO TRIGGER PULSES

(B) MULTIVIBRATOR OUTPUT
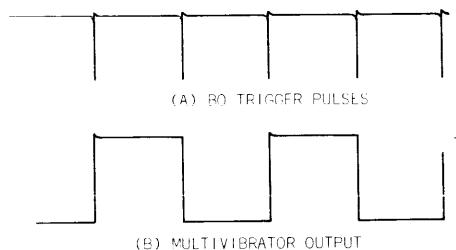
Fig. 4-7. Waveform time-relationship comparison.

blocking
oscillator

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

multi-
vibrator
(binary)

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

steering
switch

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

(A) BO TRIGGERS

(B) MULTI-OUTPUT
    SQUAREWAVE

(C) SIDE-1 STEERING-SWITCH
    OUTPUT PULSES

(D) SIDE-2 STEERING-SWITCH
    OUTPUT PULSES

PULSE WIDTH
DETERMINED BY
PULSE-WIDTH
CONTROL CIRCUITS
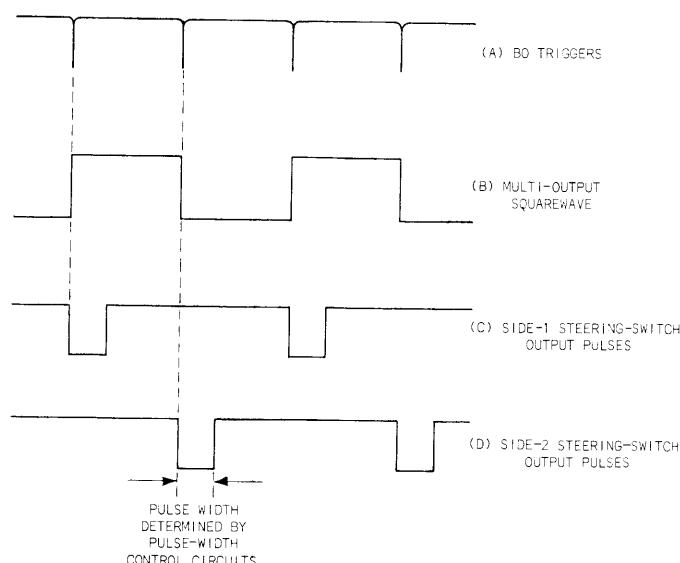
Fig. 4-8. Waveform time-relationship
comparison.

energy-
storage
switch

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

pulse-width
control
circuit

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

plus-and-
minus-DC
supply

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

Since the regulator operating potentials are generated
in the regulator itself, an external DC supply
source must be employed to initiate the regulator's
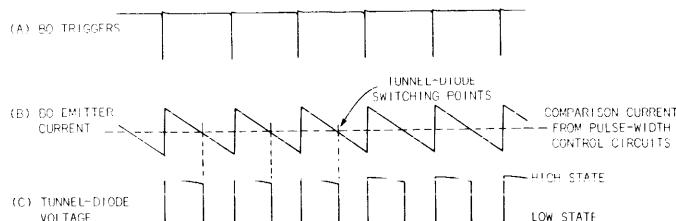
Fig. 4-9. Waveform time-relationship
comparison.

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

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

power-   When power is applied to this circuit Q1 conducts
regulator   through winding 1 of transformer T3, coupling a
blocking   positive pulse through winding 3 to the base of Q1.
oscillator   This regenerative feedback action causes a very rapid
               rise in Q1 collector current which is inductively
               coupled to winding 2 and applied to the multivibrator
               as a negative pulse. When Q1 reaches saturation,
               the magnetic field in winding 1 collapses. A
               negative pulse is thus induced in winding 3, reverse-
   biasing Q1 and cutting off its conduction. The
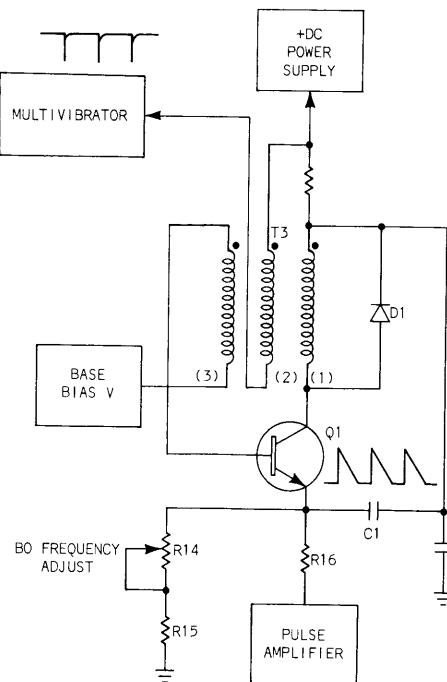
Fig. 4-10. Blocking-oscillator circuit details.

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

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

multi-
vibrator
(binary)

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

The output of the multivibrator consists of two trains of squarewave pulses (in push-pull), whose frequency is one-half that of the blocking oscillator. These pulses are applied as inputs to the steering switch (Fig. 4-12).
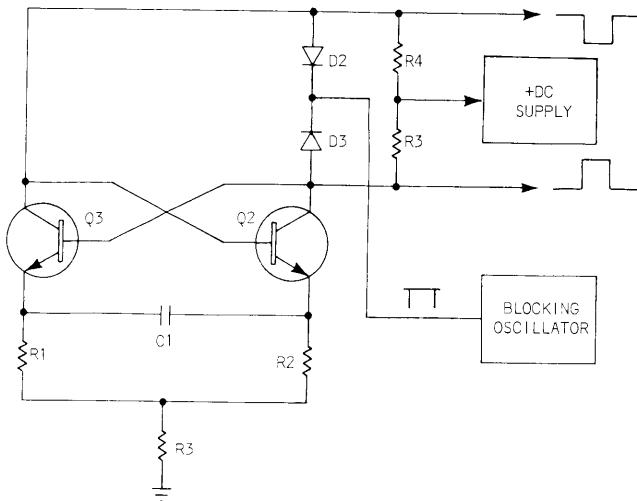
Fig. 4-11. Multivibrator (binary).
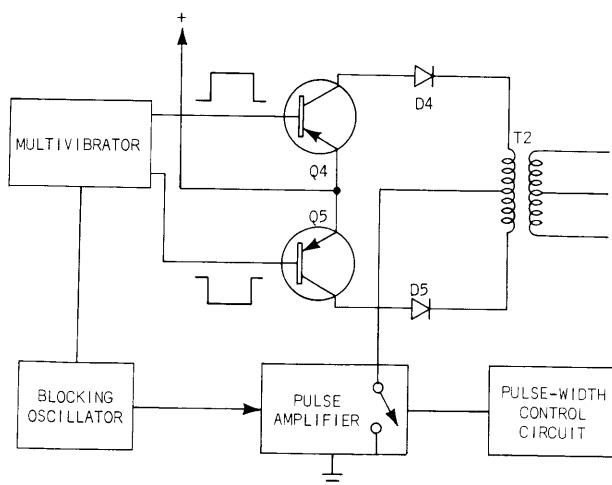

Fig. 4-12. Steering switch.
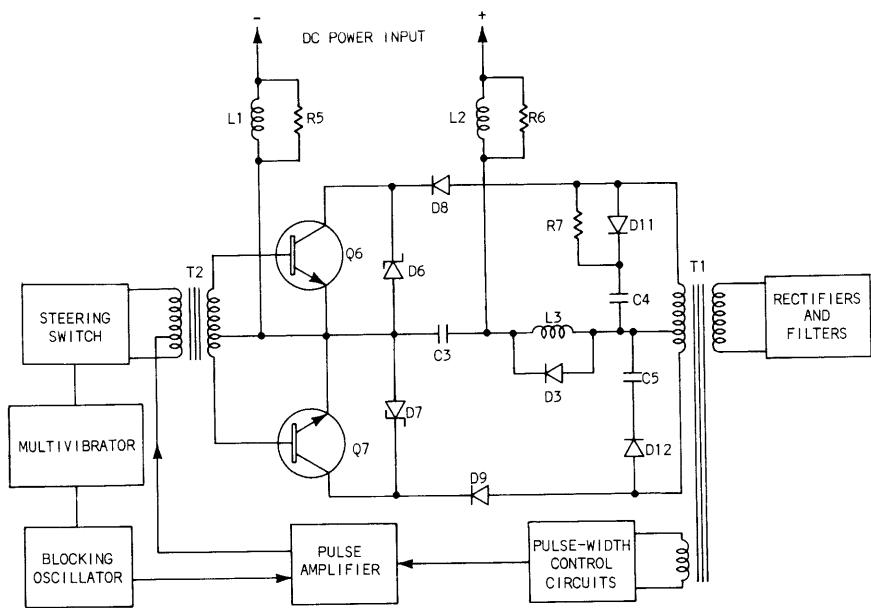
Fig. 4-13. Energy-storage switch.

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

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

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

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

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

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

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

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

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

pulse-
width
control

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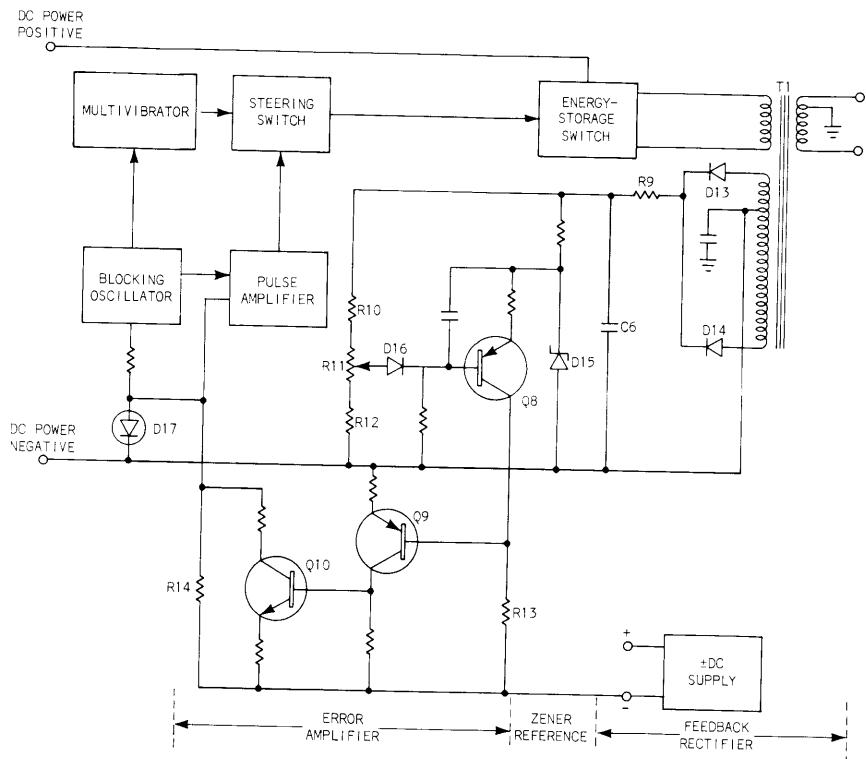
Fig. 4-14. Pulse-width control circuit.

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

feedback
rectifier

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

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

zener
reference

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

error
amplifier

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

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

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

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

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

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

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

When D17 switches to its low state, the negative step
transmitted to the base of Q11 cuts the transistor
off. Its collector voltage rises, cutting off Q12
and opening the circuit between the primary of T2
and ground. The abrupt change in current through T2
turns off the energy-storage switch and the energy
in T1's collapsing primary field is coupled to the
secondary winding.
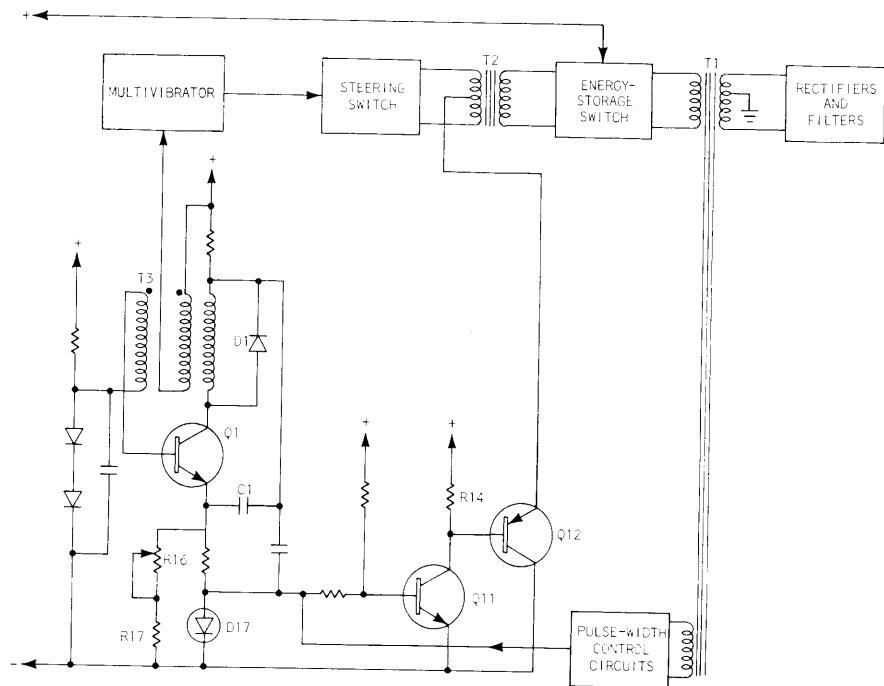
Fig. 4-15. Blocking oscillator and pulse
amplifier.
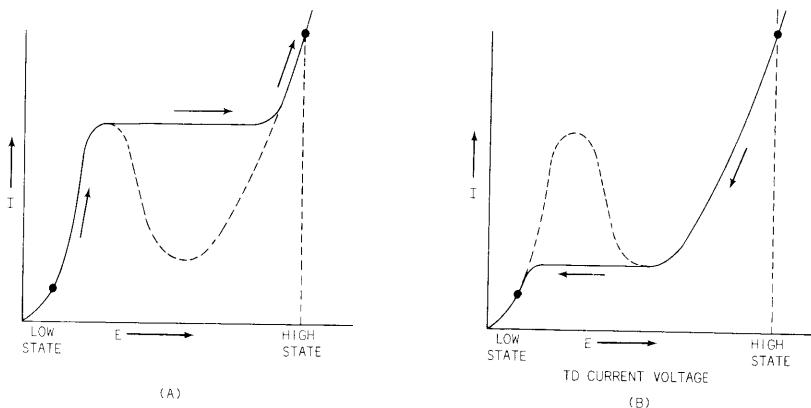
Fig. 4-16. Tunnel-diode characteristic.
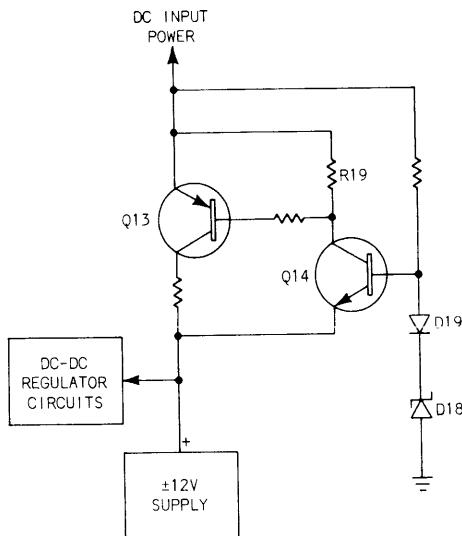
Fig. 4-17. Start circuit.

start
circuit

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

Briefly summarized, the DC-DC regulator operates as
follows: DC power at some level between 11.5 and
35 volts is applied to the energy-storage switch.
This switch is activated by pulses from the steering
switch. The energy level in the primary winding is
sampled by the pulse-width-control circuits which
determine the duration of the steering-switch pulses.
Since the duration of the steering-switch pulses
determine the level of stored energy attained by
each pulse of the energy-storage switch, a feedback
loop is established which maintains a constant energy
transfer to the secondary windings. Since the
instrument for which the regulator is designed
presents a constant load to the secondary circuits
(rectifiers), the voltage across the load remains
constant. Because power is consumed only when the
energy-storage switch is closed, a very high
efficiency is attained by this type of regulator.
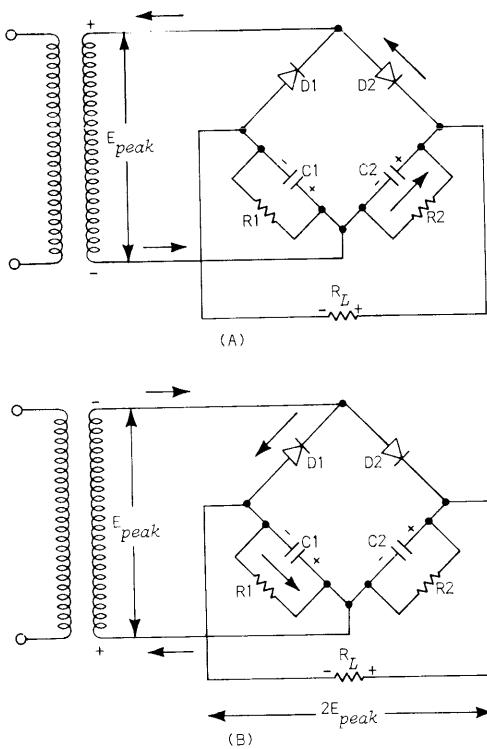
Fig. 5-1. Full-wave voltage doubler.

5

HIGH–VOLTAGE
POWER SUPPLIES

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

voltage
multiplier

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

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

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

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

cascade
multiplier

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

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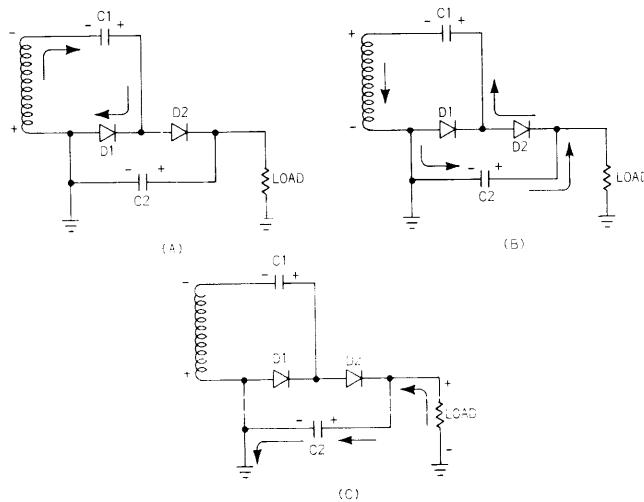
Fig. 5-2. Cascade doubler.

The voltage on C2 is twice that across the transformer secondary winding and is applied to the load. Fig. 5-2C shows the direction of discharge.

During each succeeding AC cycle, C2 receives a new charge in the manner outlined above, replenishing the charge lost in the form of load current.

cascade
tripler

The same principles govern the action of the *cascade tripler*, shown in Fig. 5-3. When the bottom of the secondary winding becomes positive, capacitor C1 charges through V1 to E_{peak} on the first alternation. On the second alternation, when the voltage on the bottom of the secondary winding becomes negative, C1 and the secondary winding series-aid to charge capacitor C2 to twice the secondary peak voltage. On the third alternation, when the bottom of the secondary winding again becomes positive, the voltage on C2 and the secondary series-aid; but due to the opposing charge on C1, *capacitor* C3 charges to only twice the secondary peak voltage, making the total charge across C1 and C3 three-times E_{peak}. This combined charge then discharges through the load. Succeeding alternations keep all capacitors charged in the manner described above, replenishing the charge lost in the form of load current.

The process described above can be extended to cascade quadruplers, pentuplers, sextuplers, etc. In vacuum-tube supplies, sufficiently high voltages for CRT operation can usually be attained by triplers or quadruplers. In semiconductor construction, several additional stages may be required to avoid high peak inverse voltages on the diodes. However, this requirement is not considered a disadvantage, due to the small size and relatively low cost of such diodes.
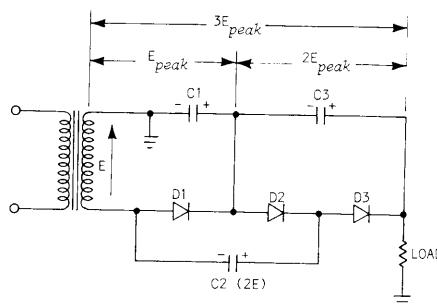

Fig. 5-3. Cascade tripler.
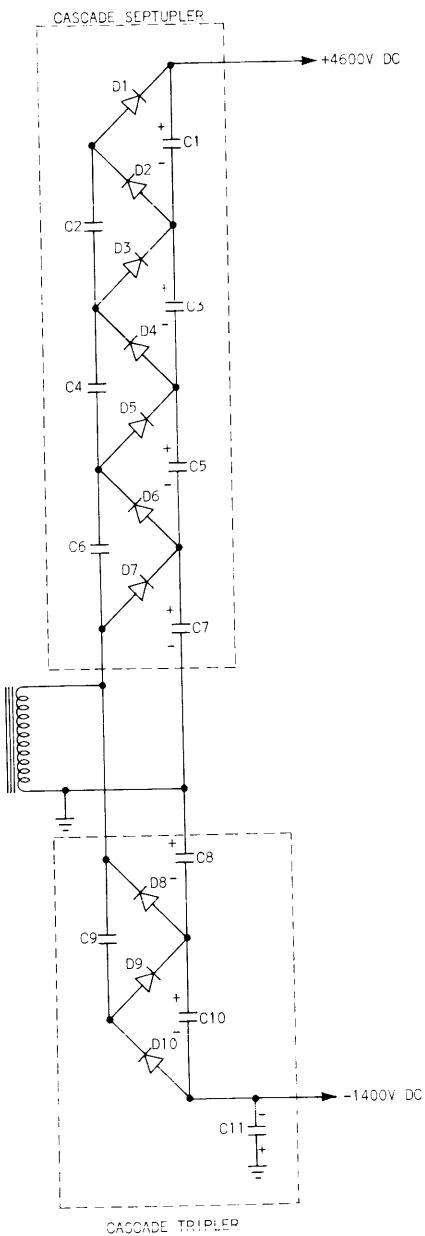
Fig. 5-4. High-voltage power supply.

high-
voltage
supply

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

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

voltage
regulation

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

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

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

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

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

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

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

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

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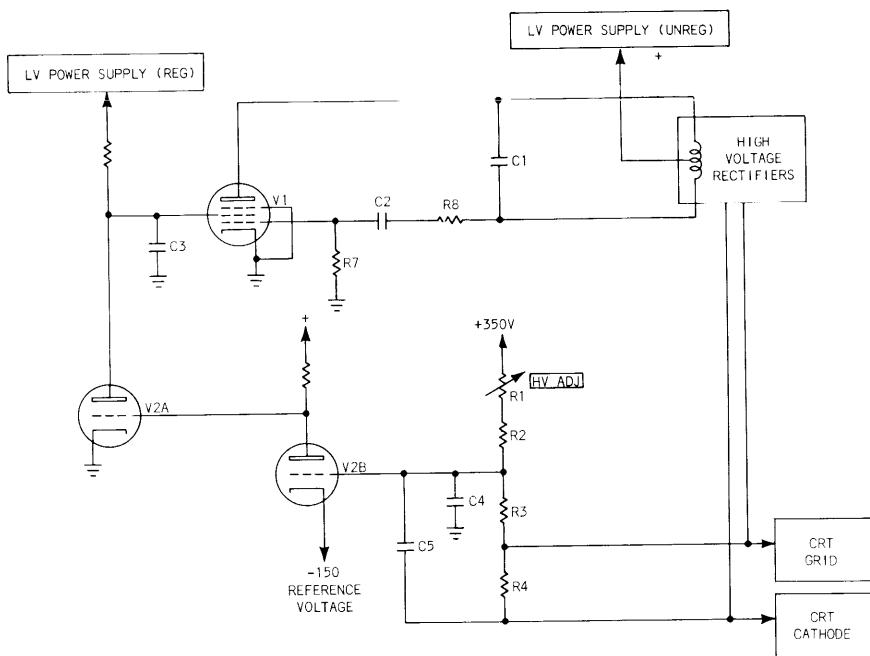
Fig. 5-5 High-voltage oscillator circuit.

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

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

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

high-
voltage
oscillator

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

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

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

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

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

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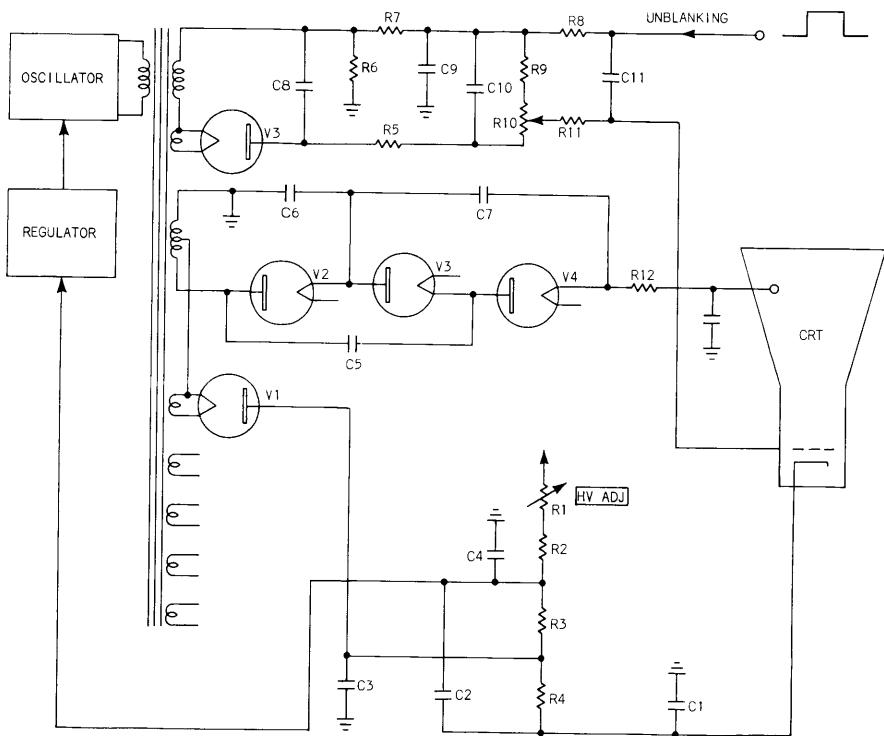
Fig. 5-6. High-voltage rectifiers.

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

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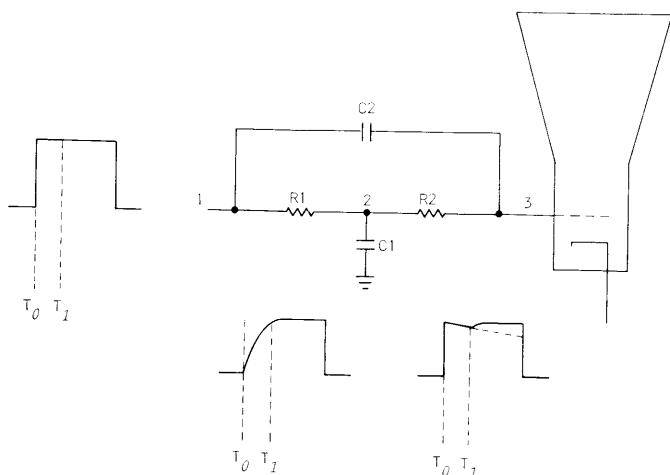
Fig. 5-7. CRT-grid-supply unblanking-pulse
correction circuit.

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

               In the actual circuit (Fig. 5-6), C11 acts as the
               coupling capacitor while resistors R7, R8, and R11
               provide the necessary isolation between transformer
               capacitance, CRT grid, and the unblanking-pulse input.
               C9 and R6 provide RF filtering for the HV-oscillator
               frequency, preventing the RF from being coupled into
               the horizontal sweep generator through the unblanking
               circuit. C8, C10, and R5 constitute an RC filter for
               the rectifier output. R2, R3, R4, and HV-adjust
               potentiometer R1 form a voltage divider across the
               cathode-supply rectifier output. The adjustment of
   R10 determines beam intensity for the CRT.
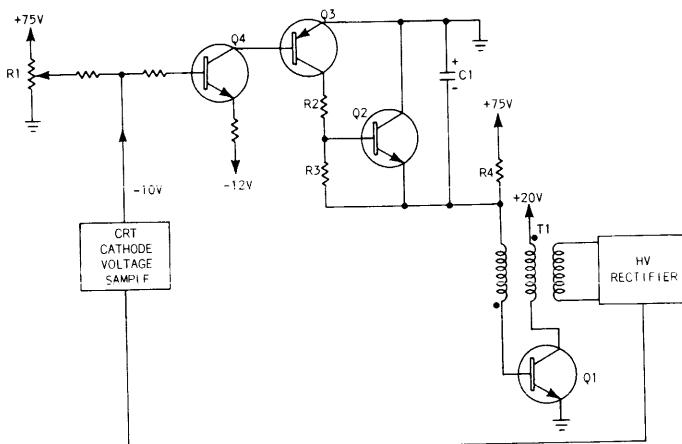
Fig. 5-8. Transistorized HV oscillator.

transistor-
ized
high-voltage
supplies

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

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

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

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

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

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

battery-
powered
high-voltage

Battery-powered instruments present special problems in their high-voltage power supplies. These instruments develop their low-voltage potentials in some types of oscillator-driven power supply (see DC-DC regulator). If their high-voltage potentials were developed in conventional fashion, an additional oscillator would be required. It is clear that a considerable savings in space and power consumption can be effected by using the output of the first oscillator to drive the high-voltage as well as the low-voltage transformer. However, this arrangement presents a problem in regulation. If a sample of the cathode voltage were used to develop a feedback error signal for the oscillator, all the low-voltage supplies would be affected by changes in cathode current. This is obviously an impossible arrangement.
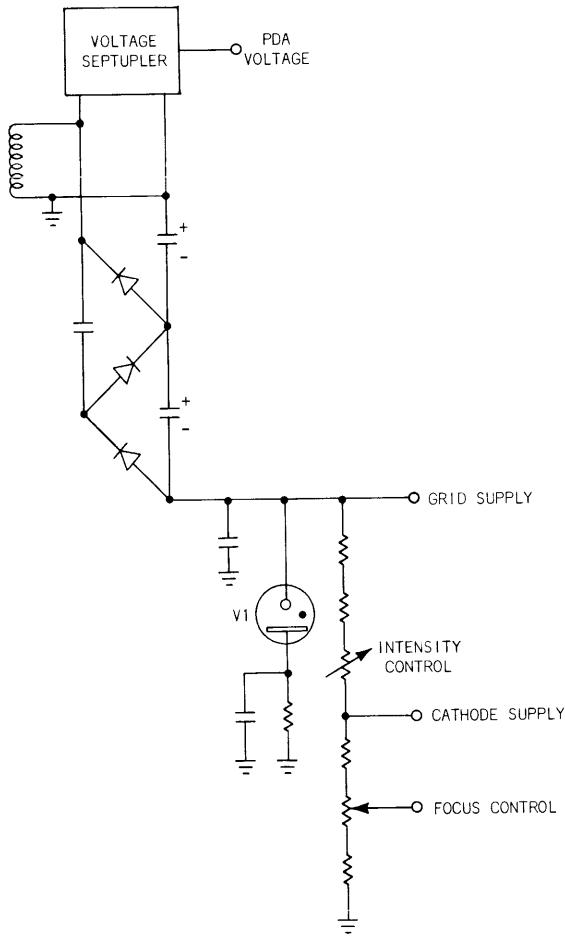
Fig. 5-9. HV supply for deflection-blanked CRT.

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

shock
regulator

corona
stabilizer

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

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

⑥

POWER SUPPLY HARDWARE

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

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

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

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

where N = number of turns in windings
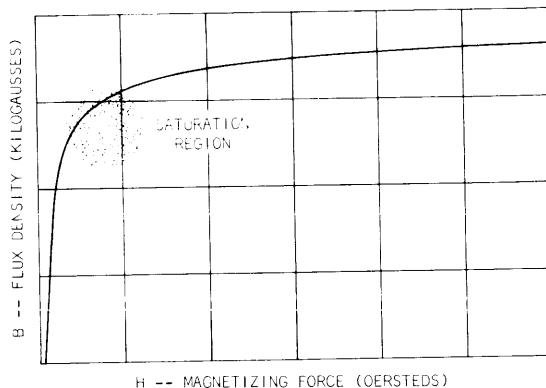
Fig. 6-1. Typical magnetization curve.

flux
density

magnetic
field
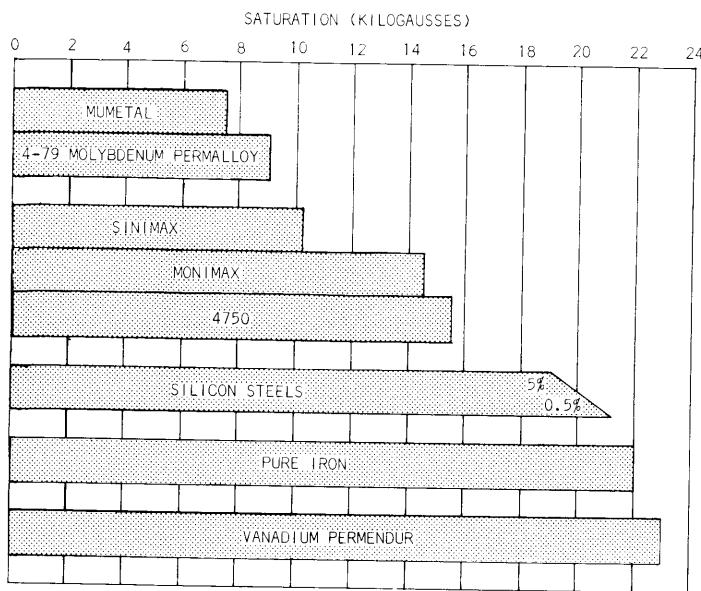
permeability

saturation

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

maximum
permeability

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

SATURATION (KILOGAUSSES)

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

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


THE RELATIONSHIP BETWEEN
B AND H IS NOT LINEAR;
THEIR RATIO B/H (PERMEA-
BILITY) ALSO VARIES.

MAXIMUM
PERMEABILITY
REGION

B -- FLUX DENSITY (KILOGAUSSES)
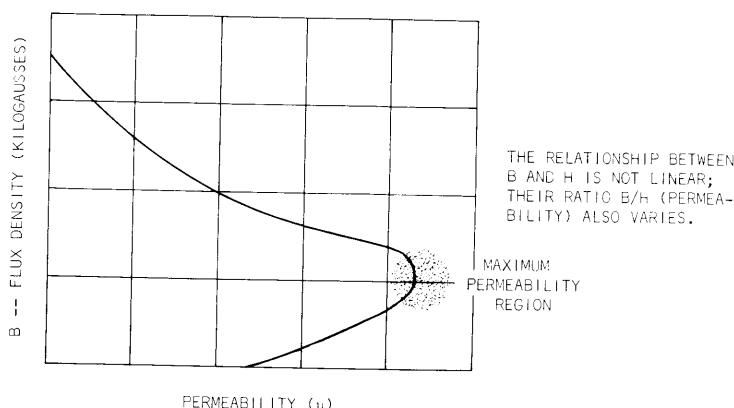
PERMEABILITY (μ)

Fig. 6-3.   Typical permeability curve.

The power losses which occur in a transformer are due
to a number of unavoidable, but reducible, effects.
The first of these arises from the fact that part of
the magnetic flux generated in the primary coil is
lost in the coupling process. Although this type of
loss can be reduced to a minimum by the proper choice
of core material, transformer geometry, etc., some
of the primary flux inevitably remains uncoupled and
therefore constitutes a small power loss.

coupling
losses

More important are the losses in the form of heat
generated in the transformer. The resistance of the
windings gives rise to what are known as "copper
losses." These losses increase with the frequency
of the applied voltage, due to "skin effect." This
is the tendency of high-frequency currents to travel
at or near the surface of a conductor, thereby
reducing the effective cross-sectional area of the
conductor. A similar phenomemon, known as "proximity
effect," is caused by distortion of the current-
carrying area in one winding by the magnetic field
of the windings closely adjacent. This effect also
increases the resistance of the windings to a small
degree.

copper
losses

skin
effect

proximity
effect

A third type of heat loss is caused by small currents
called "eddy currents" which are induced in the core
material itself by the alternating magnetic field
(Fig. 6-4). The amplitude of these currents depends
on the frequency and flux density of the field as
well as upon the specific resistance and thickness
of the core material. The resistance offered to
eddy currents is highly increased by constructing the
core from thin insulated laminations. This technique
has no adverse effect on the core's ability to
concentrate the flux field. The specific resistance
of the core material is controlled to a certain
extent by various metallurgical processes.

eddy
currents

Finally, there is the loss due to a phenomenon
called "hysteresis." As the magnetic field shifts
back and forth in polarity, the molecules of core
material align themselves first in one direction
and then in the opposite. The resistance of the
individual molecules to these constant changes in
alignment varies with the type of core material and
determines the degree of hysteresis loss.

hysteresis

EDDY CURRENTS

MAGNETIZING FORCE H

ENERGY LOSS IN THE
FORM OF HEAT

MAGNETIC FIELD B
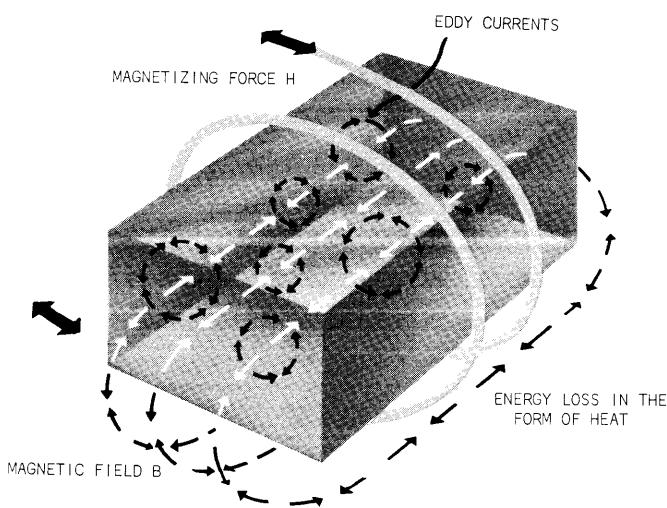
Fig. 6-4. Eddy currents.

132
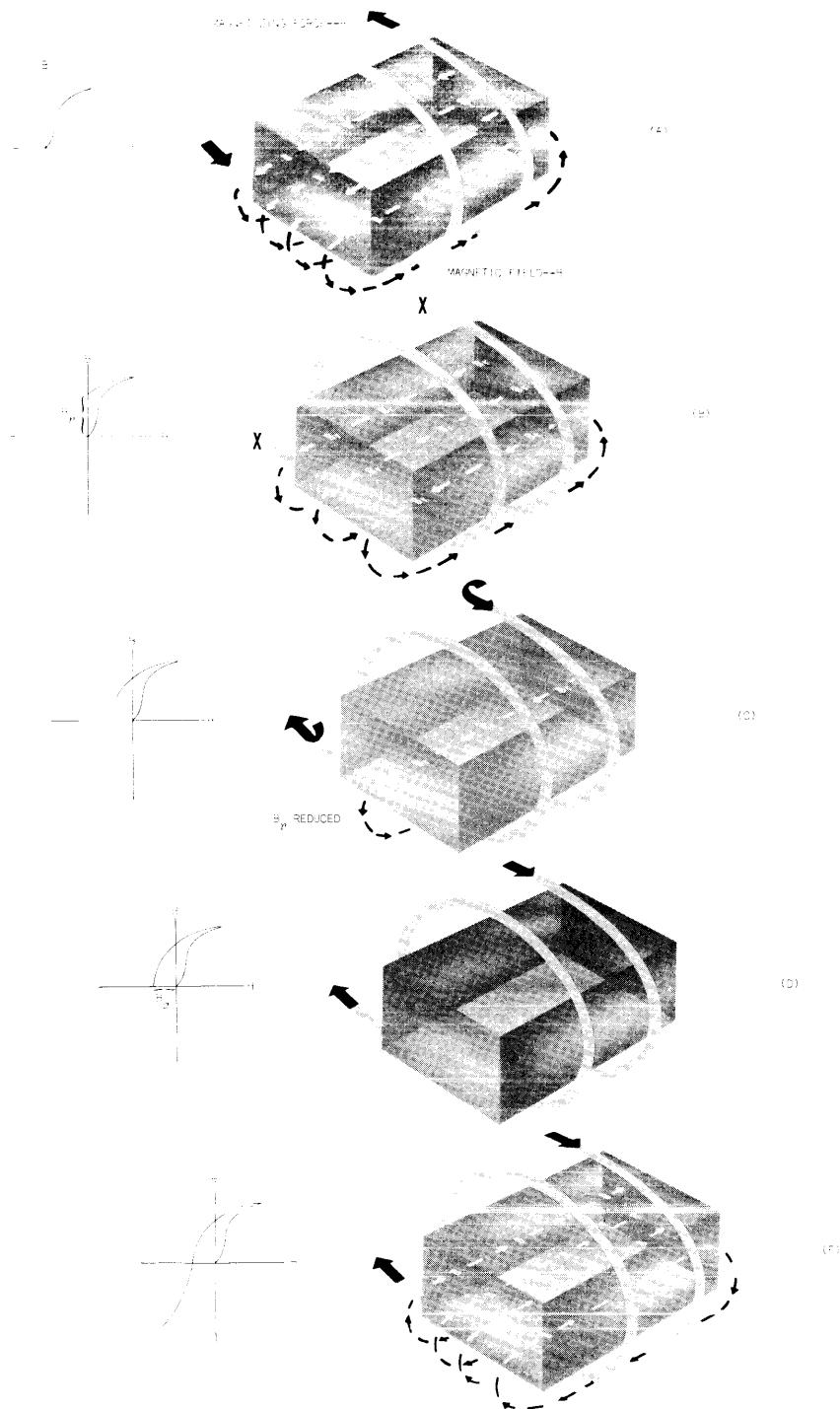
Fig. 6-5. The hysteresis cycle.

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

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

hysteresis loop

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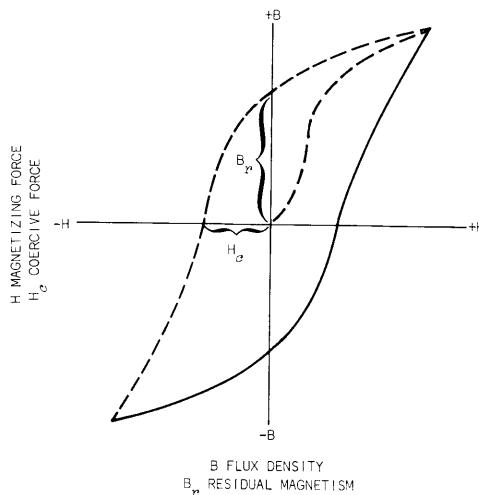

Fig. 6-6. Typical hysteresis loop.

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

core
materials

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

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

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

Fig. 6-8 shows the wide differences in permeability
of a single crystal of iron when measured along
different axes of the crystal.
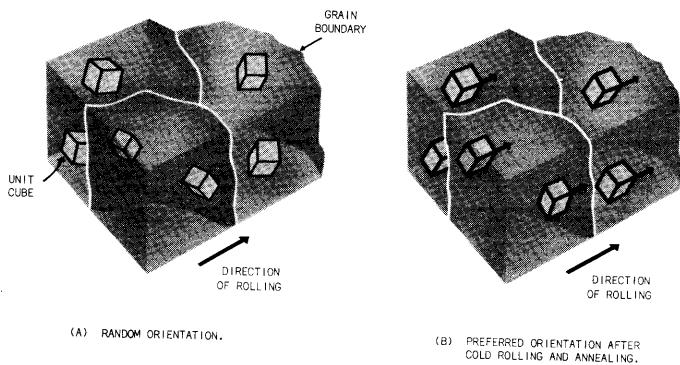
Fig. 6-7. Orientation in grains of iron.
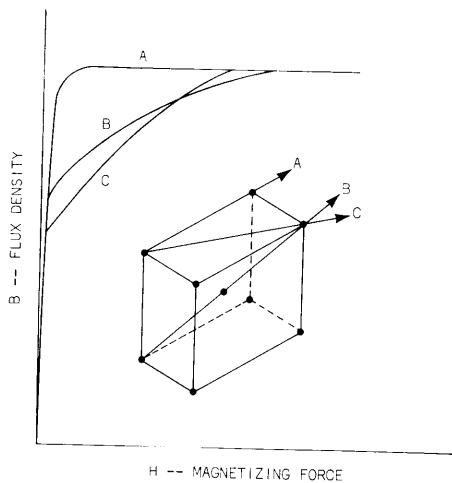
Fig. 6-8. Permeability along different directions of a single crystal of iron.

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

transformer
geometry
consider-
ations

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

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

Type of Load Resistance: Inductive loads versus capacitive loads.

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

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

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

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

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

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

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

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

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

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

transformer-design character-istics

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

1. *Low External Field:* The alternating magnetic field surrounding a power-supply transformer must be held to a minimum to prevent induced AC modulation in adjacent circuits and consequent "jitter" in the CRT display.

2. *Low Temperature Rise:* Power-supply transformers must not generate excessive heat in the instrument. High temperatures shorten the life of electronic components and change their electrical characteristics. Semiconductors, in particular, require a constant-temperature operating environment for stable performance.

3. *Low Regulation:* The term "regulation," when applied to transformer characteristics, indicates the difference in voltage on the

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

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

line-
voltage
selection

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

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

Unless otherwise specified, instruments are generally
wired for 115 V, and adaptive changes must be made
in the field. Several methods are provided for this
contingency. In the older instruments the necessary
changes are made by unsoldering the factory connections
and making the required connections according to
instructions in the instrument manual. Care must be
exercised when making these changes to assure the
proper working voltage (115 V) for the fan motor.
In other instruments the necessary changes can be
made by shifting square-pin connectors mounted on
a power board. Sometimes the nominal voltage
selection can be made by a simple snap-switch, and
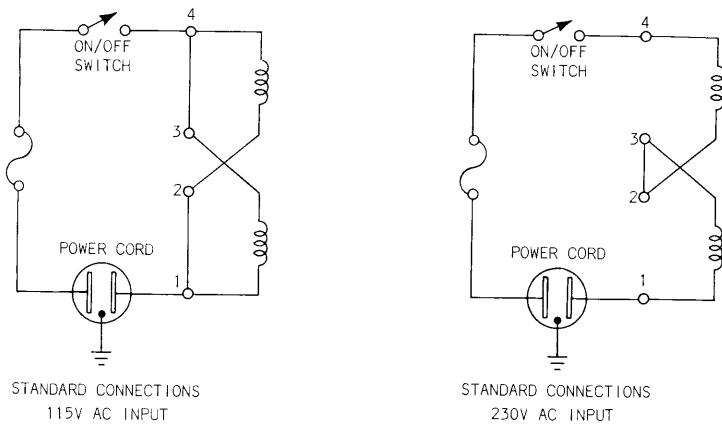
Fig. 6-9. Transformer primary connections.
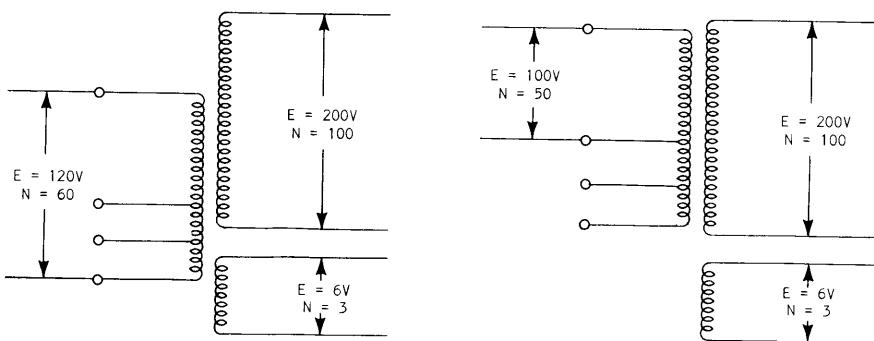
Fig. 6-10. Line-voltage compensation.
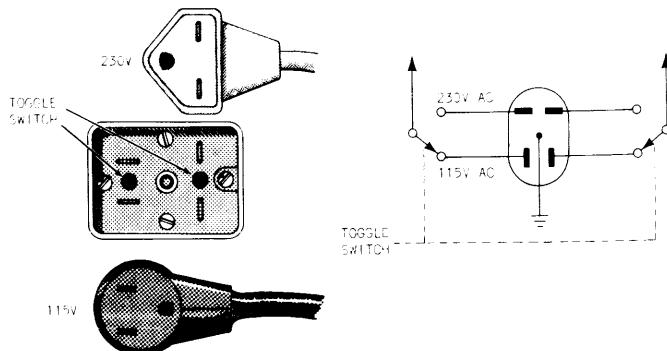
Fig. 6-11.   115 V - 230 V selector.

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

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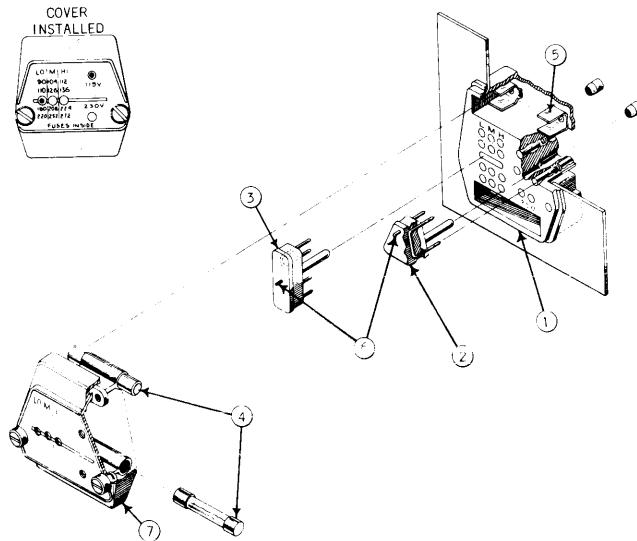
Fig. 6-12.   Voltage-selection device.

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

voltage-
regulating
transformers

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

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


$E_{peak} = 1.414$RMS $E_{peak} < 1.414$RMS $E_{peak} > 1.414$RMS

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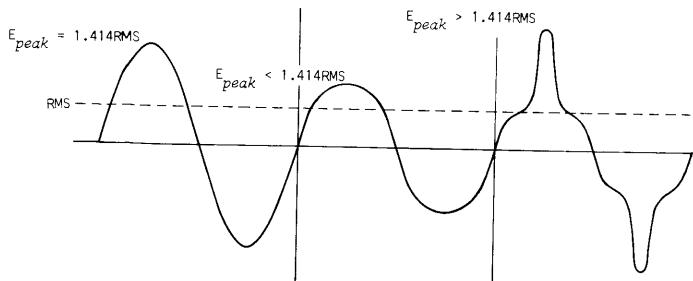
Fig. 6-13. Effects of distortion on E_{peak}.

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

filament
windings

DC
filaments

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

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

line
filters

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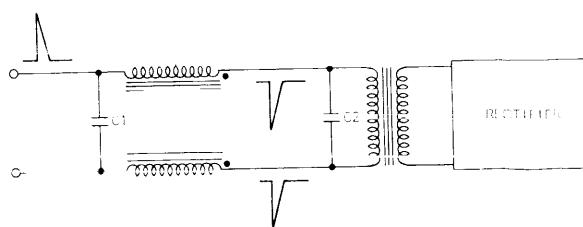
Fig. 6-14. Line (RFI) filter.

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

thermal
relays

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

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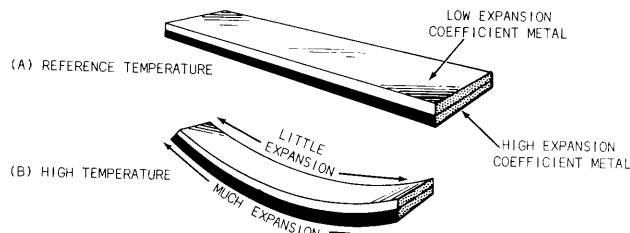

Fig. 6-15. Bimetallic strip.
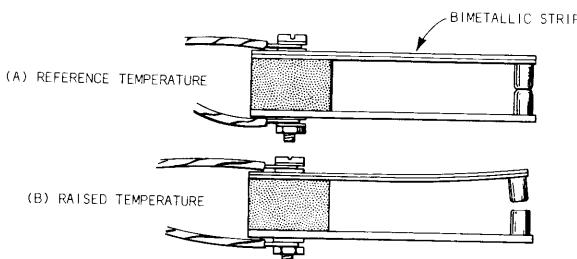
Fig. 6-16. Simple thermal relay.
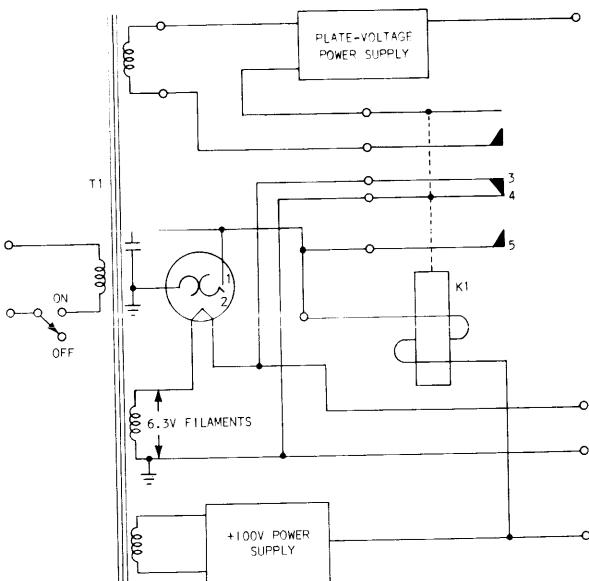
Fig. 6-17. Time-delay relay circuit.

time-delay
relay

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

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

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

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

INDEX

CIRCUIT CONCEPTS
BY INSTRUMENT

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

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

504	507	515A	516	519	520	524AD	526	528	529	531A	533A	535A	536	543B	544	545B	546	547	549	551	555	556
		●	●						●	●	●	●	●	●	●	●	●	●		●	●	●
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CIRCUIT CONCEPTS:	561A	564	565	567	568	575	581A	585A	647A	661
CRT Grid-unblanking Correction (P120-121)							•	•		
Line-voltage Selection, Interlocked (p140-141)										
Line-voltage Selection, Plug-in (p139)										
Line-voltage Selection, Switchable, (p138)										
Multiplication, Cascade Doubler (p110)										
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Oscillator-driven Power Supplies, AC-DC (p85-108)										
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Rectification, Full-wave Center-tapped (p15-16)	•	•	•	•	•	•			•	
Rectification, Half-wave (p3-8, 83, 115, 118)	•	•	•	•	•	•	•	•	•	•
RFI Filtering (p142-143)					•				•	•
Thermal Protection (p87, 143)	•	•	•	•	•	•	•	•	•	•
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Voltage Regulation, Vacuum Tube (p3, 17-69)	•	•	•	•		•	•	•		•

PL UG-INS POWER SUPPLIES DIGITAL AND PROGRAMMING

1A4	1S1	1S2	3T2	3T4	3T77A	125	127	129	132	133	160A	285	230	240	262
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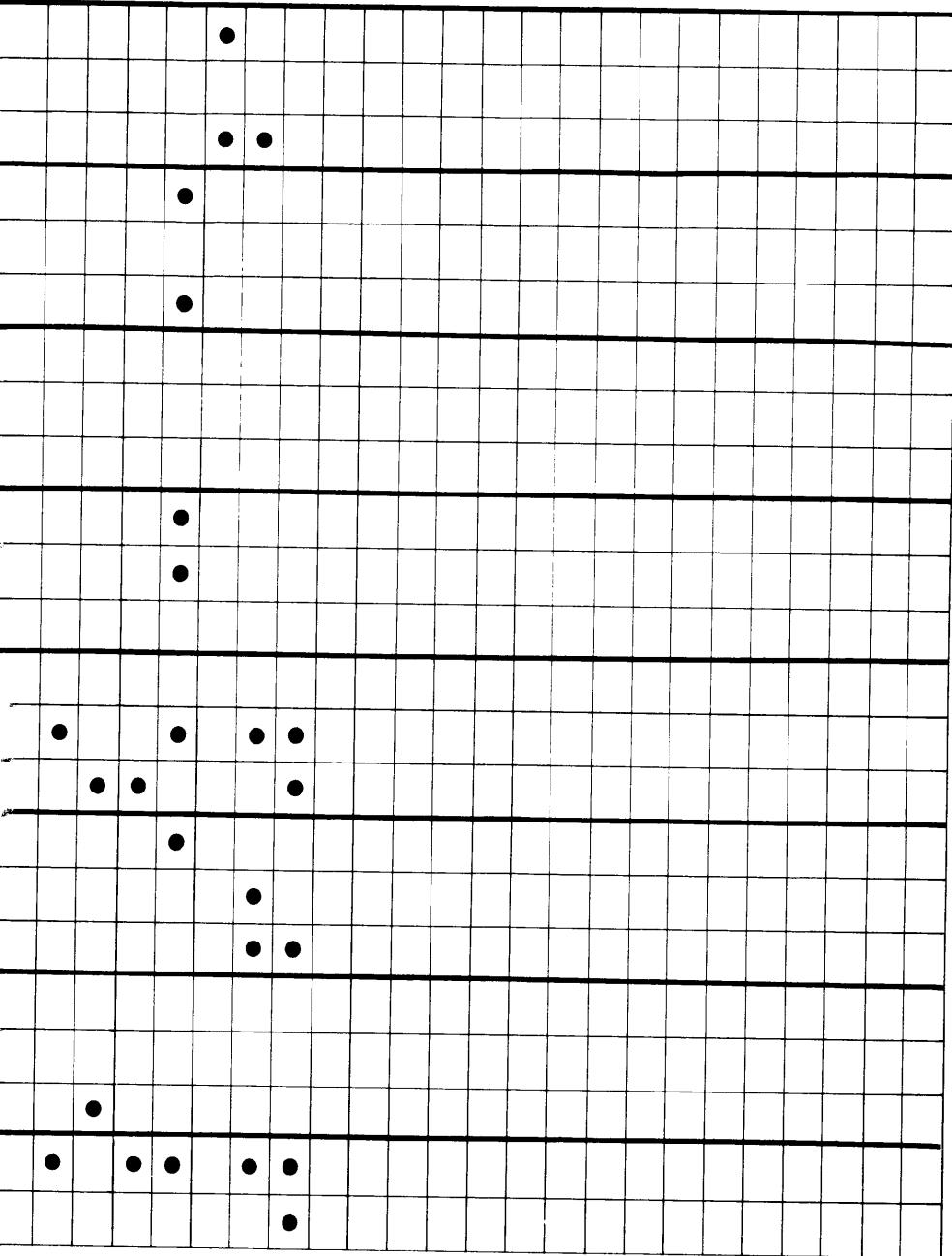
	81A	130	292	410	601	611	1121
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This book is but one of a series.
The series consists of two groups,
circuits and measurements. These
texts present a conceptual approach
to circuits and measurements which
apply to Tektronix products.
Several "concept books" are in
preparation; those now available
are:

Power Supply Circuits
062-0888-01

Oscilloscope Cathode-Ray Tubes
062-0852-01

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

Television Waveform Processing Circuits
062-0955-00

Information Display Concepts
062-1005-00

Semiconductor Devices
062-1009-00