Inductance

It is possible to show that the flow of current through a conductor is accompanied by magnetic effects; a compass needle brought near the conductor, for example, will be deflected from its normal north-south position. The current, in other words, sets up a magnetic field.

The transfer of energy to the magnetic field represents work done by the source of emf. Power is required for doing work, and since power is equal to current multiplied by voltage, there must be a voltage drop in the circuit during the time in which energy is being stored in the field. This voltage "drop" (which has nothing to do with the voltage drop in any resistance in the circuit) is the result of an opposing voltage "induced" in the circuit while the field is building up to its final value. When the field becomes constant the *induced emf* or *back emf* disappears, since no further energy is being stored.

Since the induced emf opposes the emf of the source, it tends to prevent the current from rising rapidly when the circuit is closed. The amplitude of the induced emf is proportional to the rate at which the current is changing and to a constant associated with the circuit itself, called the *inductance* of the circuit.

Inductance depends on the physical characteristics of the conductor. If the conductor is formed into a coil, for example, its inductance is increased. A coil of many turns will have more inductance than one of few turns, if both coils are otherwise physically similar. Also, if a coil is placed around an iron core its inductance will be greater than it was without the magnetic core.

The polarity of an induced emf is always such as to oppose any change in

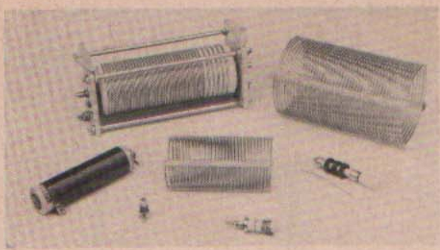


Fig. 18 — Assorted inductors. A rotary (continuously variable) coil is at the upper left. Slug-tuned inductors are visible in the lower foreground. An rf choke (three pi windings) is seen at the lower right.

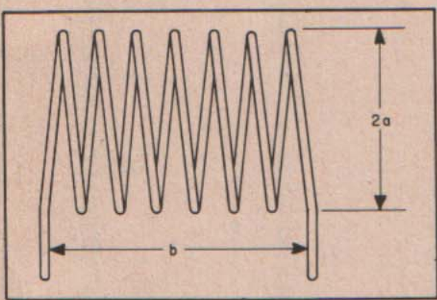


Fig. 19 — Coil dimensions used in the inductance formula. The wire diameter does not enter into the formula.

the current in the circuit. This means that when the current in the circuit is increasing, work is being done against the induced emf by storing energy in the magnetic field. If the current in the circuit tends to decrease, the stored energy of the field returns to the circuit, and thus adds to the energy being supplied by the source of emf. This tends to keep the current flowing even though the applied emf may be decreasing or be removed entirely.

The unit of inductance is the *henry*. Values of inductance used in radio equipment vary over a wide range. Inductance of several henrys is required in power-supply circuits (see chapter on power supplies), and to obtain such values of inductance it is necessary to use coils of many turns wound on iron cores. In radio-frequency circuits, the inductance values used will be measured in *millihenrys* (a mH, one one-thousandth of a henry) at low frequencies, and in microhenrys (μH, one one-millionth of a henry) at medium frequencies and higher. Although coils for radio frequencies may be wound on special iron cores (ordinary iron is not suitable), most rf coils made and used by amateurs are of the "air-core" type; that is, wound on an insulating support consisting of non-magnetic material (Fig. 18).

Every conductor has inductance, even though the conductor is not formed into a coil. The inductance of a short length of straight wire is small, but it may not be negligible because if the current through it changes its intensity rapidly enough the induced voltage may be appreciable. This will be the case in even a few inches of wire when an alternating current having a frequency of the order of 100 MHz, or higher is flowing. However, at much lower frequencies the inductance of the same wire could be ignored because the induced voltage would be negligibly small.

Calculating Inductance

The approximate inductance of single-layer air-core coils may be calculated from the simplified formula

$$L\ (\mu H) = \frac{a^2n^2}{9a + 10b}$$

where L = inductance in microhenrys
a = coil radius in inches
b = coil length in inches
n = number of turns

The notation is explained in Fig. 19. This formula is a close approximation for coils having a length equal to or greater than 0.8a.

Example: Assume a coil having 48 turns wound 32 turns per inch and a diameter of 3/4 inch. This a = 0.75/2 = 0.375, b = 48/32 = 1.5, and n = 48. Substituting,

$$L = \frac{.375 \times .375 \times 48 \times 48}{(9 \times .375) + (10 \times 1.5)} = 17.6\ \mu H$$

To calculate the number of turns of a single-layer coil for a required value of inductance,

$$n = \sqrt{\frac{L\,(9a + 10b)}{a^2}}$$

Example: Suppose an inductance of 10 μH is required. The form on which the coil is to be wound has a diameter of one inch and is long enough to accommodate a coil of 1-1/4 inches. Then a = 0.5, b = 1.25, and L = 10. Substituting,

$$n = \sqrt{\frac{10\,(4.5 + 12.5)}{0.5 \times 0.5}} = \sqrt{680}$$

$$= 26.1 \text{ turns}$$

A 26-turn coil would be close enough in practical work. Since the coil will be 1.25 inches long, the number of turns per inch will be 26.1/1.25 = 20.8. Consulting the wire table, we find that no. 17 enameled wire (or anything smaller) can be used. The proper inductance is obtained by winding the required number of turns on the form and then adjusting the spacing between the turns to make a uniformly spaced coil 1.25 inches long.

Inductance Charts

Most inductance formulas lose accuracy when applied to small coils (such as are

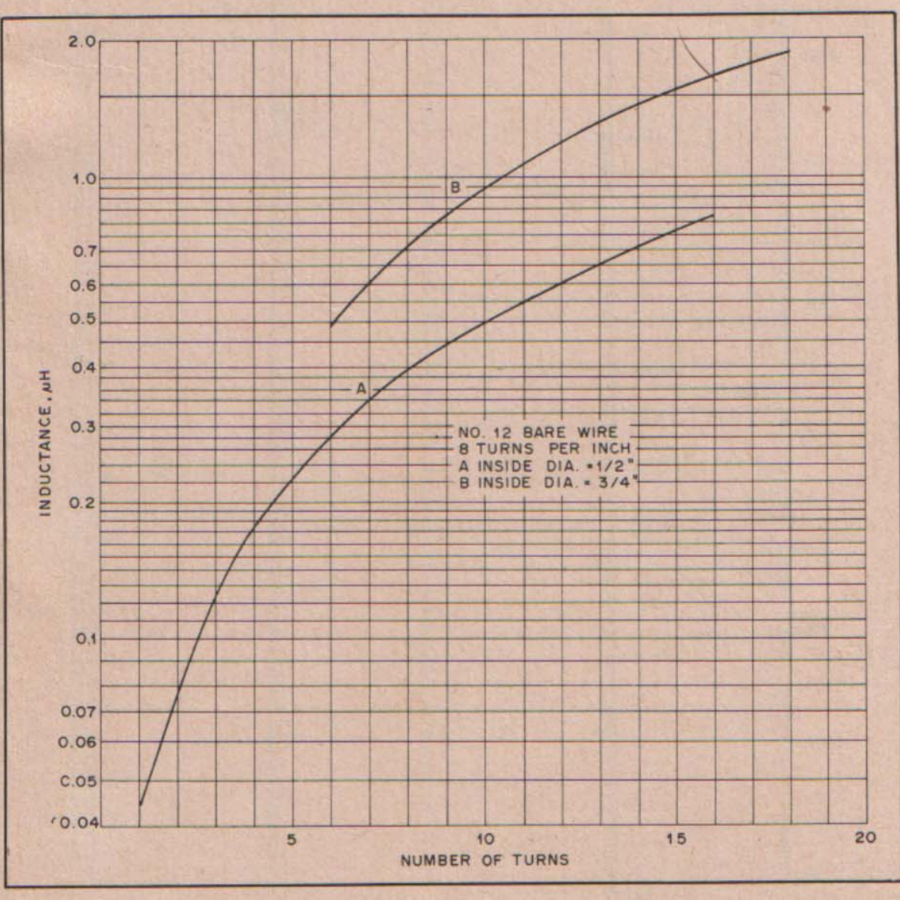
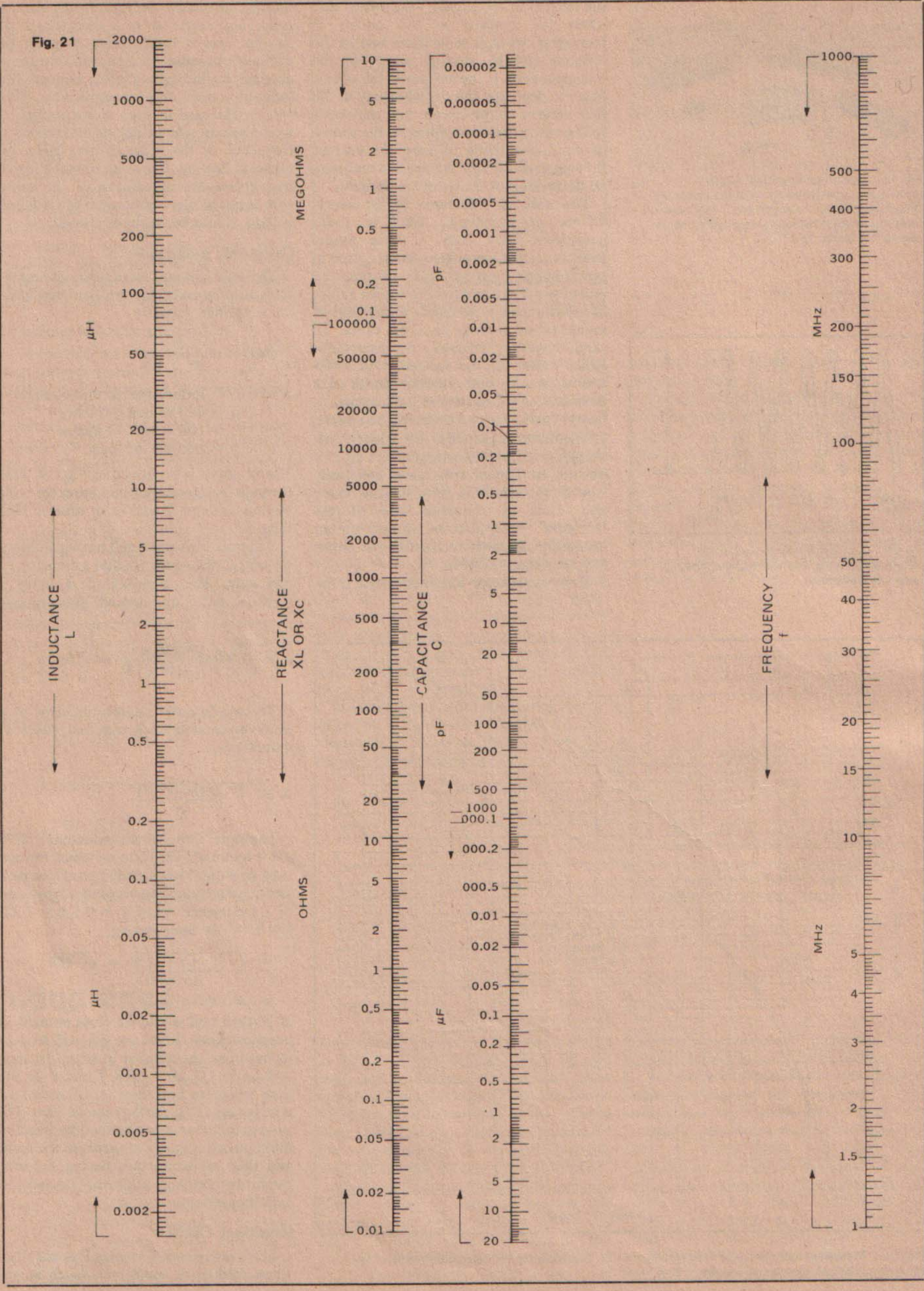


Fig. 20 — Measured inductance of coils wound with no. 12 bare wire, eight turns to the inch. The values include half-inch leads. Inches $\times$ 25.4 = mm.

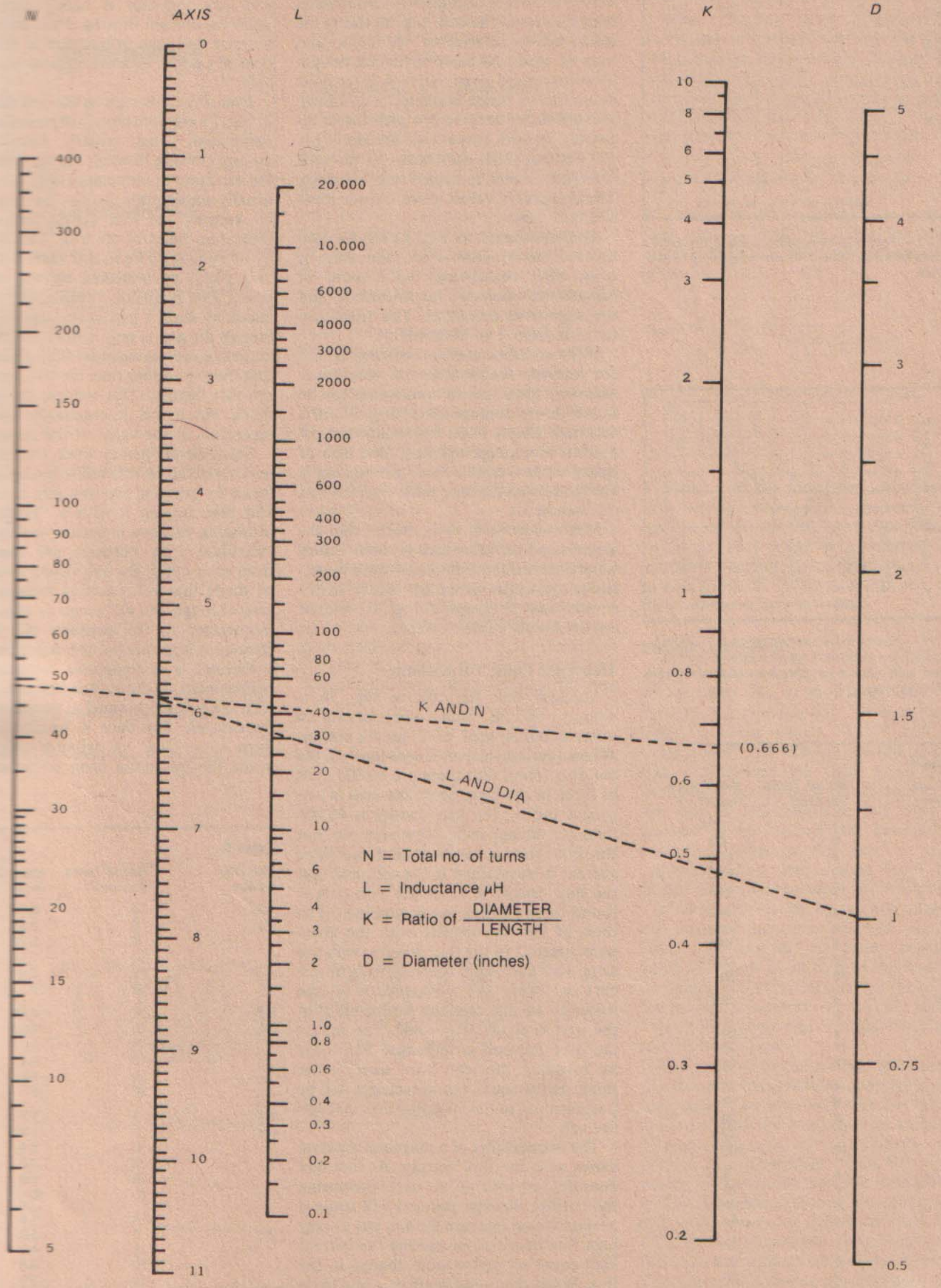
Fig. 21



INDUCTANCE L
µH
2000 1000 500 200 100 50 20 10 5 2 1 0.5 0.2 0.1 0.05 0.02 0.01 0.005 0.002
REACTANCE XL OR XC
MEGOHMS
10 5 2 1 0.5 0.2 0.1
OHMS
100000 50000 20000 10000 5000 2000 1000 500 200 100 50 20 10 5 2 1 0.5 0.2 0.1 0.05 0.02 0.01
CAPACITANCE C
pF
0.00002 0.00005 0.0001 0.0002 0.0005 0.001 0.002 0.005 0.01 0.02 0.05 0.1 0.2 0.5 1 2 5 10 20 50 100 200 500 1000
µF
000.1 000.2 000.5 0.01 0.02 0.05 0.1 0.2 0.5 1 2 5 10 20
FREQUENCY f
MHz
1000 500 400 300 200 150 100 50 40 30 20 15 10 5 4 3 2 1.5 1

Fig. 22

SINGLE-LAYER WOUND COIL CHART



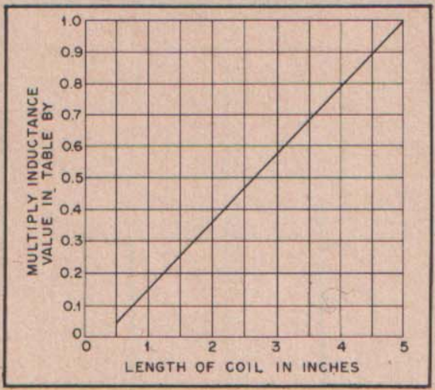


Fig. 23 — Factor to be applied to the inductance of coils listed in Table 4 for coil lengths up to five inches.

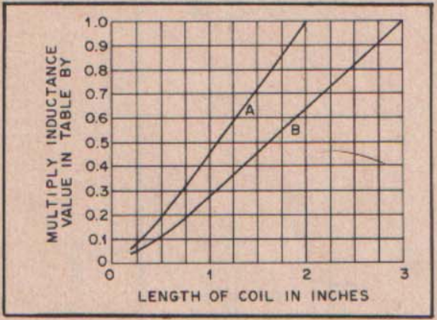


Fig. 24 — Factor to be applied to the inductance of coils listed in Table 5, as a function of coil length. Use curve A for coils marked A, and curve B for coils marked B.

Table 4

Coil Dia, Inches	*No. of Turns Per Inch*	*Inductance in μH*
1-1/4	4	2.75
	6	6.3
	8	11.2
	10	17.5
	16	42.5
1-1/2	4	3.9
	6	8.8
	8	15.6
	10	24.5
	16	63
1-3/4	4	5.2
	6	11.8
	8	21
	10	33
	16	85
2	4	6.6
	6	15
	8	26.5
	10	42
	16	108
2-1/2	4	6.6
	6	23
	8	41
	10	64
3	4	14
	6	31.5
	8	56
	10	89

Inches × 25.4 = mm.

used in vhf work and in low-pass filters built for reducing harmonic interference to television) because the conductor thickness is no longer negligible in comparison with the size of the coil. Fig. 20 shows the measured inductance of vhf coils, and may be used as a basis for circuit design. Two curves are given: curve A is for coils wound to an inside diameter of 1/2 inch; curve B is for coils of 3/4 inch inside diameter. In both curves the wire size is no. 12, winding pitch eight turns to the inch (1/8 inch center-to-center turn spacing). The inductance values given include leads 1/2 inch long.

The nomograph of Fig. 21 can be used for fast determination of the inductance of coils, their reactances, the amount of capacitance necessary for resonance, and the capacitive reactance. The frequency range is from 1 to 1000 MHz.

If the coil diameter and winding length are known, the number of wire turns necessary for a specific inductance can be found in the nomograph of Fig. 22. It is necessary also to know how many turns of a given wire gauge will fit in one inch of space, close wound. This information is available from the wire table elsewhere in this handbook.

Machine-wound coils with the diameters and turns per inch given in Tables 4 and 5 are available in many radio stores, under the trade names of "B&W Miniductor" and "Polycoils." Figs. 23 and 24 are used with Tables 4 and 5.

Iron-Core Coils: Permeability

Suppose that the coil in Fig. 25 is wound on an iron core having a cross-sectional area of 2 square inches. When a certain current is sent through the coil it is found that there are 80,000 lines of force in the core. Since the area is two square inches, the flux density is 40,000 lines per square inch. Now suppose that the iron core is removed and the same current is maintained in the coil, and that the flux density without the iron core is found to be 50 lines per square inch. The ratio of the flux density with the given core material to the flux density (with the same coil and same current) with an air core is called the *permeability* of the material. In this case the permeability of the iron is 40,000/50 = 800. The inductance of the coil is increased 800 times by inserting the iron core since, other things being equal, the inductance will be proportional to the magnetic flux through the coil.

The permeability of a magnetic material varies with the flux density. At low flux densities (or with an air core) increasing the current through the coil will cause a proportionate increase in flux, but at very high flux densities, increasing the current may cause no appreciable change in the flux. When this is so, the iron is said to be saturated. Saturation causes a rapid decrease in permeability, because it decreases the ratio of flux lines to those obtainable with the same current and an air core. Obviously, the inductance of an iron-core inductor is highly dependent upon the current flowing in the coil. In an air-core coil, the inductance is independent of current because air does not saturate.

Iron core coils such as the one sketched in Fig. 25 are used chiefly in power-supply equipment. They usually have direct current flowing through the winding, and the variation in inductance with current is usually undesirable. It may be overcome by keeping the flux density below the saturation point of the iron. This is done by opening the core so that there is a small "air gap," as indicated by the dashed lines. The magnetic "resistance" introduced by such a gap is so large — even though the gap is only a small fraction of an inch — compared with that of the iron that the gap, rather than the iron, controls the flux density. This reduces the inductance, but makes it practically constant regardless of the value of the current.

For radio-frequency work, the losses in iron cores can be reduced to a satisfactory figure by grinding the iron into a powder and then mixing it with a "binder" of insulating material in such a way that the individual iron particles are insulated from each other. By this means cores can be made that will function satisfactorily even through the vhf range — that is, at frequencies up to perhaps 100 MHz. Because a large part of the magnetic path is through a nonmagnetic material, the permeability of the iron is low compared with the values obtained at power-supply frequencies. The core is usually in the form of a "slug" or cylinder which fits inside the insulating form on which the

Table 5

Coil Dia, Inches	*No. of Turns Per Inch*	*Inductance in μH*
1/2 (A)	4	0.18
	6	0.40
	8	0.72
	10	1.12
	16	2.9
	32	12
5/8 (A)	4	0.28
	6	0.62
	8	1.1
	10	1.7
	16	4.4
	32	18
3/4 (B)	4	0.6
	6	1.35
	8	2.4
	10	3.8
	16	9.9
	32	40
1 (B)	4	1.0
	6	2.3
	8	4.2
	10	6.6
	16	16.9
	32	68

Inches × 25.4 = mm.

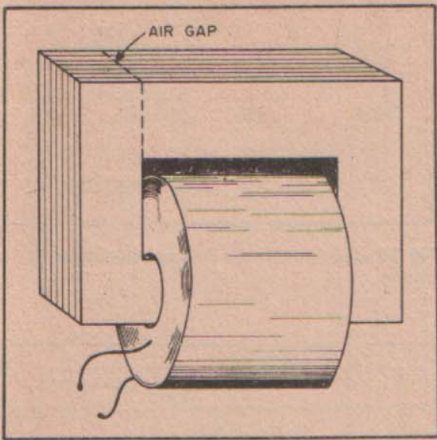


Fig. 25 — Typical construction of an iron-core inductor. The small air gap prevents magnetic saturation of the iron and thus maintains the inductance at high currents.

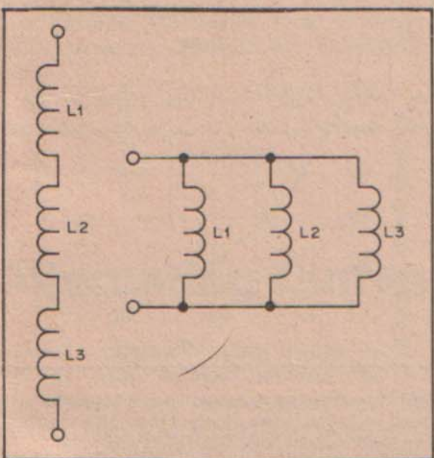


Fig. 26 — Inductances in series and parallel.

coil is wound. Despite the fact that with this construction the major portion of the magnetic path for the flux is in air, the slug is quite effective in increasing the coil inductance. By pushing the slug in and out of the coil, the inductance can be varied over a considerable range.

Eddy Currents and Hysteresis

When alternating current flows through a coil wound on an iron core an emf will be induced, as previously explained, and since iron is a conductor a current will flow in the core. Such currents (called *eddy currents*) represent a waste of power because they flow through the resistance of the iron and thus cause heating. Eddy-current losses can be reduced by *laminating* the core; that is, by cutting it into thin strips. These strips or *laminations* must be insulated from each other by painting them with some insulating material such as varnish or shellac.

There is also another type of energy loss: The iron tends to resist any change in its magnetic state, so a rapidly-changing current such as ac is forced continually to supply energy to the iron to overcome this "inertia." Losses of this sort are called *hysteresis* losses.

Eddy-current and hysteresis losses in iron increase rapidly as the frequency of the alternating current is increased. For this reason, ordinary iron cores can be used only at power and audio frequencies — up to, say, 15,000 hertz. Even so, a very good grade of iron or steel is necessary if the core is to perform well at the higher audio frequencies. Iron cores of this type are completely useless at radio frequencies.

Inductances in Series and Parallel

When two or more inductors are connected in series (Fig. 26) the total inductance is equal to the sum of the individual inductances, *provided the coils are sufficiently separated so that no coil is in the magnetic field of another.*
That is,

$$L_{total} = L1 + L2 + L3 + L4 + \ldots$$

If inductors are connected in parallel (Fig. 26) — and the coils are separated sufficiently, the total inductance is given by

$$L_{total} = \frac{1}{\frac{1}{L1} + \frac{1}{L2} + \frac{1}{L3} + \frac{1}{L4} + \cdots}$$

and for two inductances in parallel,

$$L = \frac{L1 \times L2}{L1 + L2}$$

Thus the rules for combining inductances in series and parallel are the same for resistances, if the coils are far enough apart so that each is unaffected by another's magnetic field. When this is not so the formulas given above cannot be used.

Mutual Inductance

If two coils are arranged with their axes on the same line, as shown in Fig. 27, a current sent through coil 1 will cause a magnetic field which "cuts" coil 2. Consequently, an emf will be induced in coil 2 whenever the field strength is changing. This induced emf is similar to the emf of self-induction, but since it appears in the *second* coil because of current flowing in the *first*, it is a "mutual" effect and results from the *mutual inductance* between the two coils.

If all the flux set up by one coil cuts all the turns of the other coil, the mutual inductance has its maximum possible value. If only a small part of the flux set up by one coil cuts, the turns of the other the mutual inductance is relatively small. Two coils having mutual inductance are said to be *coupled.*

The ratio of actual mutual inductance to the maximum possible value that could theoretically be obtained with two given coils is called the *coefficient of coupling* between the coils. It is frequently expressed as a percentage. Coils that have nearly the maximum possible (coefficient = 1 or 100 percent) mutual inductance are said to be *closely,* or *tightly,* coupled, but if the mutual inductance is relatively small the coils are said to be *loosely* coupled. The degree of coupling depends upon the physical spacing between the coils and how they are placed with respect to each

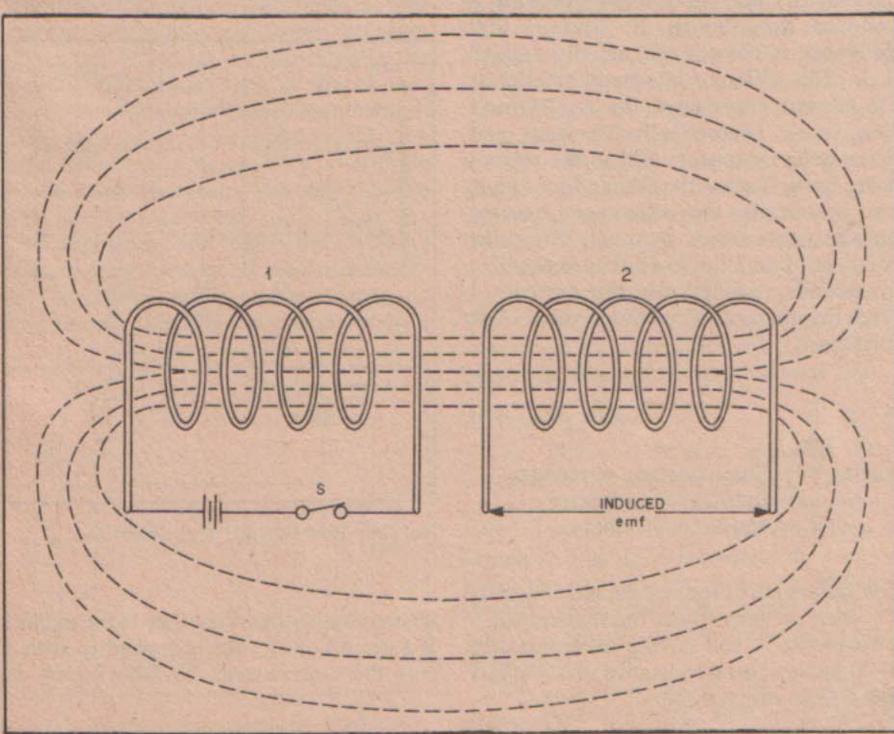


Fig. 27 — Mutual inductance. When the switch, S, is closed current flows through coil no. 1, setting up a magnetic field that induces an emf in the turns of coil no. 2.

other. Maximum coupling exists when they have a common axis and are as close together as possible (one wound over the other). The coupling is least when the coils are far apart or are placed so their axes are at right angles.

The maximum possible coefficient of coupling is closely approached only when the two coils are wound on a closed iron core. The coefficient with air-core coils may run as high as 0.6 or 0.7 if one coil is wound over the other, but will be much less if the two coils are separated.

Time Constant: Capacitance and Resistance

Connecting a source of emf to a capacitor causes the capacitor to become charged to the full emf practically instantaneously, if there is no resistance in the circuit. However, if the circuit contains resistance, as in Fig. 28A, the resistance limits the current flow and an appreciable length of time is required for the emf between the capacitor plates to build up to the same value as the emf of the source. During this "building-up" period, the current gradually decreases from its initial value, because the increasing emf stored on the capacitor offers increasing opposition to the steady emf of the source.

Theoretically, the charging process is never really finished, but eventually the charging current drops to a value that is smaller than anything that can be measured. The *time constant* of such a circuit is the length of time, in seconds, required for the voltage across the capacitor to reach 63 percent of the applied emf (this figure is chosen for mathematical reasons). The voltage across the capacitor rises with time as shown by Fig. 29.

The formula for time constant is

$$T = RC$$

where T = time constant in seconds
C = capacitance in farads
R = resistance in ohms

Example: The time constant of a 2-μF capacitor and a 250,000-ohm (0.25 M) resistor is

$$T = RC = 0.25 \times 2 = 0.5 \text{ second}$$

If the applied emf is 1000 volts, the voltage between the capacitor plates will be 630 volts at the end of 1/2 second.

If C is in microfarads and R in megohms, the time constant also is in seconds. These units usually are more convenient.

If a charged capacitor is *discharged* through a resistor, as indicated in Fig. 28B, the same time constant applies. If there were no resistance, the capacitor would discharge instantly when S was closed. However, since R limits the current flow the capacitor voltage cannot instantly go to zero, but it will decrease just as rapidly as the capacitor can rid itself of its charge through R. When the capacitor is discharging through a resistance, the time constant (calculated in the same way as above) is the time, in seconds, that it takes for the capacitor to lose 63 percent of its voltage; that is, for the voltage to drop to 37 percent of its initial value.

Example: If the capacitor of the example above is charged to 1000 volts, it will discharge to 370 volts in 1/2 second through the 250 Ω resistor.

Inductance and Resistance

A comparable situation exists when resistance and inductance are in series. In Fig. 30, first consider L to have no resistance and also assume that R is zero. Then closing S would tend to send a current through the circuit. However, the instantaneous transition from no current to a finite value, however small, represents a very rapid *change* in current, and a *back emf* is developed by the self-inductance of L that is practically equal and opposite to the applied emf. The result is that the initial current is very small.

The back emf depends upon the change in current and would cease to offer opposition if the current did not continue to increase. With no resistance in the circuit (which would lead to an infinitely large current, by Ohm's Law) the current would increase forever, always growing just fast enough to keep the emf of self-induction equal to the applied emf.

When resistance is in series, Ohm's Law sets a limit to the value that the current can reach. The back emf generated in L has only to equal the difference between E and the drop across R, because that difference is the voltage actually applied to L. This difference becomes smaller as the current approaches the final Ohm's Law value. Theoretically, the back emf never quite disappears and so the current never quite reaches the Ohm's Law value, but practically the differences become unmeasurable after a time. The time constant of an inductive circuit is the time in seconds required for the current to reach 63 percent of its final value. The formula is

$$T = \frac{L}{R}$$

where T = time constant in seconds
L = inductance in henrys
R = resistance in ohms.

The resistance of the wire in a coil acts as if it were in series with the inductance.

Example: A coil having an inductance of 20 henrys and a resistance of 100 ohms has a time constant of

$$T = \frac{L}{R} = \frac{20}{100} = 0.2 \text{ second}$$

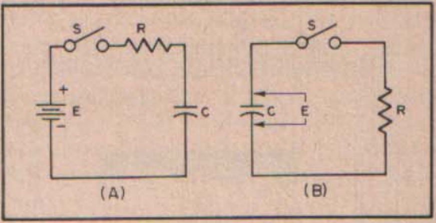


Fig. 28 — Illustrating the time constant of an RC circuit.

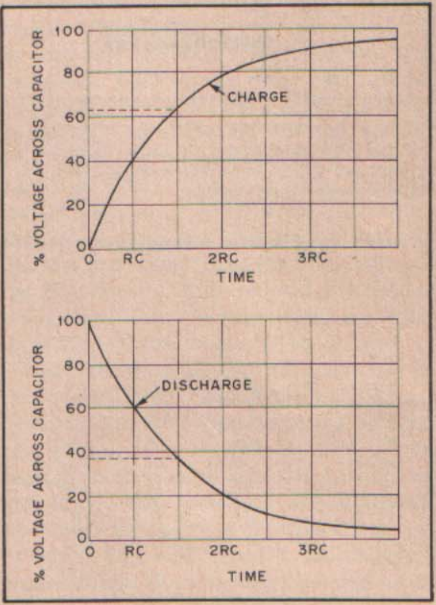


Fig. 29 — How the voltage across a capacitor rises, with time, when charged through a resistor. The lower curve shows the way in which the voltage decreases across the capacitor terminals on discharging through the same resistor.

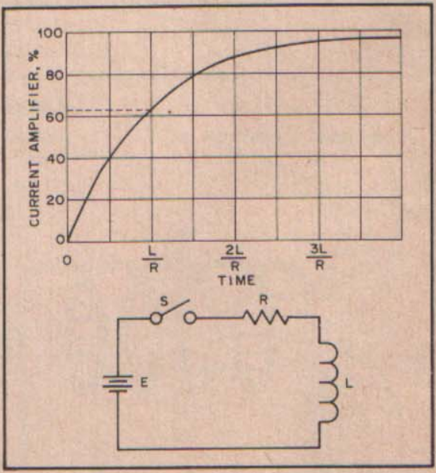


Fig. 30 — Time constant of an LR circuit.

if there is no other resistance in the circuit. If a dc emf of 10 volts is applied to such a coil, the final current, by Ohm's Law, is

$$I = \frac{E}{R} = \frac{10}{100} = 0.1 \text{ amp. or } 100 \text{ mA}$$

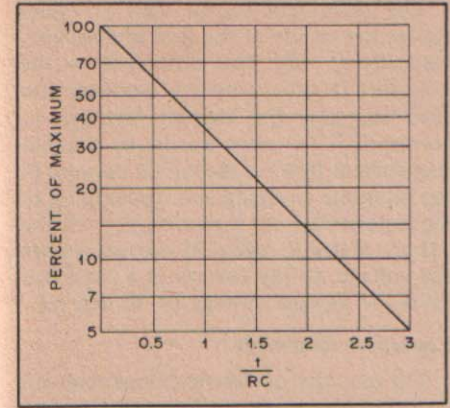


Fig. 31 — Voltage across capacitor terminals in a discharging RC circuit, in terms of the initial charged voltage. To obtain time in seconds, multiply the factor t/RC by the time constant of the circuit.

The current would rise from 0 to 63 millamperes in 0.2 second after closing the switch.

An inductor cannot be "discharged" in the same way as a capacitor, because the magnetic field disappears as soon as current flow ceases. Opening S does not leave the inductor "charged." The energy stored in the magnetic field instantly returns to the circuit when S is opened. The rapid disappearance of the field causes a very large voltage to be induced in the coil — ordinarily many times larger than the voltage applied, because the induced voltage is proportional to the *speed* with which the field changes. The common result of opening the switch in a circuit such as the one shown is that a spark or arc forms at the switch contacts at the instant of opening. If the inductance is large and the current in the circuit is high, a great deal of energy is released in a very short period of time. It is not at all unusual for the switch contacts to burn or melt under such circumstances. The spark or arc at the opened switch can be reduced or suppressed by connecting a suitable capacitor and resistor in series across the contacts.

Time constants play an important part in numerous devices, such as electronic keys, timing and control circuits, and shaping of keying characteristics by vacuum tubes. The time constants of circuits are also important in such applications as automatic gain control and noise limiters. In nearly all such applications a resistance-capacitance (RC) time constant is involved, and it is usually necessary to know the voltage across the capacitor at some time interval larger or smaller than the actual time constant of the circuit as given by the formula above. Fig. 31 can be used for the solution of such problems, since the curve gives the voltage across the capacitor, in terms of percentage of the initial charge, for percentages between 5 and 100, at any time after discharge begins.

Example: A 0.01-μF capacitor is charged to 150 volts and then allowed to discharge through a 0.1-megohm resistor. How long will it take the voltage to fall to 10 volts? In percentage, $10/150 = 6.7$ percent. From the chart, the factor corresponding to 6.7 percent is 2.7. The time constant of the circuit is equal to $RC = 0.1 \times 0.01 = 0.001$. The time is therefore $2.7 \times 0.001 = 0.0027$ second, or 2.7 milliseconds.

Transformers

Two coils having mutual inductance constitute a *transformer*. The coil connected to the source of energy is called the *primary* coil, and the other is called the *secondary* coil.

The usefulness of the transformer lies in the fact that electrical energy can be transferred from one circuit to another without direct connection, and in the process can be readily changed from one voltage ~~level~~ to another. Thus, if a device to be operated requires, for example, 7 volts ac and only a 440-volt source is available, a transformer can be used to change the source voltage to that required. A transformer can be used only with ac, since no voltage will be induced in the secondary if the magnetic field is not changing. If dc is applied to the primary of a transformer, a voltage will be induced in the secondary only at the instant of closing or opening the primary circuit, since it is only at these times that the field is changing.

The Iron-Core Transformer

As shown in Fig. 41, the primary and secondary coils of a transformer may be wound on a core of magnetic material. This increases the inductance of the coils so that a relatively small number of turns may be used to induce a given value of voltage with a small current. *A closed core* (one having a continuous magnetic path) such as that shown in Fig. 41 also tends to insure that practically all of the field set up by the current in the primary coil will cut the turns of the secondary coil. However, the core introduces a power loss because of hysteresis and eddy currents, so this type of construction is normally practicable only at power and audio frequencies. The discussion in this section is confined to transformers operating at such frequencies.

Voltage and Turns Ratio

For a given varying magnetic field, the voltage induced in a coil in the field will be proportional to the number of turns in the coil. If the two coils of a transformer are in the same field (which is the case when both are wound on the same closed core) it follows that the induced voltages will be proportional to the number of turns in each coil. In the primary the induced voltage is practically equal to, and opposes, the applied voltage, as described earlier. Hence,

$$E_s = \left(\frac{n_s}{n_p}\right) E_p$$

where E_s = secondary voltage
E_p = primary applied voltage
n_s = number of turns on secondary
n_p = number of turns on primary

The ratio, n_s/n_p is called the secondary-to-primary *turns ratio* of the transformer.

Example: A transformer has a primary of 400 turns and a secondary of 2800 turns, and an emf of 117 volts is applied to the primary.

$$E_s = \left(\frac{n_s}{n_p}\right) E_p = \frac{2800}{400} \times 117 = 7 \times 117$$

$$= 819 \text{ volts}$$

Also, if an emf of 819 volts is applied to the 2800-turn winding (which then becomes the primary) the output voltage from the 400-turn winding will be 117 volts.

Either winding of a transformer can be used as the primary, providing the winding has enough turns (enough inductance) to induce a voltage equal to the applied voltage without requiring an excessive current flow.

Effect of Secondary Current

The current that flows in the primary when no current is taken from the secondary is called the *magnetizing current* of the transformer. In any properly designed transformer the primary inductance will be so large that the magnetizing current will be quite small. The power consumed by the transformer when the secondary is "open" — that is, not delivering power — is only the amount necessary to supply the losses in the iron core and in the resistance of the wire with which the primary is wound.

When power is taken from the secondary winding, the secondary current sets

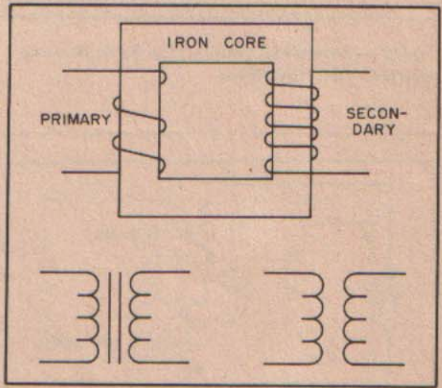


Fig. 41 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

up a magnetic field that opposes the field set up by the primary current. But if the induced voltage in the primary is to equal the applied voltage, the original field must be maintained. Consequently, the primary must draw enough additional current to set up a field exactly equal and opposite to the field set up by the secondary current.

In practical calculations on transformers it may be assumed that the entire primary current is caused by the secondary "load." This is justifiable because the magnetizing current should be very small in comparison with the primary "load" current at rated power output.

If the magnetic fields set up by the primary and secondary currents are to be equal, the primary current multiplied by the primary turns must equal the secondary current multiplied by the secondary turns. From this it follows that

$$I_p = \left(\frac{n_s}{n_p}\right) I_s$$

where I_p = primary current
I_s = secondary current
n_p = number of turns on primary
n_s = number of turns on secondary

Example: Suppose that the secondary of the transformer in the previous example is delivering a current of 0.2 ampere to a load. Then the primary current will be

$$I_p = \left(\frac{n_s}{n_p}\right) I_s = \frac{2800}{400} \times 0.2 = 7 \times 0.2$$

$$= 1.4 \text{ ampere}$$

Although the secondary voltage is higher than the primary voltage, the secondary current is lower than the primary current, and by the same ratio.

Power Relationships; Efficiency

A transformer cannot create power; it can only transfer it and change the emf. Hence, the power taken from the secondary cannot exceed that taken by the primary from the source of applied emf. There is always some power loss in the resistance of the coils and in the iron core, so in all practical cases the power taken from the source will exceed that taken from the secondary. Thus,

$$P_o = nP_i$$

where P_o = power output from secondary
P_i = power input to primary
n = efficiency factor

The efficiency, n, always is less than 1. It is usually expressed as a percentage; if n is 0.65, for instance, the efficiency is 65 percent.

Example: A transformer has an efficiency of 85 percent as its full-load output of 150 watts. The power input to the primary at full secondary load will be

$$P_i = \frac{P_o}{n} = \frac{150}{0.85} = 176.5 \text{ watts}$$

A transformer is usually designed to have the highest efficiency at the power output for which it is rated. The efficiency decreases with either lower or higher outputs. On the other hand, the *losses* in the transformer are relatively small at low output but increase as more power is taken. The amount of power that the transformer can handle is determined by its own losses, because these heat the wire and core. There is a limit to the temperature rise that can be tolerated, because a too-high temperature either will melt the wire or cause the insulation to break down. A transformer can be operated at reduced output, even though the efficiency is low, because the actual loss will be low under such conditions.

The full-load efficiency of small power transformers such as are used in radio receivers and transmitters usually lies between about 60 and 90 percent, depending upon the size and design.

Leakage Reactance

In a practical transformer not all of the magnetic flux is common to both windings, although in well-designed transformers the amount of flux that "cuts" one coil and not the other is only a small percentage of the total flux. This *leakage flux* causes an emf of self-induction; consequently, there are small amounts of *leakage inductance* associated with both windings of the transformer. Leakage inductance acts in exactly the same way as an equivalent amount of ordinary inductance inserted in series with the circuit. It has, therefore, a certain reactance, depending upon the amount of leakage inductance and the frequency. This reactance is called *leakage reactance*.

Current flowing through the leakage reactance causes a voltage drop. This voltage drop increases with increasing current, hence it increases as more power is taken from the secondary. Thus, the greater the secondary current, the smaller the secondary terminal voltage becomes. The resistances of the transformer windings also cause voltage drops when current is flowing; although these voltage drops are not in phase with those caused by leakage reactance, together they result in a lower secondary voltage under load than is indicated by the turns ratio of the transformer.

At power frequencies (60 Hz) the voltage at the secondary, with a reasonable well-designed transformer, should not drop more than about 10 percent from open-circuit conditions to full load. The drop in voltage may be considerably more than this in a transformer operating at audio frequencies because the leakage reactance increases directly with the frequency.

Impedance Ratio

In an ideal transformer — one without losses or leakage reactance — the following relationship is true:

$$Z_p = Z_s\left[\frac{N_p}{N_s}\right]^2$$

where Z_p = impedance looking into primary terminals from source of power
Z_s = impedance of load connected to secondary
N_p/N_s = turns ratio, primary to secondary

That is, a load of any given impedance connected to the secondary of the transformer will be transformed to a different value "looking into" the primary from the source of power. The impedance transformation is proportional to the square of the primary-to-secondary turns ratio.

Example: A transformer has a primary-to-secondary turns ratio of 0.6 (primary has 6/10 as many turns as the secondary) and a load of 3000 ohms is connected to the secondary. The impedance looking into the primary then will be

$$Z_p = Z_s\left[\frac{N_p}{N_s}\right] = 3000 \times (0.6)^2$$

$$= 3000 \times 0.36 = 1080 \text{ ohms}$$

By choosing the proper turns ratio, the impedance of a fixed load can be transformed to any desired value, within practical limits. If transformer losses can be neglected, the transformed or "reflected" impedance has the same phase angle as

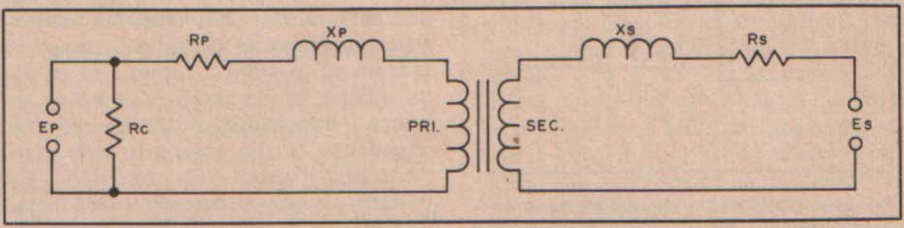


Fig. 42 — The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance R_C is an equivalent resistance representing the core losses, which are essentially constant for any given applied voltage and frequency. Since these are comparatively small, their effect may be neglected in many approximate calculations.

the actual load impedance; thus, if the load is a pure resistance, the load presented by the primary to the source of power also will be a pure resistance.

The above relationship may be used in practical work even though it is based on an "ideal" transformer. Aside from the normal design requirements of reasonably low internal losses and low leakage reactance, the only requirement is that the primary have enough inductance to operate with low magnetizing current at the voltage applied to the primary.

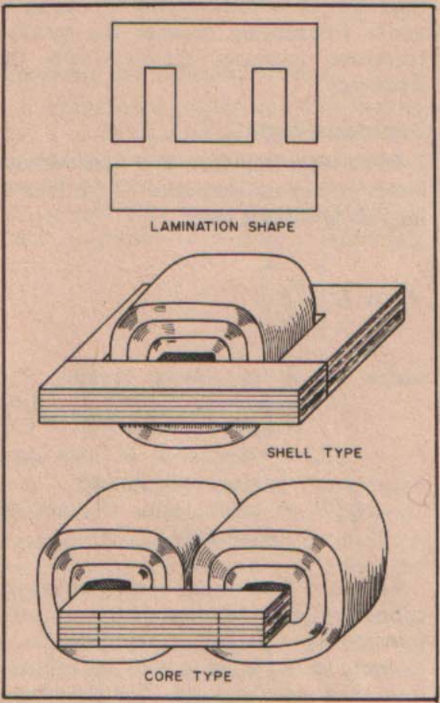


Fig. 43 — Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path.

The primary impedance of a transformer — *as it appears to the source of power* — is determined wholly by the load connected to the secondary and by the turns ratio. If the characteristics of the transformer have an appreciable effect on the impedance presented to the power source, the transformer is either poorly designed or is not suited to the voltage and frequency at which it is being used. Most transformers will operate quite well at voltages from slightly above to well below the design figure.

Impedance Matching

Many devices require a specific value of load resistance (or impedance) for optimum operation. The impedance of the actual load that is to dissipate the power may differ widely from this value, so a transformer is used to change the actual load into an impedance of the desired value. This is called *impedance matching*. From the preceding,

$$\frac{N_p}{N_s} = \sqrt{\frac{Z_p}{Z_s}}$$

where

N_p/N_s = required turns ratio, primary to secondary

Z_p = primary impedance required

Z_s = impedance of load connected to secondary

Example: A vacuum-tube af amplifier requires a load of 5000 ohms for optimum performance, and is to be connected to a loudspeaker having an impedance of 10 ohms. The turns ratio, primary to secondary, required in the coupling transformer is

$$\frac{N_p}{N_s} = \sqrt{\frac{Z_p}{Z_s}} = \sqrt{\frac{5000}{10}} = \sqrt{500} = 22.4$$

The primary therefore must have 22.4 times as many turns as the secondary.

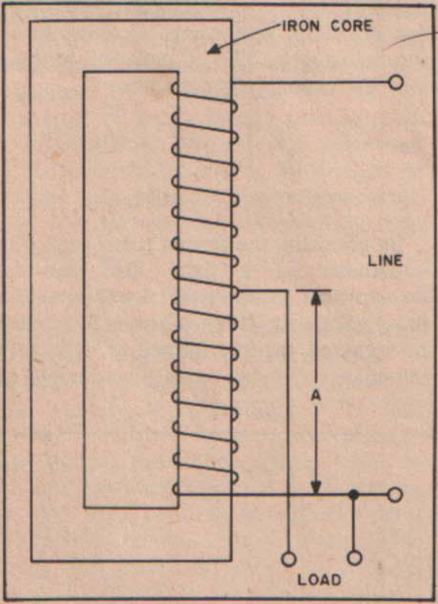


Fig. 44 — The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (A) flow in opposite directions, so that the resultant current is the difference between them. The voltage across A is proportional to the turns ratio.

Impedance matching means, in general, adjusting the load impedance — by means of a transformer or otherwise — to a desired value. However, there is also another meaning. It is possible to show that any source of power will deliver its maximum possible output when the impedance of the load is equal to the internal impedance of the source. The impedance of the source is said to be "matched" under this condition. The efficiency is only 50 percent in such a case; just as much power is used up in the source as is delivered to the load. Because of the poor efficiency, this type of impedance matching is limited to cases where only a small amount of power is available and heating from power loss in the source is not important.

Transformer Construction

Transformers usually are designed so that the magnetic path around the core is as short as possible. A short magnetic path means that the transformer will operate with fewer turns, for a given applied voltage, than if the path were long. A short path also helps to reduce flux leakage and therefore minimizes leakage reactance.

Two core shapes are in common use, as shown in Fig. 43. In the shell type both windings are placed on the inner leg, while in the core type the primary and secondary windings may be placed on separate legs, if desired. This is sometimes done when it is necessary to minimize capacitive effects between the primary and secondary, or when one of the windings must operate at very high voltage.

Core material for small transformers is usually silicon steel, called "transformer iron." The core is built up of laminations, insulated from each other (by a thin coating of shellac, for example) to prevent the flow of eddy currents. The laminations are interleaved at the ends to make the magnetic path as continuous as possible and thus reduce flux leakage.

The number of turns required in the primary for a given applied emf is determined by the size, shape and type of core material used, and the frequency. The number of turns required is inversely proportional to the cross-sectional area of the core. As a rough indication, windings of small power transformers frequently have about six to eight turns per volt on a core of 1-square-inch (645 sq. mm) cross section and have a magnetic path 10 or 12 inches (252 or 302 mm) in length. A longer path or smaller cross section requires more turns per volt, and vice versa.

In most transformers the coils are wound in layers, with a thin sheet of treated-paper insulation between each layer. Thicker insulation is used between coils and between coils and core.

Autotransformers

The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 44; the principles just discussed apply equally well. A one-winding transformer is called an *autotransformer*. The current in the common section (A) of the winding is the difference between the line (primary) and the load (secondary) currents, since these currents are out of phase. Hence, if the line and load currents are nearly equal, the common section of the winding may be wound with comparatively small wire. This will be the case only when the primary (line) and secondary (load) voltages are not very different. The autotransformer is used chiefly for boosting or reducing the power-line voltage by relatively small amounts. Continuously variable autotransformers are commercially available under a variety of trade names; "Variac" and "Powerstat" are typical examples.

Ferromagnetic Transformers and Inductors

The design concepts and general theory of transformers which is presented earlier in this chapter apply also to transformers which are wound on ferromagnetic core materials (ferrite and powdered iron). As is the case with stacked cores made of laminations in the classic I and E shapes, the core material has a specific permeability factor which determines the inductance of the windings versus the number of wire turns used. Both ferrite and powdered-iron materials are manufactured with a wide range of μ_i (initial permeability) characteristics. The value chosen by the designer will depend upon the intended operating frequency and the desired bandwidth of a given broadband transformer.

Core-Types in Common Use

For use in radio-frequency circuits especially, a suitable core type must be chosen to provide the Q required by the designer. The wrong core material destroys the Q of an rf type of inductor.

Toroid cores are useful from a few hundred hertz well into the uhf spectrum. Tape-wound steel cores are employed in some types of power supplies — notably dc-to-dc converters. The toroid core is doughnut shaped, hence the name *toroid* (Fig. 45). The principal advantage to this type of core is the self-shielding characteristic. Another feature is the compactness of a transformer or inductor, which is possible when using a toroidal format. Therefore, toroids are excellent not only in dc-to-dc converters, but at audio and radio frequencies up to at least 1000 MHz, assuming the proper core material is selected for the range of frequencies over which the device must operate. Toroid cores are available from micro-miniature sizes well up to several inches in diameter. The latter can be used, as one example, to build a 20-kW balun for use in antenna systems.

Another form taken in ferromagnetic transformers and inductors is the "pot-core" or "cup-core" device. Unlike the toroid, which has the winding over the outer surface of the core material, the pot-core winding is inside the ferromagnetic material (Fig. 46). There are two cup-shaped halves to the assembly, both made of ferrite or powdered iron, which are connected tightly together by means of a screw which is passed through a center hole. The wire for the assembly is wound on an insulating bobbin which fits inside the two halves of the pot-core unit. The advantage to this type of construction is that the core permeability can be chosen to ensure a minimum number of wire turns for a given value of inductance. This reduces the wire resistance and increases the Q as opposed to an equivalent inductance wnich is wound on a core that has relatively low permeability, or none at all. By virtue of the winding being contained inside the ferrite or powdered-iron pot core, shielding is excellent.

Still another kind of ferromagnetic-core inductor is found in today's technology — the solenoidal type (Fig. 47). Transformers and inductors fabricated in this manner consist of a cylindrical, oval or rectangular rod of material over which the wire winding is placed. This variety of device does not have a self-shielding trait. Therefore it must be treated in the same manner as any solenoidal-wound inductor (using external shield devices). An example of a ferrite-rod inductor is the built-in loop antennas found in portable radios and direction finders.

Fig. 45 — An assortment of toroid cores. A ferrite rod is placed at the top of the picture for comparison. The two light-colored, plastic-encased toroids at the upper left are tape-wound types (Hypersil steel) are suitable for audio and dc-to-dc converter transformers. The wound toroid at the right center contains two toroid cores which have been stacked atop one another to increase the power capability.

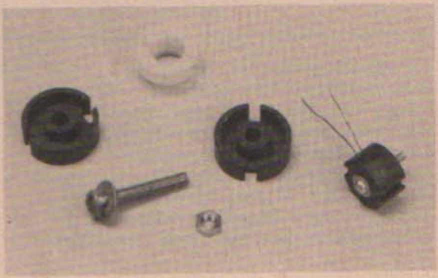


Fig. 46 — Breakaway view of a pot-core assembly (left) and an assembled pot core (right).

Fig. 47 — A bc-band ferrite rod loop antenna is at the top of the picture (J. W. Miller Co.) A blank ferrite rod is seen at the center and a flat bc-band ferrite loop antenna is in the lower foreground.

Core Size

The cross-sectional area of ferromagnetic core is chosen to prevent saturation from the load seen by the transformer. This means that the proper thickness and diameter are essential parameters to consider. For a specific core the maximum operational ac excitation can be detemined by

$$B_{op\,(ac)} = \frac{E_{rms} \times 10^8}{4.44\ fN_pA_e} \text{ (gauss)}$$

where A_e = equivalent area of the magnetic path in cm^2
E_{rms} = applied voltage
N_p = number of core turns
f = operating frequency in Hz
B_{max} = maximum flux density in gauss

The foregoing equation is applicable to inductors which do not have dc flowing in the winding along with ac. When both ac and dc currents flow

$$B_{op(total)} = \frac{E_{rms} \times 10^8}{4.444\ fN_pA_e} + \frac{N_pI_{dc}A_L}{10A_e}$$

where I_{dc} = the dc current through the winding
A_L = the manufacturer's index for the core being used

The latter can be obtained for the core in use by consulting the manufacturer's data sheet.

Types of Transformers

The most common ferromagnetic transformers used in amateur radio work are the narrow-band, broadband, conventional and transmission-line varieties. *Narrow-band* transformers are used when selectivity is desired in a tuned circuit, such as an audio peaking or notching circuit, a resonator in an rf filter, or a tuned circuit associated with an rf amplifier. *Broadband* transformers are employed in circuits which must have uniform response over a substantial spread of frequency, as in a 2- to 30-MHz broadband amplifier. In such an example the reactance of the windings should be at least four times the impedance the winding is designed to look into. Therefore, a transformer which has a 300-ohm primary and a 50-ohm secondary load should have winding reactances (X_L) of at least 1200 ohms and 200 ohms, respectively. The windings, for all practical purposes, can be regarded as rf chokes, and the same rules apply. The permeability of the core material plays a vital role in designing a good broadband transformer. The performance of the transformer at the low-frequency end of the operating range

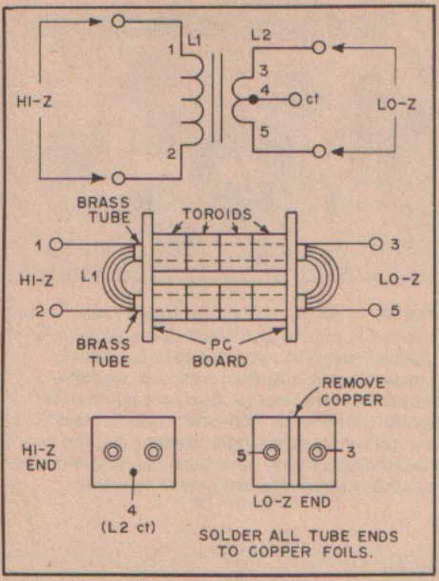


Fig. 48 — Schematic and pictorial representations of one type of "conventional" transformer. This style is used frequently at the input and output ports of rf power amplifiers which use transistors. The magnetic material consists of two rows of 950-mu toroid cores for use from 1.8 to 30 MHz. The primary and secondary windings are passed through the center holes of the toroid-stack rows as shown.

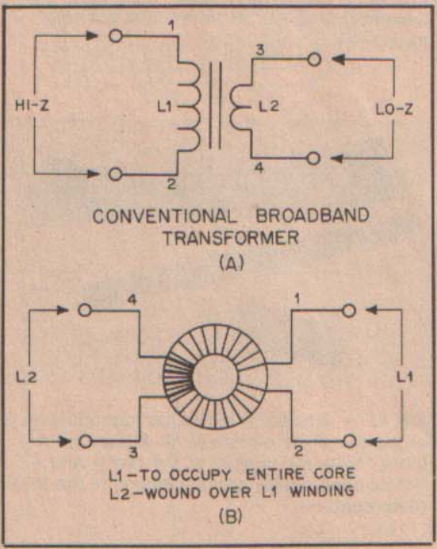


Fig. 49 — Another conventional transformer. Primary and secondary windings are wound over the outer surface of a toroid core.

depends on the permeability. That is, the μ_e (effective permeability) must be high enough in value to provide ample winding reactance at the low end of the operating range. As the operating frequency is increased, the effects of the core tend to disappear progressively until there are scarcely any core effects at the upper limit of the operating range. For this reason it is common to find a very low frequency core material utilized in a transformer that is contained in a broadband circuit which reaches well into the upper hf region, or

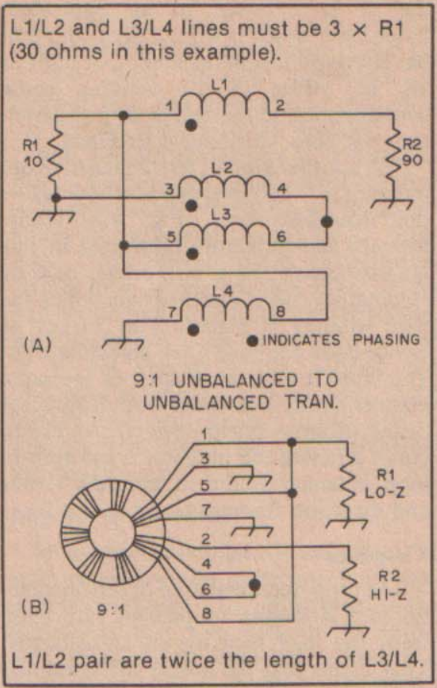


Fig. 50 — Schematic and pictorial presentations of a transmission-line transformer in which the windings need to be configured for a specific impedance.

even into the vhf spectrum. By way of simple explanation, at high frequency the low-frequency core material becomes inefficient and tends to vanish electrically. This desirable trait makes possible the use of ferromagnetics in broadband applications.

Conventional transformers are those that are wound in the same manner as a power transformer. That is, each winding is made from a separate length of wire, with one winding being placed over the previous one with suitable insulation in between (Figs. 48 and 49). A *transmission-line transformer* is, conversely, one that uses windings which are configured to simulate a piece of transmission line of a specific impedance. This can be achieved by twisting the wires together a given number of times per inch, or by laying the wires on the core (adjacent to one another) at a distance apart which provides a two-wire line impedance of a particular value. In some applications these windings are called *bifilar*. A three-wire winding is known as a *trifilar* one, and so forth (Fig. 50). It can be argued that a transmission-line transformer is more efficient than a conventional one, but in practice it is difficult to observe a significant difference in the performance characteristics. An interesting technical paper on the subject of toroidal broadband transformers was published by Sevick, W2FMI.[1] The classic reference on the subject is by Ruthroff.[2]

Ferrite Beads

Another form of toroidal inductor is the *ferrite bead*. This component is available in various μ_i values and sizes, but most beads are less than 0.25-inch (6.3-mm) diameter. Ferrite beads are used principally as vhf/uhf parasitic suppressors at the input and output terminals of amplifiers. Another practical application for them is in decoupling networks which are used to prevent unwanted migration of rf energy from one section of a circuit to another. They are used also in suppressing RFI and TVI in hi-fi and television sets. In some circuits it is necessary to only place one or more beads over a short length of wire to obtain ample inductive reactance for creating an rf choke. A few turns of small-diameter enameled wire can be looped through the larger beads to increase the effective inductance. Ferrite beads are suitable as low-Q base impedances in solid-state vhf and uhf amplifiers. The low-Q characteristics prevents self-oscillation that might occur if a high Q solenoidal rf choke was used in place of one which was made from beads. Miniature broadband transformers are sometimes fashioned from ferrite beads. For the most part, ferrite beads can be regarded as small toroid cores.

Number of Turns

The number of wire turns used on a toroid core can be calculated by knowing the A_L of the core and the desired inductance. The A_L is simply the *inductance index* for the core size and permeability being used. Table 6 provides information of interest concerning a popular assortment of powdered-iron toroid cores. The complete number for a given core is composed of the core-size designator in the upper left column, plus the corresponding mix number. For example, a half-inch diameter core with a no. 2 mix would be designated at a T-50-2 unit. The A_L would be 49 and the suggested operating frequency would be from 1 to 30 MHz. The μ_i for that core is 10.

The required number of wire turns for a specified inductance on a given type of core can be determined by

$$\text{Turns} = 100\sqrt{\text{desired L } (\mu\text{H}) \div A_L}$$

where A_L is obtained from Table 6. The table also indicates how many turns of a particular wire gauge can be close wound to fill a specified core. For example, a T-68 core will contain 49 turns of no. 24 enameled wire, 101 turns of no. 30 enameled wire, and so on. Generally speaking, the larger the wire gauge the higher the unloaded Q of the toroidal inductor. The inductance values are based on the winding covering the entire circumference of the core. When there is space between the turns of wire, some control over the net inductance can be effected by compressing the turns or

[1]Sevick, "Simple Broadband Matching Networks," *QST*, January 1976.
[2]Ruthroff, "Some Broadband Transformers," *Proc. IRE*, Vol. 47, August 1959, p. 137.

Table 6

Powdered-Iron Toroidal Cores — A_L Values (µH/100 turns)

Core Size	41-Mix Green µ = 75	3-Mix Grey µ = 35 0.05-0.5 MHz	15-Mix Rd & Wh µ = 25 0.1-2 MHz	1-Mix Blue µ = 20 0.5-5 MHz	2-Mix Red µ = 10 1-30 MHz	6-Mix Yellow µ = 8 10-90 MHz	10-Mix Black µ = 6 60-150 MHz	12-Mix Gn & Wh µ = 3 100-200 MHz	0-Mix Tan µ = 1 150-300 MHz
T-200	755	360	NA	250*	120	100*	NA	NA	NA
T-184	1640	720	NA	500*	240	195	NA	NA	NA
T-157	970	420	360*	320*	140	115	NA	NA	NA
T-130	785	330	250*	200	110	96	NA	NA	15.0
T-106	900	405	345*	325*	135	116	NA	NA	19.0*
T- 94	590	248	200*	160	84	70	58	32	10.6
T- 80	450	180	170	115	55	45	32*	22	8.5
T- 68	420	195	180	115	57	47	32	21	7.5
T- 50	320	175	135	100	49*	40	31	18	6.4
T- 44	229	180	160	105	52*	42	33	NA	6.5
T- 37	308	120*	90	80	40*	30	25	15	4.9
T- 30	375	140*	93	85	43	36	25	16	6.0
T- 25	225	100	85	70	34	27	19	13	4.5
T- 20	175	90	65	52	27	22	16	10	3.5
T- 16	130	61	NA	44	22	19	13	8	3.0
T- 12	112	60	50*	48	20*	17*12	7.5	3.0	

NA — Not available in that size.
Turns = $100\sqrt{L_{\mu H} \div A_L \text{Value (above.)}}$
All frequency figures optimum.

*Updated values (1979) from Micrometals, Inc.

Number of Turns vs. Wire Size and Core Size
Approximate maximum of turns — single layer wound enameled wire

Wire Size	T-200	T-130	T-106	T-94	T-80	T-68	T-50	T-37	T-25	T-12
10	33	20	12	12	10	6	4	1		
12	43	25	16	16	14	9	6	3		
14	54	32	21	21	18	13	8	5	1	
16	69	41	28	28	24	17	13	7	2	
18	88	53	37	37	32	23	18	10	4	1
20	111	67	47	47	41	29	23	14	6	1
22	140	86	60	60	53	38	30	19	9	2
24	177	109	77	77	67	49	39	25	13	4
26	223	137	97	97	85	63	50	33	17	7
28	281	173	123	123	108	80	64	42	23	9
30	355	217	154	154	136	101	81	54	29	13
32	439	272	194	194	171	127	103	68	38	17
34	557	346	247	247	218	162	132	88	49	23
36	683	424	304	304	268	199	162	108	62	30
38	875	544	389	389	344	256	209	140	80	39
40	1103	687	492	492	434	324	264	178	102	51

Physical Dimensions

Core Size	Outer Dia. (in.)	Inner Dia. (in.)	Height (in.)	Cross Sect. Area cm²	Mean Length cm	Core Size	Outer Dia. (in.)	Inner Dia. (in.)	Height (in.)	Cross Sect. Area cm²	Mean Length cm
T-200	2.000	1.250	0.550	1.330	12.97	T- 50	0.500	0.303	0.190	0.121	3.20
T-184	1.840	0.950	0.710	2.040	11.12	T- 44	0.440	0.229	0.159	0.107	2.67
T-157	1.570	0.950	0.570	1.140	10.05	T- 37	0.375	0.205	0.128	0.070	2.32
T-130	1.300	0.780	0.437	0.733	8.29	T- 30	0.307	0.151	0.128	0.065	1.83
T-106	1.060	0.560	0.437	0.706	6.47	T- 25	0.255	0.120	0.096	0.042	1.50
T- 94	0.942	0.560	0.312	0.385	6.00	T- 20	0.200	0.088	0.067	0.034	1.15
T- 80	0.795	0.495	0.250	0.242	5.15	T- 16	0.160	0.078	0.060	0.016	0.75
T- 68	0.690	0.370	0.190	0.196	4.24	T- 12	0.125	0.062	0.050	0.010	0.74

Inches × 25.4 = mm. Courtesy of Amidon Assoc., N. Hollywood, CA 91607 and Micrometals, Inc.

spreading them. The inductance will increase if compression is used and will decrease when the turns are spread farther apart.

Table 7 contains data for ferrite cores. The number of turns for a specified inductance in mH versus the A_L can be determined by

$$\text{Turns} = 1000\sqrt{\text{desired L (mH)} \div A_L}$$

where the A_L for a specific core can be taken from Table 7. Thus, if one required a 1-mH inductor and chose a no. FT-82-43 toroid core, the number of turns would be

$$\text{Turns} = 1000\sqrt{1 \div 557}$$
$$= 1000\sqrt{0.001795}$$
$$= 1000 \times 0.0424 = 42.4 \text{ turns}$$

For an FT-82 size core no. 22 enameled wire would be suitable as indicated in Table 6 (using the T-80 core size as the nearest one to an FT-82). If the toroid core has rough edges (untumbled), it is suggested that insulating tape (3M glass epoxy tape or Mylar tape) be wrapped through the core before the wire is added. This will prevent the rough edges of the core from abrading the enameled wire.

Checking RF Toroidal Devices

The equations given previously will provide the number of wire turns needed for a particular inductance, plus or minus 10 percent. However, slight variations in core permeability may exist from one production run to another. Therefore, for

Table 7

Ferrite Toroids **A_L - Chart (mH per 1000 turns) Enameled Wire**

Core Size	*63-Mix* $\mu = 40$	*61-Mix* $\mu = 125$	*43-Mix* $\mu = 950$	*72-Mix* $\mu = 2000$	*75-Mix* $\mu = 5000$
FT- 23	7.9	24.8	189.0	396.0	990.0
FT- 37	17.7	55.3	420.0	884.0	2210.0
FT- 50	22.0	68.0	523.0	1100.0	2750.0
FT- 82	23.4	73.3	557.0	1172.0	2930.0
FT-114	25.4	79.3	603.0	1268.0	3170.0

Number turns $= 1000\sqrt{\text{desired L (mH)} \div A_L \text{ value (above)}}$

Ferrite Magnetic Properties

Property	*Unit*	*63-Mix*	*61-Mix*	*43-Mix*	*72-Mix*	*75-Mix*
Initial Perm. (μi)		40	125	950	2000	5000
Maximum Perm.		125	450	3000	3500	8000
Saturation Flux Density @ 13 oer	Gauss	1850	2350	2750	3500	3900
Residual Flux Density	Gauss	750	1200	1200	1500	1250
Curie Temp.	°C	500	300	130	150	160
Vol. Resistivity	ohm/cm	1×10^8	1×10^8	1×10^5	1×10^2	5×10^2
Opt. Freq. Range	MHz	15-25	.2-10	.01-1	.001-1	.001-1
Specific Gravity		4.7	4.7	4.5	4.8	4.8
Loss Factor	$\frac{1}{u\,O}$	9.0×10^{-5} @ 25 MHz.	2.2×10^{-5} @ 2.5 MHz	2.5×10^{-5} @ .2 MHz	9.0×10^{-6} @ .1 MHz	5.0×10^{-6} @ .1 MHz
Coercive Force	Oer.	2.40	1.60	0.30	0.18	0.18
Temp. Co-eff of initial Perm.	%/°C 20-70°C	0.10	0.10	0.20	0.60	

Ferrite Toroids **Physical Properties**

Core Size	*OD*	*ID*	*Height*	A_e	l_e	V_e	A_s	A_w
FT- 23	0.230	0.120	0.060	0.00330	0.529	0.00174	0.1264	0.01121
FT- 37	0.375	0.187	0.125	0.01175	0.846	0.00994	0.3860	0.02750
FT- 50	0.500	0.281	0.188	0.02060	1.190	0.02450	0.7300	0.06200
FT- 82	0.825	0.520	0.250	0.03810	2.070	0.07890	1.7000	0.21200
FT-114	1.142	0.748	0.295	0.05810	2.920	0.16950	2.9200	0.43900

OD - Outer diameter (inches)
ID - Inner diameter (inches)
Hgt - Height (inches)
A_W - Total window area $(in)^2$
A_e - Effective magnetic cross-sectional area $(in)^2$
l_e - Effective magnetic path length (inches)
V_e - Effective magnetic volume $(in)^3$
A_s - Surface area exposed for cooling $(in)^2$

Inches $\times$ 25.4 = mm. Courtesy of Amidon Assoc., N. Hollywood, CA 91607

circuits which require exact values of inductance it is necessary to check the toroid winding by means of an RCL bridge or an RX meter. If these instruments are not available, close approximations can be had by using a dip meter, standard capacitor (known value, stable type, such as a silver mica) and a calibrated receiver against which to check the dipper frequency. Fig. 51A shows how to couple a dip meter to a completed toroid for testing. The coupling link in the illustration is necessary because the toroid has a self-shielding characteristic. The latter makes it difficult, and often impossible, to secure a dip in the meter reading when coupling the instrument directly to the toroidal inductor or transformer. The inductance can be determined by X_L since $X_L = X_c$ at resonance. Therefore,

$$X_c = \frac{1}{2\pi fC} \text{ and } L_{(\mu H)} = \frac{X_L}{2\pi f}$$

where X_c is the reactance of the known capacitor value, f is in MHz and C is in μF. Using an example, where f is 3.5 MHz (as noted on a dip meter) and C is 100 PF, L is determined by

$$X_c = \frac{1}{6.28 \times 3.5 \times 0.0001} = 455 \text{ ohms}$$

Since $X_L = X_c$ at resonance,

$$L_{(\mu H)} = \frac{455}{6.26 \times 3.5} = 20.7\ \mu H$$

It is assumed, for the purpose of accuracy, that the dip-meter signal is checked for precise frequency by means of a calibrated receiver.

Practical Considerations

Amateurs who work with toroidal inductors and transformers are sometimes confused by the winding instructions given in construction articles. For the most part, winding a toroid core with wire is less complicated than it is when winding a cylindrical single-layer coil.

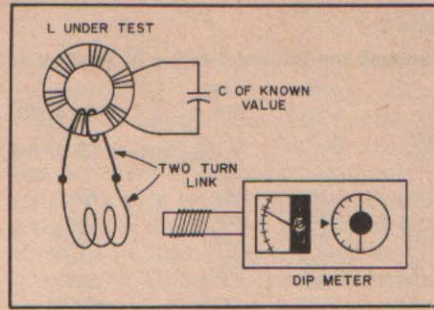


Fig. 51A — Method for checking the inductance of a toroid winding by means of a dip meter, known capacitance value and a calibrated receiver. The self-shielding properties of a toroidal inductor prevent dip-meter readings when the instrument is coupled directly to the toroid. Sampling is done by means of a coupling link as illustrated.

When many turns of wire are required, a homemade winding shuttle can be used to simplify the task. Fig. 51B-(A) illustrates how this method may be employed. The shuttle can be fashioned from a piece of circuit-board material. The wire is wound on the shuttle after determining how many inches are required to provide the desired number of toroid turns. (A sample turn around the toroid core will reveal the wire length per turn.) Once the shuttle is loaded, it is passed through the toroid center again and again until the winding is completed. The edges of the shuttle should be kept smooth to prevent abrasion of the wire insulation.

How to Wind Toroids

The *effective inductance* of a toroid coil or a transformer winding is dependent in part upon the distributed capacitance between the coil turns and between the ends of the winding. When a large number of turns are used (e.g., 500 or 1000), the distributed capacitance can be as great as 100 pF. Ideally, there would be no distributed or "parasitic" capacitance, but this is not possible. Therefore, the unwanted capacitance must be kept as low as possible in order to take proper advantage of the A_L factors discussed earlier in this section. The greater the distributed capacitance the more restrictive the transformer or inductor becomes when applied in a broadband circuit. In the case of a narrow-band application, the Q can be affected by the distributed capacitance. The pictorial illustration at Fig. 51B-(B) shows the inductor turns distributed uniformly around the toroid core, but a gap of approximately 30° is maintained between the ends of the winding. This method is recommended to reduce the distributed capacitance of the winding. The closer the ends of the winding are to one another, the greater the unwanted capacitance. Also, in order to closely approximate the desired toroid inductance when using the A_L formula, the winding should be spread over the core as shown. When the turns of the winding are not close wound, they can be spread apart to *decrease* the effective inductance (this lowers the distributed C). Conversely, as the turns are pushed closer together, the effective inductance is *increased* by virtue of the greater distributed capacitance. This phenomenon can be used to advantage during final adjustment of narrow-band circuits in which toroids are used.

The proper method for counting the turns on a toroidal inductor is shown in Fig. 51B-(C). The core is shown as it would appear when stood on its edge with the narrow dimension toward the viewer. In this example a four-turn winding has been placed on the core.

Some manufacturers of toroids recommend that the windings on toroidal transformers be spread around all of the core in the manner shown in Fig. 51B-(B). That is, the primary and secondary windings should each be spread around most of the core. This is a proper method when winding conventional broadband transformers. However, it is not recommended when narrow-band transformers are being built. It is better to place the low-impedance winding (L1 of Fig. 51B-(D)) at the "cold" or ground end of L2 on the core. This is shown in pictorial and schematic form at Fig. 51B-(D). The windings are placed on the core in the same rotational sense, and L1 is wound over L2 at the grounded end of L2. The purpose of this winding method is to discourage unwanted *capacitive* coupling between the windings — an aid to the reduction of spurious energy (harmonics, etc.) which might be present in the circuit where T1 is employed.

In circuits which have a substantial amount of ac and/or dc voltage present in the transformer windings, it is prudent to use a layer of insulating material between the toroid core and the first winding. Alternatively, the wire can have high-dielectric insulation, such as Teflon. This procedure will prevent arcing between the winding and the core. Similarly, a layer of insulating tape (3-M glass tape, mylar or Teflon) can be placed between the primary and secondary windings of the toroidal transformer (Fig. 51B-(D)). Normally, these precautions are not necessary at impedance levels under a few hundred ohms at rf power levels below 100 watts.

Once the inductor or transformer is wound and tested for proper performance, a coating or two of high-dielectric cement should be applied to the winding(s) of the toroid. This will protect the wire insulation from abrasion, hold the turns in place and seal the assembly against moisture and dirt. Polystyrene Q Dope is excellent for the purpose.

The general guidelines given for toroidal components can be applied to pot cores and rods when they are used as foundations for inductors or transformers. The important thing to remember is that all of the powdered-iron and ferrite core materials are brittle. They break easily under stress.

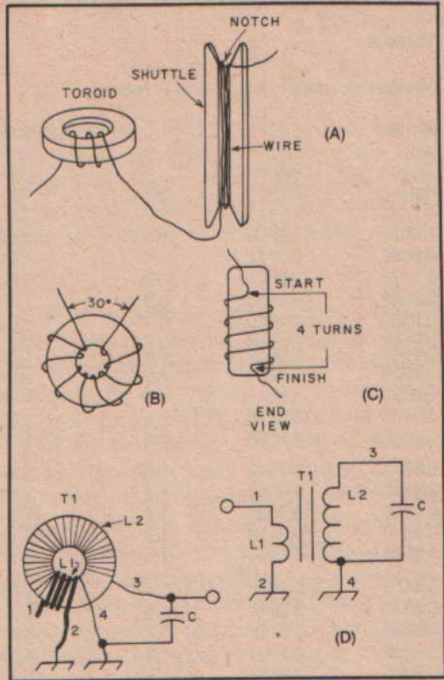


Fig. 51B — (A) Illustration of a homemade winding shuttle for toroids. The wire is stored on the shuttle and the shuttle is passed through the center hole of the toroid, again and again, until the required number of turns is in place. (B) It is best to leave a 30° gap between the ends of the toroid winding. This will reduce the distributed capacitance considerably. (C) Edgewise view of a toroid core, illustrating the method for counting the turns accurately. (D) The low-impedance winding of a toroidal transformer is usually wound over the larger winding, as shown. For narrow-band applications the link should be wound over the cold end of the main winding (see text).

Ac-Operated Power Supplies

Power-line voltages have been "standardized" throughout the U.S. at 117 and 234 volts in residential areas where a single phase voltage is supplied. These figures represent nominal voltages. "Normal" line voltage in a particular area may be between approximately 110 and 125 volts, but generally will be above 115 volts. In many states the service is governed by a PUC (public utilities commission). The voltage average across the country is approximately 117.

The ac-current capability of the service is a factor of line length from the dwelling to the nearest pole transformer, plus the conductor size of the line. Many older homes are supplied with a 60-ampere service while most new homes have 100 amperes. Houses equipped with electric heat will have services ranging from 150 to 200 amperes.

The electrical power required to operate Amateur Radio equipment is usually taken from the ac lines when the equipment is operated where power is available. For mobile operation the source of power is almost always the car storage battery.

Dc voltages used in transmitters, receivers and other related equipment are derived from the commercial ac lines by using a transformer-rectifier-filter system. The transformer changes the ac voltage to a suitable value and the rectifier converts the ac to pulsating dc. A filter is used to smooth out these pulsations to an acceptably low level. Essentially pure direct current is required to prevent 60- or 120-Hz hum in most pieces of amateur equipment. Transmitters must be operated from a pure dc supply as dictated by federal regulations. If a constant voltage is required under conditions of changing load or ac-line voltage, a regulator is used following the filter.

When the prime power source is dc (a battery), the dc is used directly or is first changed to ac and is then followed by the transformer-rectifier-filter combination. The latter system has lost considerable popularity with the advent of low-voltage semiconductor devices.

Transformerless power supplies are used in some applications (notably ac-dc radios and some television receivers). Supplies of this sort operate directly from the power line, making it necessary to connect the chassis or common-return point of the circuit directly to one side of the ac line. This type of power supply represents a shock hazard when the equipment is connected to other units in the amateur station or when the chassis is exposed. For safety reasons, an isolation transformer should be used with such equipment.

Power-Line Considerations: Connections

In most residential systems, three wires are brought in from the outside to the distribution board, while in a few older systems there are only two wires. In the three-wire system, the third wire is the *neutral*, which is grounded. The voltage between the two wires normally is 234, while half of this voltage appears between each of these wires and neutral, as indicated in Fig. 1A. In systems of this type the 117-volt household load is divided as evenly as possible between the two sides of the circuit, half of the load being connected between one wire and the neutral, while the other half of the load is connected between the other wire and neutral. Heavy appliances, such as electric stoves and heaters are designed for 234-volt operation and therefore are connected across the two ungrounded wires. While both ungrounded wires should be fused, a fuse should never be used in the neutral wire, nor should a switch be used in this side of the line. The reason for this is that opening the neutral wire does not disconnect the equipment. It simply leaves the equipment on one side of the 234-volt circuit in series with whatever load may be across the other side of the circuit, as shown in Fig. 1B. Furthermore, with the neutral open, the voltage will then be divided between the two sides in inverse proportion to the load resistance, the voltage on one side dropping below normal, while it soars on the other side, unless the loads happen to be equal.

The usual line running to baseboard outlets is rated at 15 amperes. Considering the power consumed by filaments, lamps, transmitter, receiver and other auxiliary

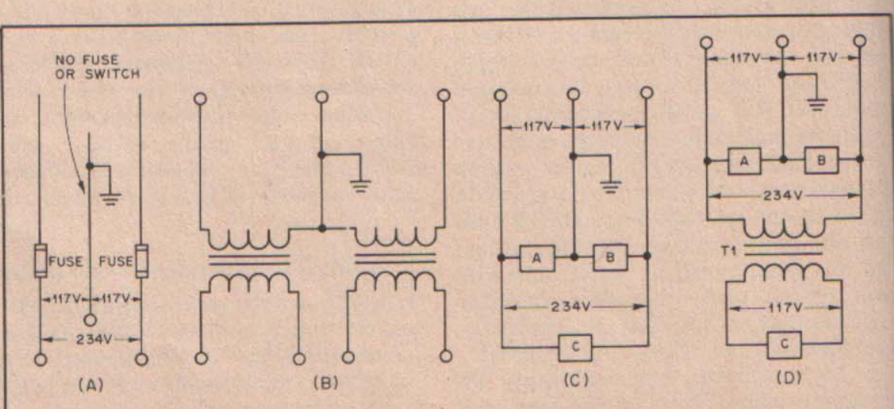


Fig. 1 — Three-wire power-line circuits. At A — Normal three-wire-line termination. No fuse should be used in the grounded (neutral) line. B — A switch in the neutral does not remove voltage from either side of the line. C — Connections for both 117- and 234-volt transformers. D — Operating a 117-volt plate transformer from the 234-volt line to avoid light blinking. T1 is a 2:1 step-down transformer.

equipment, it is not unusual to find this 15-A rating exceeded by the requirements of a station of only moderate power. It must also be kept in mind that the same branch may be in use for other household purposes through another outlet. For this reason, and to minimize light blinking when keying or modulating the transmitter, a separate heavier line should be run from the distribution board to the station whenever possible. A 3-volt drop in line voltage will cause noticeable blinking of lights.

If the system is of the three-wire, 234-V type, the three wires should be brought into the station so that the load can be distributed to keep the line balanced. The voltage across a fixed load on one side of the circuit will increase as the load current on the other side is increased. The rate of increase will depend upon the resistance introduced by the neutral wire. If the resistance of the neutral is low, the increase will be correspondingly small. When the currents in the two circuits are balanced, no current flows in the neutral wire and the system is operating at maximum efficiency.

Light blinking can be minimized by using transformers with 234-volt primaries in the power supplies for the keyed or intermittent part of the load, connecting them across the two ungrounded wires with no connection to the neutral, as shown in Fig. 1C. The same can be accomplished by the insertion of a step-down transformer with its primary operating at 234 volts and secondary delivering 117 volts. Conventional 117-volt transformers may be operated from the secondary of the step-down transformer (see Fig. 1D).

When a special heavy-duty line is to be installed, the local power company should be consulted as to local requirements. In some localities it is necessary to have such a job done by a licensed electrician, and there may be special requirements to be met. Some amateurs terminate the special line to the station at a switch box, while others may use electric-stove receptacles as the termination. The power is then distributed around the station by means of conventional outlets at convenient points. All circuits should be properly fused.

Three-Wire 117-V Power Cords

To meet the requirements of state and national codes, electrical tools, appliances and many items of electronic equipment now being manufactured to operate from the 117-volt line, they must be equipped with a three-conductor power cord. Two of the conductors carry power to the device in the usual fashion, while the third conductor is connected to the case or frame.

When plugged into a properly wired mating receptacle, the three-contact polarized plug connects this third conductor to an earth ground, thereby grounding the chassis or frame of the appliance and preventing the possibility of electrical shock to the user. All commercially manufactured items of electronic test equipment and most ac-operated amateur equipment are being supplied with these three-wire cords. Adapters are available for use where older electrical installations do not have mating receptacles. For proper grounding, the lug of the green wire protruding from the adapter must be attached underneath the screw securing the cover plate of the outlet box where connection is made, and the outlet box itself must be grounded.

Fusing

All transformer primary circuits should be properly fused. To determine the approximate current rating of the fuse or circuit breaker to be used, multiply each current being drawn from the supply in amperes by the voltage at which the current is being drawn. Include the current taken by bleeder resistances and voltage dividers. In the case of series resistors, use the source voltage, not the voltage at the equipment end of the resistor. Include filament power if the transformer is supplying filaments. After multiplying the various voltages and currents, add the individual products. Then divide by the line voltage and add 10 or 20 percent. Use a fuse or circuit breaker with the nearest larger current rating.

Line-Voltage Adjustment

In certain communities trouble is sometimes experienced from fluctuations in line voltage. Usually these fluctuations are caused by a variation in the load on the line. Since most of the variation comes at certain fixed times of the day or night, such as the times when lights are turned on at evening, they may be taken care of by the use of a manually operated compensating device. A simple arrangement is shown in Fig. 2A. A tapped transformer is used to boost or buck the line voltage as required. The transformer should have a secondary varying between 6 and 20 volts in steps of 2 or 3 volts and its secondary should be capable of carrying the full load current.

The secondary is connected in series with the line voltage and, if the phasing of the windings is correct, the voltage applied to the primaries of the transmitter transformers can be brought up to the rated 117 volts by setting the transformer tap switch on the right tap. If the phasing of the two windings of the transformer happens to be reversed, the voltage will be reduced instead of increased. This connection may be used in cases where the line voltage may be above 117 volts. This method is preferable to using a resistor in the primary of a power transformer since it does not affect the voltage regulation as seriously. The circuit of 2B illustrates the use of a variable autotransformer (Variac) for adjusting line voltage.

Constant-Voltage Transformers

Although comparatively expensive, special transformers called *constant-voltage transformers* are available for use in cases where it is necessary to hold line voltage and/or filament voltage constant with fluctuating supply-line voltage. These are static-magnetic voltage regulating transformers operating on principles of ferroresonance. They have no tubes or moving parts, and require no manual adjustments. These transformers are rated over a range of less than 1 volt-ampere (VA) at 5 volts output up to several thousand VA at 117 or 234 volts. On the average they will hold their output voltages within one percent under an input voltage variation of ±15 percent.

Safety Precautions

All power supplies in an installation should be fed through a single main power-line switch so that all power may be cut off quickly, either before working on the equipment, or in case of an accident. Spring-operated switches or relays are not sufficiently reliable for this important service. Foolproof devices for cutting off all power to the transmitter and other equipment are shown in Fig. 3. The arrangements shown in Figs. 3A and B are similar circuits for two-wire (117-volt) and three-wire (234-volt) systems. S is an enclosed double-throw switch of the sort usually used as the entrance switch in house installations. J is a standard ac outlet and P a shorted plug to fit the outlet. The switch should be located

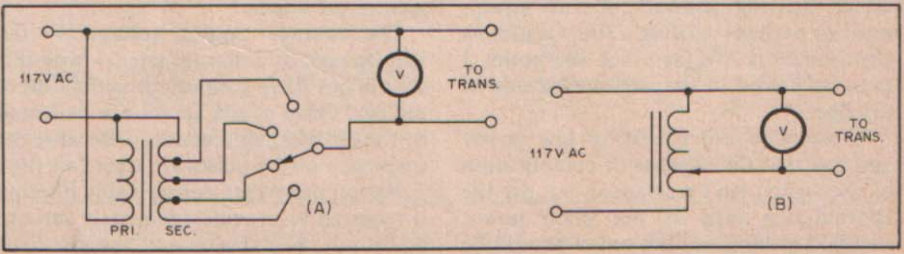


Fig. 2 — Two methods of transformer primary control. At A is a tapped transformer which may be connected so as to boost or buck the line voltage as required. At B is indicated a variable transformer or autotransformer (Variac) which feeds the transformer primaries.

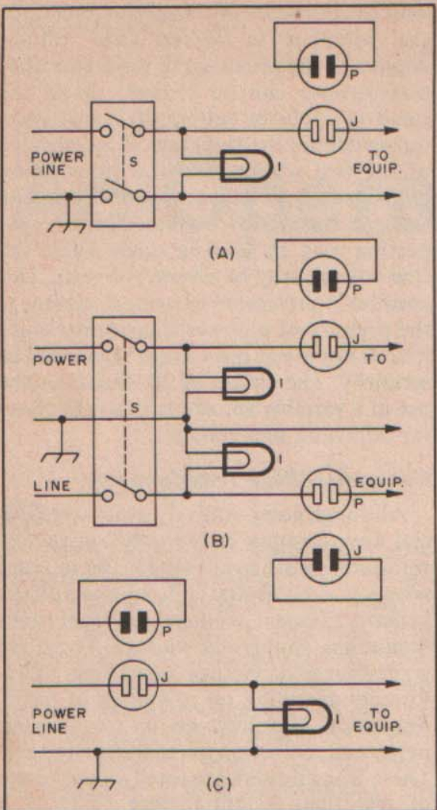


Fig. 3 — Reliable arrangements for cutting off all power to the transmitter. S is an enclosed double-pole power switch, J a standard ac outlet, P a shorted plug to fit the outlet and I a red lamp.
A is for a two-wire 117-volt line, B for a three-wire 234-volt system, and C a simplified arrangement for low-power stations.

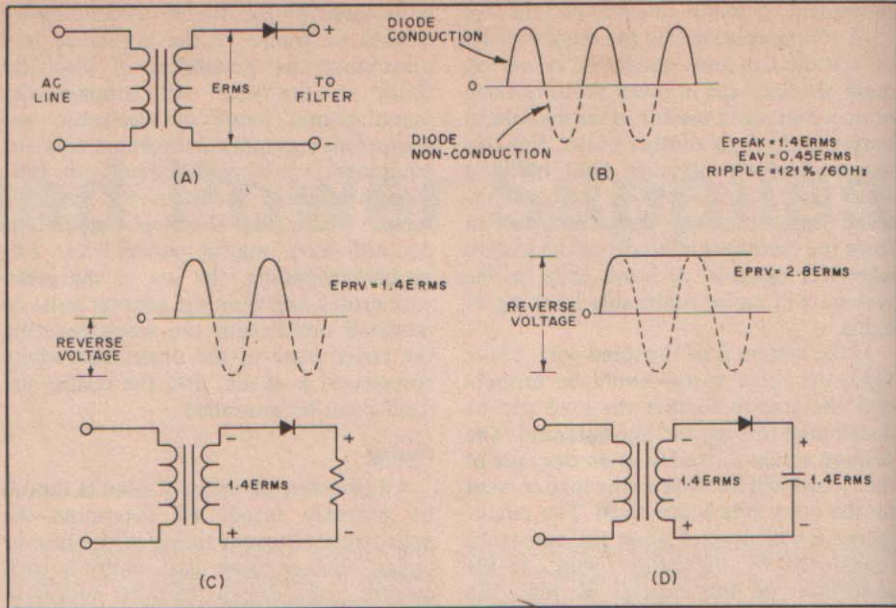


Fig. 4 — Half-wave rectifier circuit. A illustrates the basic circuit and B displays the diode conduction and nonconduction periods. The peak-reverse voltage impressed across the diode is shown at C and D with a simple resistor load at C and a capacitor load at D. E_{prv} for the resistor load is 1.4 E_{rms} and 2.8 E_{rms} for the capacitor load.

prominently in plain sight, and members of the household should be instructed in its location and use. I is a red lamp located alongside the switch. Its purpose is not so much to serve as a warning that the power is on as it is to help in identifying and quickly locating the switch should it become necessary for someone else to cut the power off in an emergency.

The outlet J should be placed in some corner out of sight where it will not be a temptation for children or others to play with. The shorting plug can be removed to open the power circuit if there are others around who might inadvertently throw the switch while the operator is working on the rig. If the operator takes the plug with him, it will prevent someone from turning on the power in his absence and either hurting himself or the equipment or perhaps starting a fire. Of utmost importance is the fact that the outlet J *must* be placed in the *ungrounded* side of the line.

Those who are operating low power and feel that the expense or complication of the switch isn't warranted can use the shorted-plug idea as the main power switch. In this case, the outlet should be located prominently and identified by a signal light, as shown in Fig. 3C.

The test bench should be fed through the main power switch, or a similar arrangement at the bench, if the bench is located remotely from the transmitter.

A bleeder resistor with a power rating which gives a considerable margin of safety should be used across the output of all transmitter power supplies, so that the filter capacitors will be discharged when the high-voltage is turned off.

Rectifier Circuits: Half-Wave

Fig. 4 shows a simple half-wave rectifier circuit. As pointed out in the semiconductor chapter a rectifier (in this case a semiconductor diode) will conduct current in one direction but not the other. During one half of the ac cycle the rectifier will conduct and current will flow through the rectifier to the load (indicated by the solid line in Fig. 4B). During the other half cycle the rectifier is reverse biased and no current will flow (indicated by the dotted line in Fig. 4B) to the load. As shown, the output is in the form of pulsed dc and current always flows in the same direction. A filter can be used to smooth out these variations and provide a higher average dc voltage from the circuit. This idea will be covered in the next section on filters.

The average output voltage — the voltage read by a dc voltmeter — with this circuit (no filter connected) is 0.45 times the rms value of the ac voltage delivered by the transformer secondary. Because the frequency of the pulses is rather low (one pulsation per cycle), considerable filtering is required to provide adequately smooth dc output. For this reason the circuit is usually limited to applications where the current required is small, as in a transmitter bias supply.

The peak reverse voltage (PRV), the voltage that the rectifier must withstand when it isn't conducting, varies with the load. With a resistive load it is the peak ac voltage (1.4 E_{rms}) but with a capacitor filter and a load drawing little or no current it can rise to 2.8 E_{rms}. The reason for this is shown in Figs. 4C and 4D. With a resistive load as shown at C the amount of reverse voltage applied to the diode is that voltage on the lower side of the Zero-axis line or 1.4 E_{rms}. A capacitor connected to the circuit (shown at D) will store the peak positive voltage when the diode conducts on the positive pulse. If the circuit is not supplying any current the voltage across the capacitor will remain at that same level. The peak reverse voltage impressed across the diode is now the sum of the voltage stored in the capacitor plus the peak negative swing of voltage from the transformer secondary. In this case the PRV is 2.8 E_{rms}.

Full-Wave Center-Tap Rectifier

A commonly used rectifier circuit is shown in Fig. 5. Essentially an arrangement in which the outputs of two half-wave rectifiers are combined, it makes use of both halves of the ac cycle. A transformer with a center-tapped secondary is required with the circuit.

The average output voltage is 0.9 times the rms voltage of half the transformer secondary; this is the maximum that can be obtained with a suitable choke-input filter. The peak output voltage is 1.4 times the rms voltage of half the transformer secondary; this is the maximum voltage that can be obtained from a capacitor-input filter.

As can be seen in Fig. 5C the PRV

impressed on each diode is independent of the type load at the output. This is because the peak reverse voltage condition occurs when diode A conducts and diode B does not conduct. The positive and negative voltage peaks occur at precisely the same time, a different condition than exists in the half-wave circuit. As diodes A and B cathodes reach a positive peak (1.4 E_{rms}), the anode of diode B is at a negative peak, also 1.4 E_{rms}, but in the opposite direction. The total peak reverse voltage is therefore 2.8 E_{rms}.

Fig. 5B shows that the frequency of the output pulses is twice that of the half-wave rectifier. Comparatively less filtering is required. Since the rectifiers work alternately, each handles half of the load current: The current rating of each rectifier need be only half the total current drawn from the supply.

Two separate transformers, with their primaries connected in parallel and secondaries connected in series (with the proper polarities), may be used in this circuit. However, if this substitution is made, the primary volt-ampere rating must be reduced to about 40 percent less than twice the rating of one transformer.

Full-Wave Bridge Rectifier

Another commonly used rectifier circuit is illustrated in Fig. 6. In this arrangement, two rectifiers operate in series on each half of the cycle, one rectifier being in the lead to the load, the other being in the return lead. As shown in Figs. 6A and B, when the top lead of the transformer secondary is positive with respect to the bottom lead diodes A and C will conduct while diodes B and D are reverse biased. On the next half cycle when the top lead of the transformer is negative with respect to the bottom diodes B and D will conduct while diodes A and C are reverse biased.

The output wave shape is the same as that from the simple center-tap rectifier circuit. The maximum output voltage into a resistive load or choke-input filter is 0.9 times the rms voltage delivered by the transformer secondary; with a capacitor filter and a light load the output voltage is 1.4 times the secondary rms voltage.

Fig. 6C shows the peak reverse voltage to be 2.8 E_{rms} for each pair of diodes. Since the diodes are connected in series each diode has 1.4 E_{rms} as the reverse voltage impressed across it. Each pair of diodes works alternately so each handles half of the load current. The rectifier in this circuit should have a minimum current rating of one half the total load current to be drawn from the supply.

Filtering

The pulsating dc waves from the rectifiers are not sufficiently constant in amplitude to prevent hum corresponding to the pulsations. Filters are required between the rectifier and the load to smooth out the pulsations into an essentially constant dc voltage. Also, the design of the filter depends to a large extent on the dc voltage output, the voltage regulation of the power supply, and the maximum load current that can be drawn from the supply without exceeding the peak-current rating of the rectifier. Power supply filters are low-pass devices using series inductors and shunt capacitors.

Load Resistance

In discussing the performance of power-supply filters, it is sometimes convenient to express the load connected to the output terminals of the supply in terms of resistance. The load resistance is equal to the output voltage divided by the total current drawn, including the current drawn by the bleeder resistor.

Voltage Regulation

The output voltage of a power supply always decreases as more current is drawn, not only because of increased voltage drops on the transformer, filter chokes and the rectifier (if high-vacuum rectifiers are used) but also because the output voltage at light loads tends to soar to the peak value of the transformer voltage as a result of charging the first capacitor. By proper filter design the latter effect can be eliminated. The change in output voltage with load is called *voltage regulation* and is expressed as a percentage.

$$\text{Percent regulation} = \frac{100\,(E_1 - E_2)}{E_2}$$

where

E_1 = the no-load voltage
E_2 = the full-load voltage

A steady load, such as that represented by a receiver, speech amplifier or unkeyed stages of a transmitter, does not require good (low) regulation as long as the proper voltage is obtained under load conditions. However, the filter capacitors

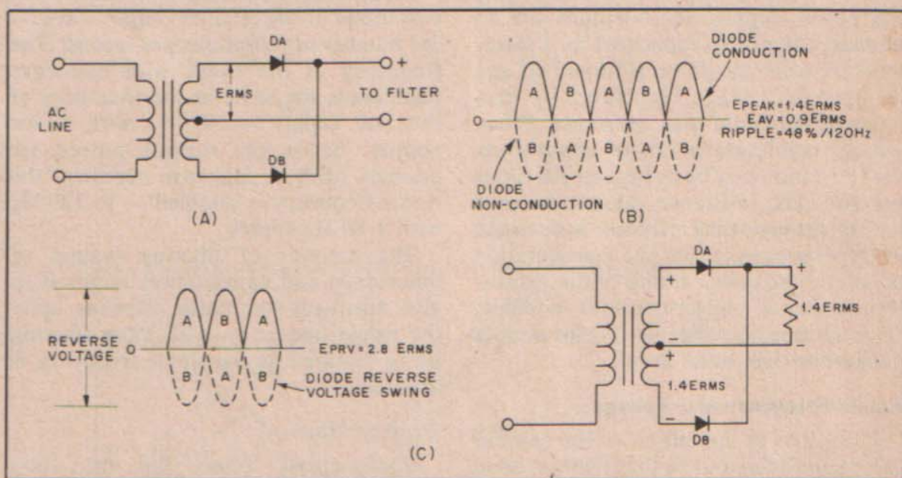


Fig. 5—Full-wave center-tap rectifier circuit. A illustrates the basic circuit. Diode conduction is shown at B with diodes A and B alternately conducting. The peak-reverse voltage for each diode is 2.8 E_{rms} as depicted at C.

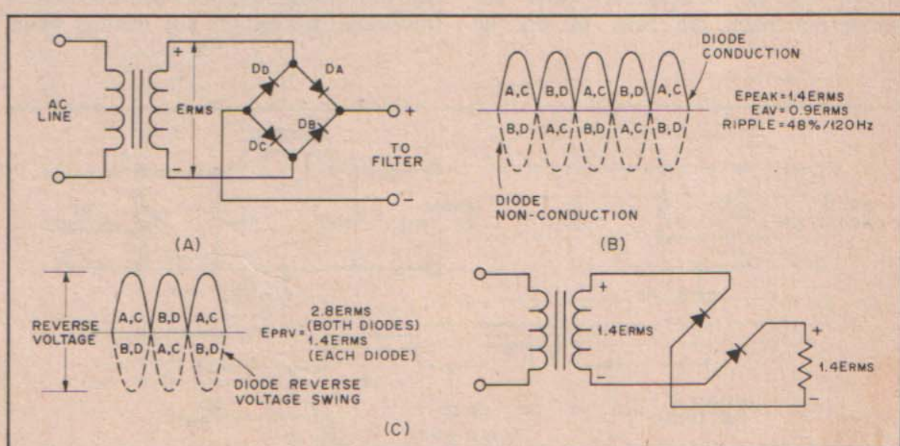


Fig. 6 — Full-wave bridge rectifier circuit. The basic circuit is illustrated at A. Diode conduction and nonconduction times are shown at B. Diodes A and C conduct on one half of the input cycle while diodes B and D conduct on the other. C displays the peak-reverse voltage for one-half cycle. Since this circuit uses two diodes essentially in series, the 2.8 E_{rms} is divided between two diodes, or, 1.4 E_{rms} PRV for each diode.

must have a voltage rating safe for the highest value to which the voltage will soar when the external load is removed.

A power supply will show more (higher) regulation with long-term changes in load resistance than with short temporary changes. The regulation with long-term changes is often called the *static regulation*, to distinguish it from the *dynamic regulation* (short temporary load changes). A load that varies at a syllabic or keyed rate, as represented by some audio and rf amplifiers, usually requires good dynamic regulation (15 percent or less) if distortion products are to be held to a low level. The dynamic regulation of a power supply is improved by increasing the value of the output capacitor.

When essentially constant voltage regardless of current variation is required (for stabilizing an oscillator, for example), special voltage-regulating circuits described later in this chapter are used.

Bleeder

A bleeder resistor is a resistance connected across the output terminals of the power supply. Its functions are to discharge the filter capacitors as a safety measure when the power is turned off and to improve voltage regulation by providing a minimum load resistance. When voltage regulation is not of importance, the resistance may be as high as 100 ohms per volt. The resistance value to be used for voltage-regulating purposes is discussed in later sections. From the consideration of safety, the power rating of the resistor should be as conservative as possible, since a burned-out bleeder resistor is more dangerous than none at all!

Ripple Frequency and Voltage

Pulsations at the output of the rectifier can be considered to be the resultant of an alternating current superimposed on a steady direct current. From this viewpoint, the filter may be considered to consist of shunt capacitors which short-circuit the ac component while not interfering with the flow of the dc component. Series chokes will readily pass dc but will impede the flow of the ac component.

The alternating component is called *ripple*. The effectiveness of the filter can be expressed in terms of percent ripple, which is the ratio of the rms value of the ripple to the dc value in terms of percentage.

$$\text{Percent ripple (rms)} = \frac{100\,E_1}{E_2}$$

where

E_1 = the rms value of ripple voltage
E_2 = the steady dc voltage

Any multiplier or amplifier supply in a code transmitter should have less than five percent ripple. A linear amplifier can tolerate about three percent ripple on the plate voltage. Bias supplies for linear amplifiers, and modulator and modulated-amplifier plate supplies, should have less than one percent ripple. VFOs, speech amplifiers and receivers may require a ripple reduction to 0.01 percent.

Ripple frequency is the frequency of the pulsations in the rectifier output wave — the number of pulsations per second. The frequency of the ripple with half-wave rectifiers is the same as the frequency of the line supply — 60 Hz with 60-Hz supply. Since the output pulses are doubled with a full-wave rectifier, the ripple frequency is doubled — to 120 Hz with a 60-Hz supply.

The amount of filtering (values of inductance and capacitance) required to give adequate smoothing depends upon the ripple frequency, with more filtering being required as the ripple frequency is lowered.

Type of Filter

Power-supply filters fall into two classifications, capacitor input and choke input. Capacitor-input filters are characterized by relatively high output voltage in respect to the transformer voltage. Advantage of this can be taken when silicon rectifiers are used or with any rectifier when the load resistance is high. Silicon rectifiers have a higher allowable peak-to-dc ratio than do thermionic rectifiers. This permits the use of capacitor-input filters at ratios of input capacitor to load resistance that would seriously shorten the life of a thermionic rectifier system. When the series resistance through a rectifier and filter system is appreciable, as when high-vacuum rectifiers are used, the voltage regulation of a capacitor-input power supply is poor.

The output voltage of a properly designed choke-input power supply is less than would be obtained with a capacitor-input filter from the same transformer. Generally speaking, a choke-input filter will permit a higher load current to be drawn from a thermionic rectifier without exceeding the peak rating of the rectifier.

Capacitive-Input Filters

Capacitive-input filter systems are shown in Fig. 7. Disregarding voltage drops in the chokes, all have the same characteristics except in respect to ripple. Better ripple reduction will be obtained when LC sections are added as shown in Figs. 7B and C.

Output Voltage

To determine the approximate dc voltage output when a capacitive-input filter is used, the graphs shown in Fig. 8 will be helpful. An example of how to use the graph is given below.

Example:
Full-wave rectifier (use graph at B)
Transformer rms voltage = 350
Load resistance = 2000 ohms
Series resistance = 200 ohms
Input capacitance = 20 μF

$$\frac{R}{RS} = \frac{200}{2000} = 0.1 \quad \frac{RC}{1000} = \frac{2000 \times 20}{1000} = 40$$

From curve 0.1 and RC = 40, the dc voltage is (350 × 1.06) = 370.

In many cases it is desirable to know the amount of capacitance required for a power supply given certain performance criteria. This is especially true when designing a power supply for an application such as powering a solid-state transceiver. The following example should give the builder a good handle on how to arrive at circuit values for a power supply using a single capacitor filter.

Fig. 9 is the circuit diagram of the power supply to be used.

Requirements:
Output voltage = 12.6
Output current = 1 ampere
Maximum ripple = 2 percent
Load regulation = 5 percent

The rms secondary voltage of T1 must be the desired output voltage plus the voltage drops across D2 and D4 divided by 1.41.

$$E_{SEC} = \frac{12.6 + 1.4}{1.41} = 9.93$$

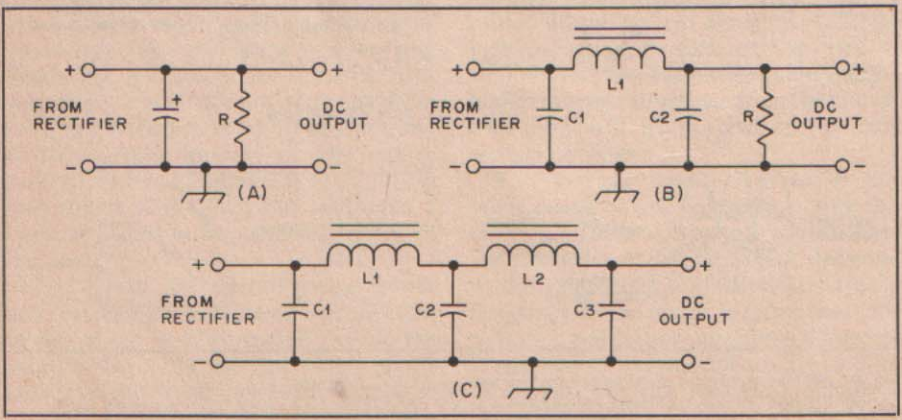


Fig. 7 — Capacitive-input filter circuits. At A is a simple capacitor filter. B and C are single- and double-section filters, respectively.

In practice the nearest standard transformer (10 V) would work fine. Alternatively, the builder could wind his own transformer, or remove secondary turns from a 12-volt transformer to obtain the desired rms secondary voltage.

A two percent ripple referenced to 12.6 volts is 0.25 V rms. The peak-to-peak value is therefore $0.25 \times 2.8 = 0.7$ V. This value is required to calculate the required capacitance for C1.

Also needed for determining the value of C1 is the time interval (t) between the full-wave rectifier pulses which is calculated as follows:

$$t = \frac{1}{f_{(Hz)}} = \frac{1}{120} = 8.3 \times 10^{-3}$$

where t is the time between pulses and f is the frequency in Hz. Since the circuit makes use of a full-wave rectifier a pulse occurs twice during each cycle. With half-wave rectification a pulse would occur only once a cycle. Thus 120 Hz is used as the frequency for this calculation.

C1 is calculated from the following equation:

$$C_{(\mu F)} = \left[\frac{I_L t}{E_{rip}(pk\text{-}pk)}\right] 10^6$$

$$= \left[\frac{1A \times 8.3 \times 10^{-3}}{0.7}\right] 10^6$$

$$= 11{,}857\ \mu F$$

where I_L is the current taken by the load. The nearest standard capacitor value is 12,000 μF. It will be an acceptable one to use, but since the tolerance of electrolytic capacitors is rather loose, the builder may elect to use the next larger standard value.

Diodes D1-D4, inclusive, should have a PRV rating of at least two times the transformer secondary peak voltage. Assuming a transformer secondary rms value of 10 volts, the PRV should be at least 28 volts. Four 50-volt diodes will provide a margin of safety. The forward current of the diodes should be at least twice the load current. For a 1-A load, the diodes should be rated for at least 2 A.

The load resistance, R_L, is determined by E_o/I_L, which in this example is $12.6/1 = 12.6$ ohms. This factor must be known in order to find the necessary series resistance for five-percent regulation. Calculate as follows:

$$R_{S\,(max)} = \text{Load regulation}\left(\frac{R_L}{10}\right)$$

$$= 0.05\left(\frac{12.6}{10}\right) = 0.063 \text{ ohm}$$

Therefore, the transformer secondary dc resistance should be no greater than 0.063

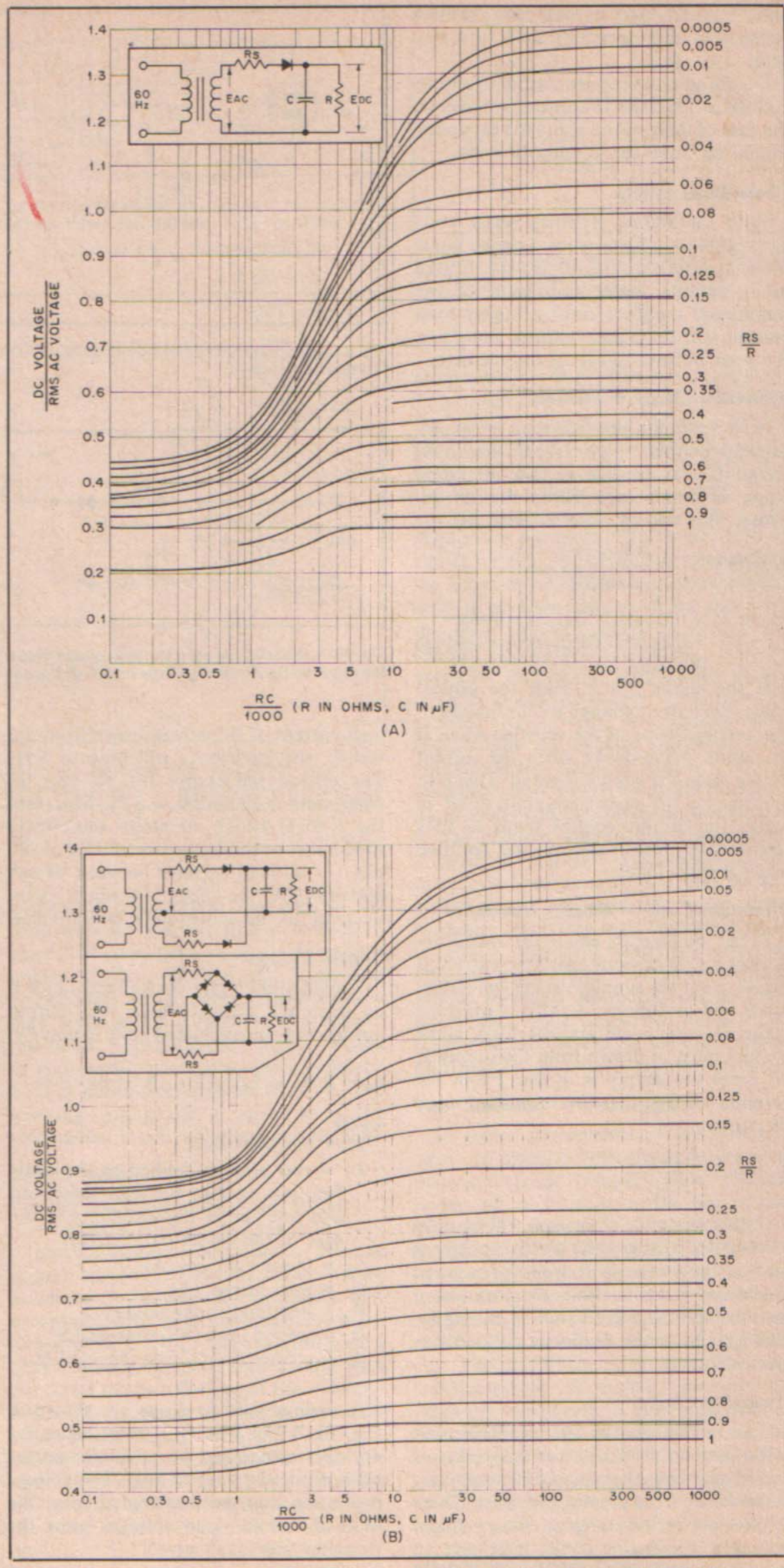


Fig. 8 — Dc output voltages from a half- and full-wave rectifier circuit as a function of the filter capacitance and load resistance (half-wave shown at A and full-wave shown at B). R_s includes transformer winding resistance and rectifier forward resistance. For the ratio R_s/R, both resistances are in ohms; for the RC product, R is in ohms and C is in μF.

ohm. The secondary current rating should be equal to or greater than the I_L = 1 ampere.

C1 should have a minimum working voltage of 1.4 times the output voltage. In the case of this power supply the capacitor should be rated for at least 18 volts.

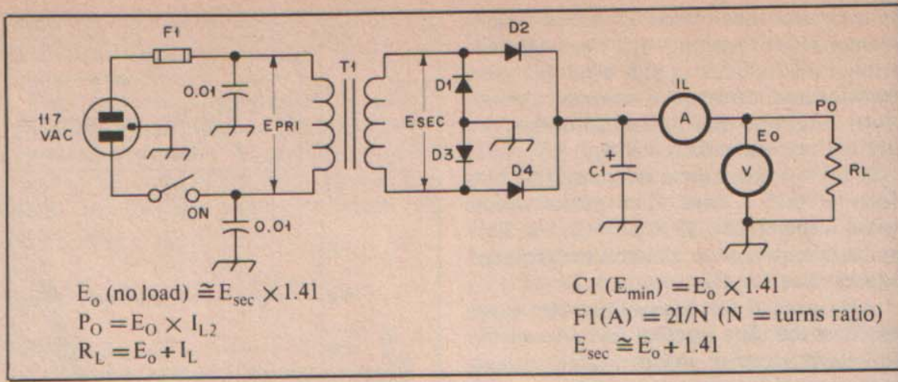


Fig. 9 — This figure illustrates how to design a simple unregulated power supply. See text for a thorough discussion.

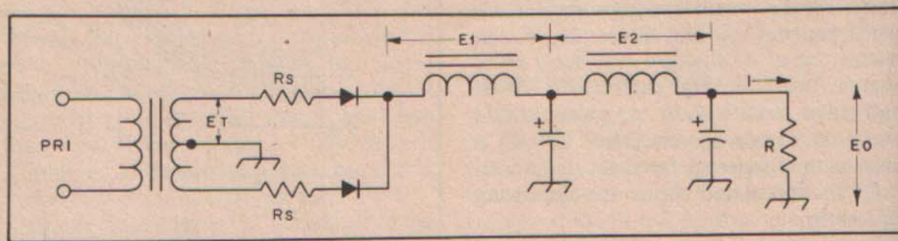


Fig. 10 — Diagram showing various voltage drops that must be taken into consideration in determining the required transformer voltage to deliver the desired output voltage.

Choke-Input Filters

With thermionic rectifiers better voltage regulation results when a choke-input filter, as shown in Fig. 10, is used. Choke input permits better utilization of the thermionic rectifier, since a higher load current can be drawn without exceeding the peak current rating of the rectifier.

Minimum Choke Inductance

A choke-input filter will tend to act as a capacitive-input filter unless the input choke has at least a certain minimum value of inductance called the *critical* value. This critical value is given by

$$L_{crit}\text{ (henrys)} = \frac{\text{E (volts)}}{\text{I (mA)}}$$

where E = the supply output voltage
I = the current being drawn through the filter.

If the choke has at least the critical value, the output voltage will be limited to the average value of the rectified wave at the input to the choke when the current drawn from the supply is small. This is in contrast to the capacitive-input filter in which the output voltage tends to soar toward the peak value of the rectified wave at light loads.

Minimum-Load — Bleeder Resistance

From the formula above for critical inductance, it is obvious that if no current is drawn from the supply, the critical inductance will be infinite. So that a practical value of inductance may be used, some current must be drawn from the supply at all times the supply is in use. From the formula we find that this minimum value of currents is

$$\text{I (mA)} = \frac{\text{E (volts)}}{L_{crit}}$$

In the majority of cases it will be most convenient to adjust the bleeder resistance so that the bleeder will draw the required minimum current. From the formula, it may be seen that the value of critical inductance becomes smaller as the load current increases.

Swinging Chokes

Less costly chokes are available that will maintain at least the critical value of inductance over the range of current likely to be drawn from practical supplies. These chokes are called *swinging chokes.* As an example, a swinging choke may have an inductance rating of 5/25 H and a current rating of 200 mA. If the supply delivers 1000 volts, the minimum load current should be 1000/25 = 40 mA. When the full load current of 200 mA is drawn from the supply, the inductance will drop to 5 H. The critical inductance for 200 mA at 1000 volts is 1000/200 = 5 H. Therefore the 5/25 H choke maintains the critical inductance at the full current rating of 200 mA. At all load currents between 40 mA and 200 mA, the choke will adjust its inductance to the approximate critical value.

Output Voltage

Provided the input-choke inductance is at least the critical value, the output voltage may be calculated quite closely by:

$$E_o = 0.9E_t - (I_B + I_L) \times (R1 + R2) - E_r$$

where
E_o = output voltage
E_t = rms voltage applied to the rectifier (rms voltage between center-tap and one end of the secondary in the case of the center-tap rectifier)
I_B = bleeder current (A)
I_L = load current (A)
R_1 = first filter choke resistance
R_2 = second filter choke resistance
E_r = voltage drop across the rectifier.

The various voltage drops are shown in Fig. 10. At no load I_L is zero; hence the no-load voltage may be calculated on the basis of bleeder current only. The voltage regulation may be determined from the no-load and full-load voltages using the formulas previously given.

Output Capacitor

Whether the supply has a choke- or capacitor-input filter, if it is intended for use with a Class A af amplifier, the reactance of the output capacitor should be low for the lowest audio frequency; 16 μF or more is usually adequate. When the supply is used with a Class B amplifier (for modulation or for ssb amplification) or a cw transmitter, increasing the output capacitance will result in improved dynamic regulation of the supply. However, a region of diminishing returns can be reached, and 20 to 30 μF will usually suffice for any supply subjected to large changes at a syllabic (or keying) rate.

Resonance

Resonance effects in the series circuit across the output of the rectifier, formed by the first choke and first filter capacitor, must be avoided, since the ripple voltage would build up to large values. This not only is the opposite action to that for which the filter is intended, but may also cause excessive rectifier peak currents and abnormally high peak-reverse voltages. For full-wave rectification the ripple frequency will be 120 Hz for a 60-Hz supply, and resonance will occur when the product of choke inductance in henrys times capacitor capacitance in microfarads is equal to 1.77. At least twice this product of inductance and capacitance should be used to ensure against resonance effects. With a swinging choke, the minimum rated inductance of the choke should be used.

Ratings of Filter Components

In a power supply using a choke-input filter and properly designed choke and bleeder resistor, the no-load voltage

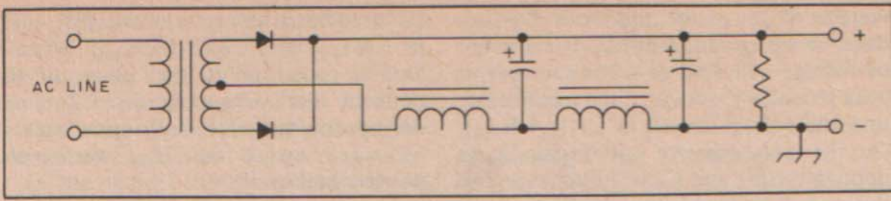


Fig. 11 — In most applications, the filter chokes may be placed in the negative instead of the positive side of the circuit. This reduces the danger of a voltage breakdown between the choke winding and core.

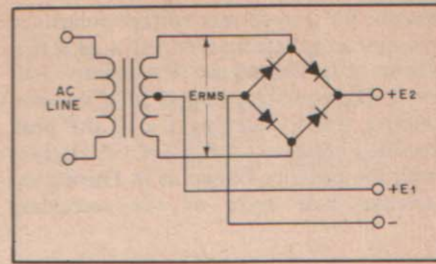


Fig. 12 — The "economy" power supply circuit is a combination of the full-wave and bridge-rectifier circuits.

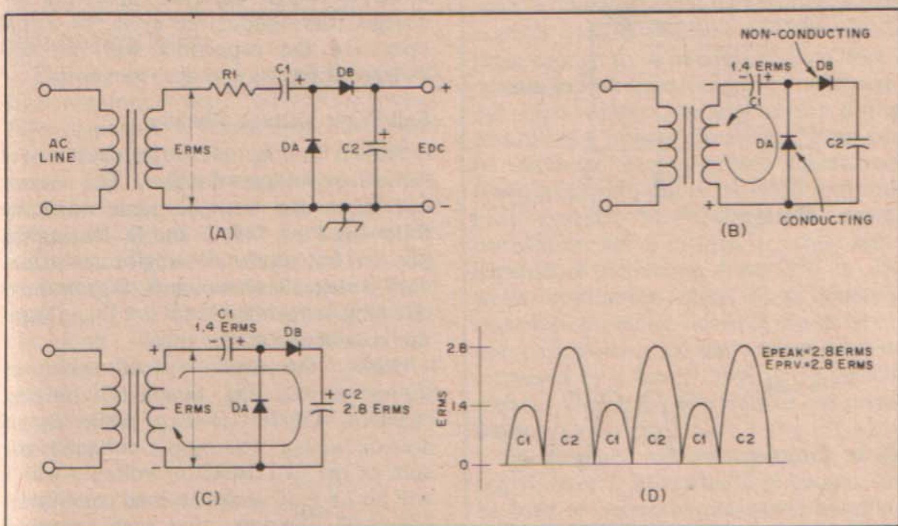


Fig. 13 — Illustrated at A is a half-wave voltage-doubler circuit. B displays how the first half cycle of input voltage charges C1. During the next half cycle (shown at C) capacitor C2 is charged with the transformer secondary voltage plus that voltage stored in C1 from the previous half cycle. D illustrates the levels to which each capacitor is charged throughout the cycle.

across the filter capacitors will be about nine-tenths of the ac rms voltage. Nevertheless, it is advisable to use capacitors rated for the *peak* transformer voltage. This large safety factor is suggested because the voltage across the capacitors can reach this peak value if the bleeder should burn out and there is no load on the supply.

In a capactive-input filter, the capacitors should have a working-voltage rating at least as high, and preferably somewhat higher, than the peak voltage from the transformer. Thus, in the case of a center-tap rectifier having a transformer delivering 550 volts each side of the center tap, the minimum safe capacitor voltage rating will be 550 × 1.41 or 775 volts. An 800-volt capacitor should be used, or preferably a 1000-volt unit.

Filter Capacitors in Series

Filter capacitors are made in several different types. Electrolytic capacitors, which are available for peak voltages up to about 800, combine high capacitance with small size, since the dielectric is an extremely thin film of oxide on aluminum foil. Capacitors of this type may be connected in series for higher voltages, although the filtering capacitance will be reduced to the resultant of the two capacitances in series. If this arrangement is used, it is important that *each* of the capacitors be shunted with a resistor of about 100 ohms per volt of supply voltage applied to the individual capacitors, with an adequate power rating. These resistors may serve as all or part of the bleeder resistance. Capacitors with higher voltage ratings usually are made with a dielectric of thin paper impregnated with oil. The working voltage of a capacitor is the voltage that it will withstand continuously.

Filter Chokes

Filter chokes or inductances are wound on iron cores, with a small gap in the core to prevent magnetic saturation of the iron at high currents. When the iron becomes saturated its permeability decreases, and consequently the inductance also decreases. Despite the air gap, the inductance of a choke usually varies to some extent with the direct current flowing in the winding; hence it is necessary to specify the inductance at the current which the choke is intended to carry. Its inductance with little or no direct current flowing in the winding will usually be considerably higher than the value when full load current is flowing.

Negative-Lead Filtering

For many years it has been almost universal practice to place filter chokes in the positive leads of plate power supplies. This means that the insulation between the choke winding and its core (which should be grounded to chassis as a safety measure) must be adequate to withstand the output voltage of the supply. This voltage requirement is removed if the chokes are placed in the negative lead as shown in Fig. 11. With this connection, the capacitance of the transformer secondary to ground appears in parallel with the filter chokes tending to bypass the chokes. However, this effect will be negligible in practical application except in cases where the output ripple must be reduced to a very low figure. Such applications are usually limited to low-voltage devices such as receivers, speech amplifiers and VFOs where insulation is no problem and the chokes may be placed in the positive side in the conventional manner. In higher-voltage applications, there is no reason why the filter chokes should not be placed in the negative lead to reduce insulation requirements. Choke terminals, negative capacitor terminals and the transformer center-tap terminal should be well protected against accidental contact, since these will assume full supply voltage to chassis should a choke burn out or the chassis connection fail.

The "Economy" Power Supply

In many transmitters of the 100-watt class, an excellent method for obtaining plate and screen voltages without wasting power in resistors is by the use of the "economy" power-supply circuit. Shown in Fig. 12, it is a combination of the full-wave and bridge-rectifier circuits. The voltage at E1 is the normal voltage obtained with the full-wave circuit, and the voltage at E2 is that obtained with the bridge circuit. The *total* dc power obtained from the transformer is, of course, the same as when the transformer is used in its normal manner. In cw and ssb applications, additional power can usually be drawn without excessive heating, especially if the transformer has a rectifier filament winding that isn't being used.

Half-Wave Voltage Doubler

Fig. 13 shows the circuit of half-wave voltage doubler. Figs. 13B, C and D illustrate the circuit operation. For clarity, assume the transformer voltage polarity at the moment the circuit is activated is that shown at B. During the first negative half cycle D_A conducts (D_B is in a nonconductive state), charging C1 to the peak rectified voltage (1.4 E_{rms}). C1 is charged with the polarity shown at B. During the positive half cycle of the secondary voltage, D_A is cutoff and diode D_B conducts charging capacitor C2. The amount of voltage delivered to C2 is the sum of peak secondary voltage of the transformer plus the voltage stored in C1 (1.4 E_{rms}). On the next negative half cycle, D_B is nonconducting and C2 will discharge into the load. If no load is connected across C2 the capacitors will remain charged — C1 to 1.4 E_{rms} and C2 to 2.8 E_{rms}. When a load is connected to the output of the doubler, the voltage across C2 drops during the negative half cycle and is recharged up to 2.8 E_{rms} during the positive half cycle.

The output waveform across C2 resembles that of a half-wave rectifier circuit in that C2 is pulsed once every cycle. The drawing at Fig. 13D illustrates the levels to which the two capacitors are charged throughout the cycle. In actual operation the capacitors will not discharge all the way to zero as shown.

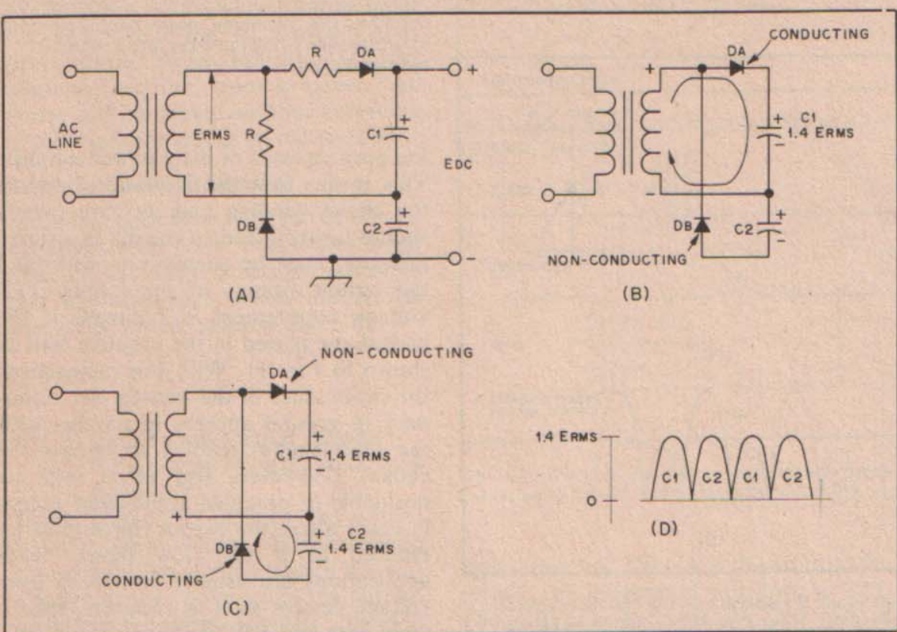


Fig. 14 — A full-wave voltage doubler is displayed at A. One half cycle is shown at B and the next half cycle at C. Each capacitor receives a charge during every cycle of input voltage. D illustrates how each capacitor is alternately charged.

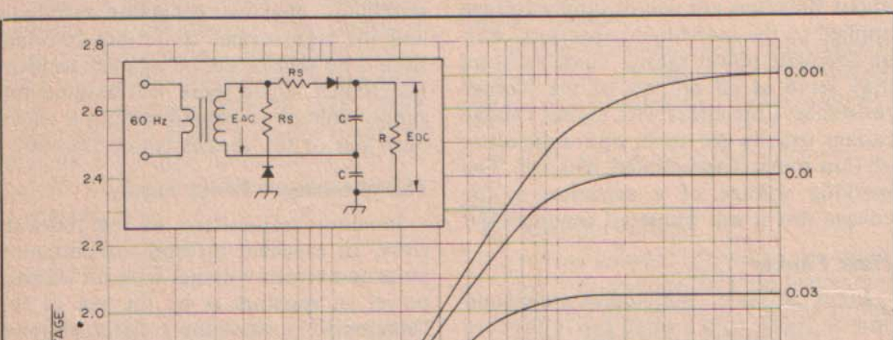


Fig. 15 — Dc output voltages from a full-wave voltage-doubling circuit as a function of the filter capacitances and load resistance. For the ratio R_s/R and for the RC product, resistances are in ohms and capacitance is in microfarads. Equal resistance values for R_s and equal capacitance values for C are assumed.

Full-Wave Voltage Doubler

Shown in Fig. 14 is the circuit of a full-wave voltage doubler. The circuit operation can best be understood by following Figs. 14B, C and D. During the positive half cycle of transformer secondary voltage, as shown at B, D_A conducts charging capacitor C1 to 1.4 E_{rms}. D_B is not conducting at this time.

During the negative half cycle, as shown at C, D_B conducts charging capacitor C2 to 1.4 E_{rms} while D_A is nonconducting. The output voltage is the sum of the two capacitor voltages which will be 2.8 E_{rms} under no-load conditions. Fig. 14D illustrates that each capacitor alternately receives a charge once per cycle. The effective filter capacitance is that of C1 and C2 in series, which is less than the capacitance of either C1 or C2 alone.

Resistors R in Fig. 14A are used to limit the surge current through the rectifiers. Their values are based on the transformer voltage and the rectifier surge-current rating, since at the instant the power supply is turned on the filter capacitors look like a short-circuited load. Provided the limiting resistors can withstand the surge current, their current-handling capacity is based on the maximum load current from the supply. Output voltages approaching twice the peak voltage of the transformer can be obtained with the voltage doubling circuit shown in Fig. 14. Fig. 15 shows how the voltage depends upon the ratio of the series resistance to the load resistance, and the load resistance times the filter capacitance. The peak reverse voltage across each diode is 2.8 E_{rms}.

Voltage Tripling and Quadrupling

A voltage-tripling circuit is shown in Fig. 16A. On one half of the ac cycle C1 and C3 are charged to the source voltage through D1, D2 and D3. On the opposite half of the cycle D2 conducts and C2 is charged to twice the source voltage, because it sees the transformer plus the charge in C1 as its source. (D1 is cut off during this half cycle.) At the same time, D3 conducts, and with the transformer and the charge in C2 as the source, C3 is charged to three times the transformer voltage.

The voltage-quadrupling circuit of

Fig. 16B works in substantially similar fashion. In either of the circuits of Fig. 16, the output voltage will approach an exact multiple of the peak ac voltage when the output current drain is low and the capacitance values are high.

In the circuits shown, the negative leg of the supply is common to one side of the transformer. The positive leg can be made common to one side of the transformer by reversing the diodes and capacitors.

Plate and Filament Transformers: Volt-Ampere Rating

The number of volt-amperes delivered by a transformer depends upon the type of filter (capacitor or choke input) used, and upon the type of rectifier used (full-wave center tap, or full-wave bridge). With a capacitive-input filter the heating effect in the secondary is higher because of the high ratio of peak-to-average current. The volt-amperes handled by the transformer may be several times the watts delivered to the load. With a choke-input filter, provided the input choke has at least the critical inductance, the secondary volt-amperes can be calculated quite closely by the equation:

$$\text{(Full-wave ct) Sec VA} = \frac{0.707\,E\,I}{1000}$$

$$\text{(Full-wave bridge) Sec VA} = \frac{E\,I}{1000}$$

where

E = *total* rms voltage of the secondary (between the outside ends in the case of a center-tapped winding)

I = dc output current in milliamperes (load current plus bleeder current)

The primary volt-amperes will be somewhat higher because of transformer losses.

Broadcast and Television Replacement Transformers

Small power transformers of the type sold for replacement in broadcast and television receivers are usually designed for service in terms of use for several hours continuously with capacitor-input filters. In the usual type of amateur transmitter service, where most of the power is drawn intermittently for periods of several minutes with equivalent intervals in between, the published ratings can be exceeded without excessive transformer heating.

With a capacitor-input filter, it should be safe to draw 20 to 30 percent more current than the rated value. With a choke-input filter, an increase in current of about 50 percent is permissible. If a bridge rectifier is used, the output voltage will be approximately doubled. In this case, it should be possible in amateur transmitter service to draw the rated current, thus obtaining about twice the rated output power from the transformer.

This does not apply, of course, to amateur transmitter plate transformers, which usually are rated for intermittent service.

Rewinding Power Transformers

Although the home winding of power transformers is a task that few amateurs undertake, the rewinding of a transformer secondary to give some desired voltage for powering filaments or a solid-state device is not difficult. It involves a matter of only a small number of turns and the wire is large enough to be handled easily. Often a receiver power transformer with a burned-out high-voltage winding or the power transformer from a discarded TV set can be converted into an entirely satisfactory transformer without great effort and with little expense. The average TV power transformer for a 17-inch or larger set is capable of delivering from 350 to 450 watts, continuous duty. If an amateur transmitter is being powered, the service is not continuous, so the ratings can be increased by a factor of 40 or 50 percent without danger of overloading the transformer.

The primary volt-ampere rating of the transformer to be rewound, if known, can be used to determine its power-handling capability. The secondary volt-ampere rating will be 10 to 20 percent less than the primary rating. The power rating may also be determined approximately from the cross-sectional area of the core which is *inside* the windings. Fig. 17 shows the method of determining the area, and Fig. 18 may be used to convert this information into a power rating.

Before disconnecting the winding leads from their terminals, each should be marked for identification. In removing the core laminations, care should be taken to note the manner in which the core is assembled, so that the reassembling will be done in the same manner. Most transformers have secondaries wound over the primary, while in some the order is reversed. In case the secondaries are on the inside, the turns can be pulled out from the center after slitting and removing the fiber core.

The turns removed from one of the original filament windings of known voltage should be carefully counted as the winding is removed. This will give the number of turns per volt and the same figure should be used in determining the number of turns for the new secondary. For instance, if the old filament winding was rated at 5 volts and had 15 turns, this is 15/5 = 3 turns per volt. If the new secondary is to deliver 18 volts, the required number of turns on the new winding will be 18 × 3 = 54 turns.

In winding a transformer, the size of wire is an important factor in the heat developed in operation. A cross-sectional area of 1000 circular mils per ampere is conservative. A value commonly used in amateur-service transformers is 700 cmil/A. The larger the cmil/A figure, the cooler the transformer will run. The current rating in amperes of various wire sizes is shown in the copper-wire table in another chapter. If the transformer being rewound is a filament transformer, it may be necessary to choose the wire size carefully to fit the small available space. On the other hand, if the transformer is a power unit with the high-voltage winding removed, there should be plenty of room for a size of wire that will conservatively

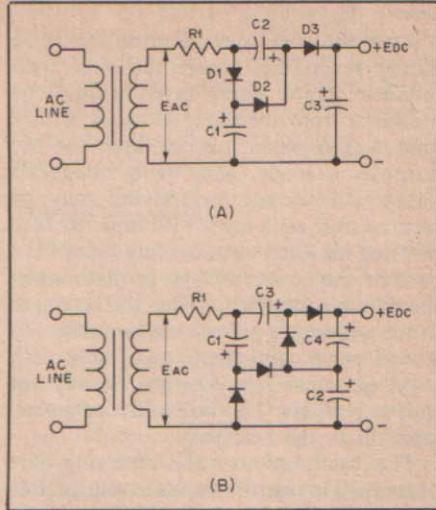


Fig. 16 — Voltage-multiplying circuits with one side of transformer secondary common. (A) Voltage tripler; (B) voltage quadrupler.
Capacitances are typically 20 to 50 µF depending upon output current demand. Dc ratings of capacitors are related to E_{peak} (1.4 E_{ac}).
C1 — Greater than E_{peak}
C2 — Greater than $2E_{peak}$
C3 — Greater than $3E_{peak}$
C4 — Greater than $4E_{peak}$

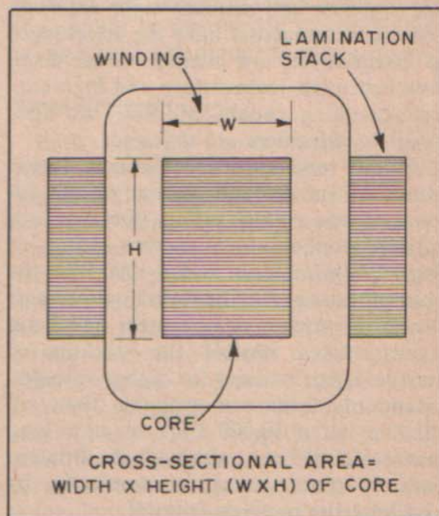


Fig. 17 — Cross-sectional drawing of a typical power transformer. Multiplying the height (or thickness of the laminations) by the width of the central core area in inches gives the value to be applied to Fig. 18.

handle the required current.

After the first layer of turns is put on during rewinding, secure the ends with cellulose tape. Each layer should be insulated from the next; ordinary household waxed paper can be used for the purpose, *a single layer being adequate.* Sheets cut to size beforehand may be secured over each layer with tape. Be sure to bring all leads out the same side of the core so the covers will go in place when the unit is completed. When the last layer of the winding is put on, use two sheets of waxed paper, and then cover those with vinyl electrical tape, keeping the tape as taut as possible. This will add mechanical strength to the assembly.

The laminations and housing are assembled in just the opposite sequence to that followed in disassembly. Use a light coating of shellac between each lamination. During reassembly, the lamination stack may be compressed by clamping in a vise. If the last few lamination strips cannot be replaced, it is better to omit them than to force the unit together.

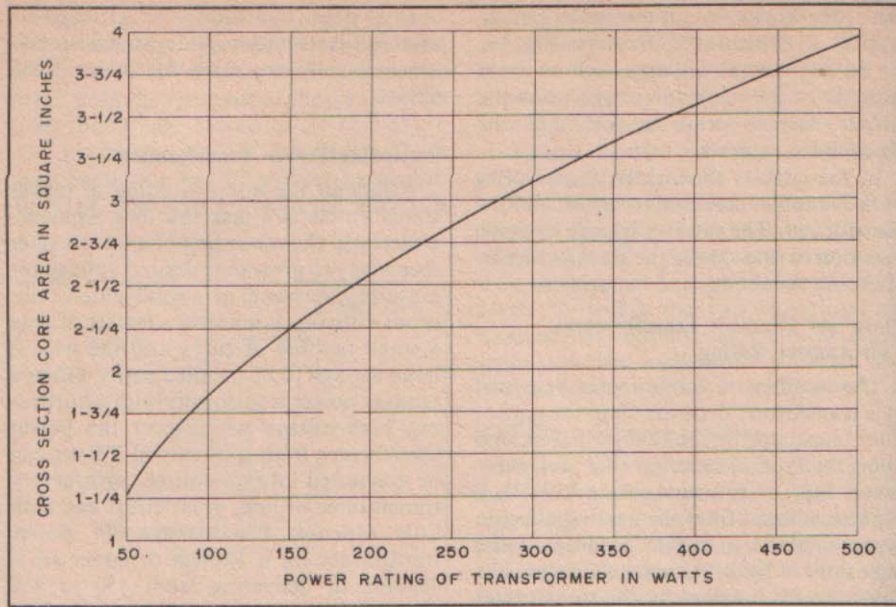


Fig. 18 — Power-handling capability of a transformer versus cross-sectional area of core.

Rectifier Ratings: Semiconductors

Silicon rectifiers are being used almost exclusively in power supplies for amateur equipment. Types are available to replace high-vacuum and mercury-vapor rectifiers. The semiconductors have the advantages of compactness, low internal voltage drop, low operating temperature and high current-handling capability. Also, no filament transformers are required.

Silicon rectifiers are available in a wide range of voltage and current ratings. In peak reverse voltage ratings of 600 or less, silicon rectifiers carry current ratings as high as 400 amperes, and at 1000 PRV the current ratings may be several amperes or so. The extreme compactness of silicon types makes feasible the stacking of several units in series for higher voltages. Standard stacks are available that will handle up to 10,000 PRV at a dc load current of 500 mA, although the amateur can do much better, economically, by stacking the rectifiers himself.

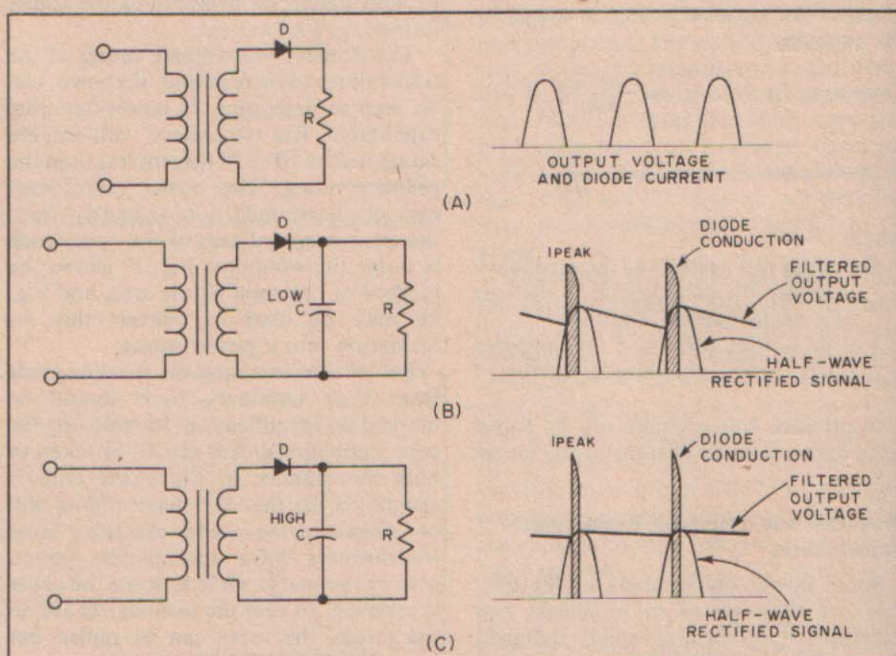


Fig. 19 — The circuit shown at A is a simple half-wave rectifier with a resistive load. The waveform shown to the right is that of output voltage and diode current. B illustrates how the diode current is modified by the addition of a capacitor filter. The diode conducts only when the rectified voltage is greater than stored capacitor voltage. Since this time period is usually only a short portion of a cycle, the peak current will be quite high. C shows an even higher peak current. This is due to the larger capacitor which effectively shortens the conduction period of the diode.

Protection of Silicon Power Diodes

The important specifications of a silicon diode are

1) PRV (or PIV), the peak reverse (or peak inverse) voltage.

2) I_o, the average dc current rating.

3) I_{REP}, the peak repetitive forward current.

4) I_{SURGE}, the peak one-cycle surge current. The first two specifications appear in most catalogs. The last two often do not, but they are very important.

Since the rectifier never allows current to flow more than half the time, when it does conduct it has to pass at least twice the average direct current. With a capacitor-input filter, the rectifier conducts much less than half the time, so that when it does conduct, it may pass as much as 10 to 20 times the average dc current, under certain conditions. This is shown in Fig. 19. At A is a simple half-wave rectifier with a resistive load. The waveform to the right of the drawing shows the output voltage along with the diode current. At B and C there are two periods of operation to consider. After the capacitor is charged to the peak-rectified voltage a period of diode nonconduction elapses while the output voltage discharges through the load. As the voltage begins to rise on the next positive pulse a point is reached where the rectified voltage equals the stored voltage in the capacitor. As the voltage rises beyond that point the diode begins to supply current. The diode will continue to conduct until the waveform reaches the crest, as shown. Since the diode must pass a current equal to that of the load over a short period of a cycle the current will be high. The larger the capacitor for a given load,

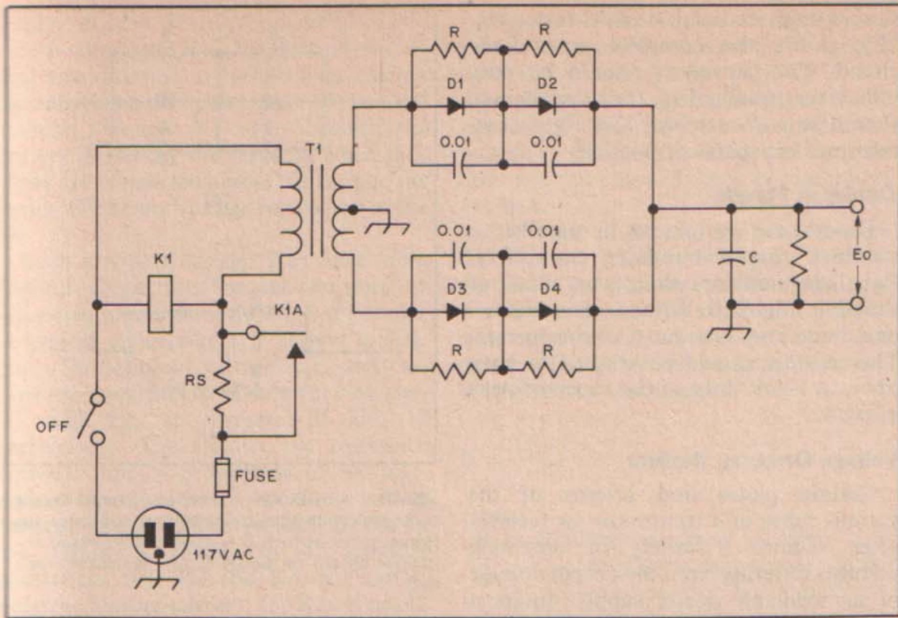


Fig. 20 — The primary circuit of T1 shows how a 117-volt ac relay and a series dropping resistor, R_s, can provide surge protection while C charges. When silicon rectifiers are connected in series for high-voltage operation, the inverse voltage does not divide equally. The reverse voltage drops can be equalized by using equalizing resistors, as shown in the secondary circuit. To protect against voltage "spikes" that may damage an individual rectifier, each rectifier should be bypassed by a 0.01-μF capacitor. Connected as shown two 400-PRV silicon rectifiers can be used as an 800-PRV rectifier, although it is preferable to include a safety factor and call it a "750-PRV" rectifier. The rectifiers, D1 through D4, should be the same type (same type number and ratings).

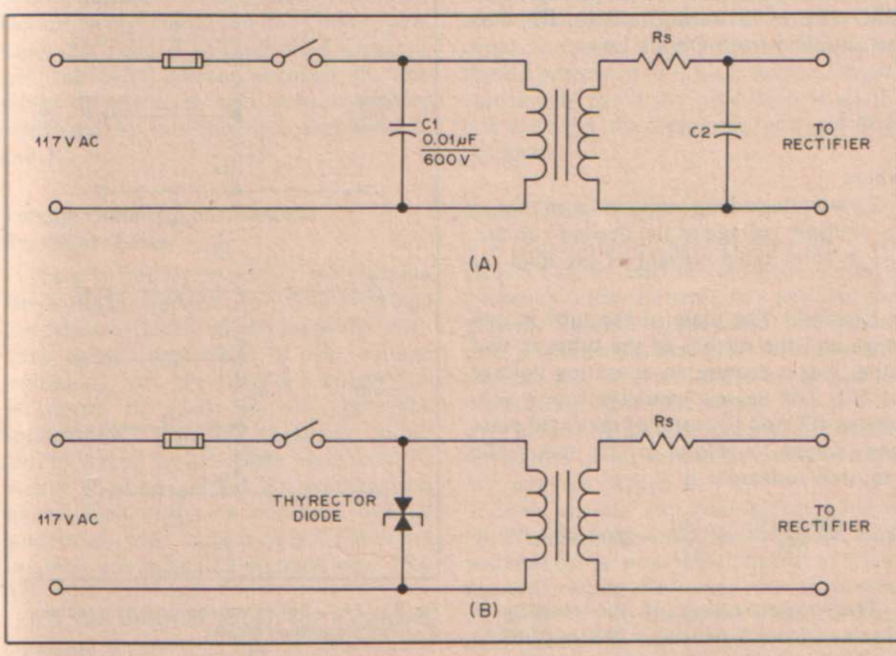


Fig. 21 — Methods of suppressing line transients. See text.

the shorter the diode conduction time and the higher the peak repetitive current (I_{REP}).

When the supply is first turned on, the discharged input capacitor looks like a dead short, and the rectifier passes a very heavy current. This is I_{SURGE}. The maximum I_{SURGE} rating is usually for a duration of one cycle (at 60 Hz), or about 16.7 milliseconds.

If a manufacturer's data sheet is not available, an educated guess about a diode's capability can be made by using these rules of thumb for silicon diodes of the type commonly used in amateur power supplies:

Rule 1) The maximum I_{REP} rating can be assumed to be approximately four times the maximum I_o rating.

Rule 2) The maximum I_{SURGE} rating can be assumed to be approximately 12 times the maximum I_o rating. (This should provide a reasonable safety factor. Silicon rectifiers with 750-mA dc ratings, as an example, seldom have 1-cycle surge ratings of less than 15 amperes; some are rated up to 35 amperes or more.) From this then, it can be seen that the rectifier should be selected on the basis of I_{SURGE} and not on I_o ratings.

Thermal Protection

The junction of a diode is quite small, hence it must operate at a high current density. The heat-handling capability is, therefore, quite small. Normally, this is not a prime consideration in high-voltage, low-current supplies. When using high-current rectifiers at or near their maximum ratings (usually 2-ampere or larger stud-mount rectifiers), some form of heat sinking is necessary. Frequently, mounting the rectifier on the main chassis — directly, or by means of thin mica insulating washers — will suffice. If insulated from the chassis, a thin layer of silicone grease should be used between the diode and the insulator, and between the insulator and the chassis to assure good heat conduction. Large high-current rectifiers often require special heat sinks to maintain a safe operating temperature. Forced-air cooling is sometimes used as a further aid. Safe case temperatures are usually given in the manufacturer's data sheets and should be observed if the maximum capabilities of the diode are to be realized.

Surge Protection

Each time the power supply is activated, assuming the input filter capacitor has been discharged, the rectifiers must look into what represents a dead short. Some form of surge protection is usually necessary to protect the diodes until the input capacitor becomes nearly charged. Although the dc resistance of the transformer secondary can be relied upon in some instances to provide ample surge-current limiting, it is seldom enough on high-voltage power supplies to be suitable. Series resistors can be installed between the secondary and the rectifier strings, but are a deterrent to good voltage regulation. By installing a surge-limiting device in the primary circuit of the plate transformer, the need for series resistors in the secondary circuit can be avoided. A practical method for primary-circuit surge control is shown in Fig. 20. The resistor, R_s introduces a voltage drop in the primary feed to T1 until C is nearly charged. Then, after C becomes partially charged, the voltage drop across R_s lessens and allows K1 to pull in, thus applying full primary power to T1 as K1A shorts out R_s. R_s is usually a 25-watt resistor whose resistance is somewhere between 15 and 50 ohms, depending upon the power supply characteristics.

Transient Problems

A common cause of trouble is transient voltages on the ac power line. These are short spikes, mostly, that can temporarily increase the voltage seen by the rectifier to values much higher than the normal transformer voltage. They come from distant lightning strokes, electric motors turning on and off, and so on. Transients cause unexpected, and often unexplained, loss of silicon rectifiers.

It's always wise to suppress line transients, and it can be easily done. Fig. 21 A shows one way. C1 looks like 280,000 ohms at 60 Hz, but to a sharp transient (which has only high-frequency components), it is an effective bypass. C2 provides additional protection on the secondary side of the transformer. It should be 0.01 μF for transformer voltages of 100 or less, and 0.001 μF for high-voltage transformers.

Fig. 21B shows another transient-suppression method using selenium suppressor diodes. The diodes do not conduct unless the peak voltage becomes abnormally high. Then they clip the transient peaks. General Electric sells protective diodes under the trade name, "Thyrector." Sarkes-Tarzian uses the descriptive name, "Klipvolt."

Transient voltages can go as high as twice the normal line voltage before the suppressor diodes clip the peaks. Capacitors cannot give perfect suppression either. Thus, it is a good idea to use power-supply rectifiers rated at about twice the expected PRV.

Diodes in Series

Where the PRV rating of a single diode is not sufficient for the application, similar diodes may be used in series. (Two 500-PRV diodes in series will withstand 1000 PRV, and so on.) When this is done, a resistor and a capacitor should be placed across each diode in the string to equalize the PRV drops and to guard against transient voltage spikes, as shown in Fig. 22A. Even though the diodes are of the same type and have the same PRV rating, they may have widely different back resistances when they are cut off. The reverse voltage divides according to Ohm's Law, and the diode with the higher back resistance will have the higher voltage developed across it. The diode may break down.

If we put a swamping resistor across each diode, R as shown in Fig. 22A, the resultant resistance across each diode will be almost the same, and the back voltage will divide almost equally. A good rule of thumb for resistor size is this: Multiply the PRV rating of the diode by 500 ohms. For example, a 500-PRV diode should be shunted by 500 × 500, or 250,000 ohms.

The shift from forward conduction to high back resistance does not take place instantly in a silicon diode. Some diodes take longer than others to develop high back resistance. To protect the "fast" diodes in a series string until all the diodes are properly cut off, a 0.01-μF capacitor should be placed across each diode. Fig. 22A shows the complete series-diode circuit. The capacitors should be non-inductive, ceramic disk, for example, and should be well matched. Use 10-percent-tolerance capacitors if possible.

Diodes in Parallel

Diodes can be placed in parallel to increase current-handling capability. Equalizing resistors should be added as shown in Fig. 22B. Without the resistors, one diode may take most of the current. The resistors should be selected to have about a 1-volt drop at the expected peak current.

Voltage Dropping Resistor

Certain plates and screens of the various tubes in a transmitter or receiver often require a variety of operating voltages differing from the output voltage of an available power supply. In most cases, it is not economically feasible to provide a separate power supply for each of the required voltages. If the current drawn by an electrode (or combination of electrodes operating at the same voltage) is reasonably constant under normal operating conditions, the required voltage may be obtained from a supply of higher voltage by means of a voltage-dropping resistor in series, as shown in Fig. 23A. The value of the series, resistor, R1, may be obtained from Ohm's Law,

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

where
E_d = voltage drop required from the supply voltage to the desired voltage.
I = total rated current of the load

Example: The plate of the tube in one stage and the screens of the tubes in two other stages require an operating voltage of 250. The nearest available supply voltage is 400 and the total of the rated plate and screen currents is 75 mA. The required resistance is

$$R = \frac{400 - 250}{.075} = \frac{150}{.075} = 2000 \text{ ohms}$$

The power rating of the resistor is obtained from P (watts) = $I^2R = (0.075)^2 \times (2000) = 11.2$ watts. A 20-watt resistor is the nearest safe rating to be used.

Voltage Dividers

The regulation of the voltage obtained in this manner obviously is poor, since any change in current through the resistor will cause a directly proportional change in the voltage drop across the resistor. The regulation can be improved somewhat by connecting a second resistor from the low-voltage end of the first to the negative power-supply terminal, as shown in Fig. 23B. Such an arrangement constitutes a

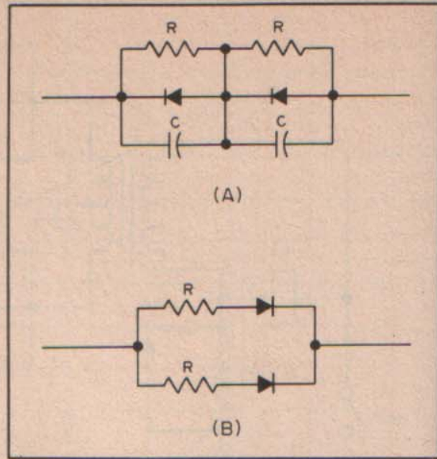


Fig. 22 — A — Diodes connected in series should be shunted with equalizing resistors and spike-suppressing capacitors. B — Diodes connected in parallel should be series current equalizing resistors.

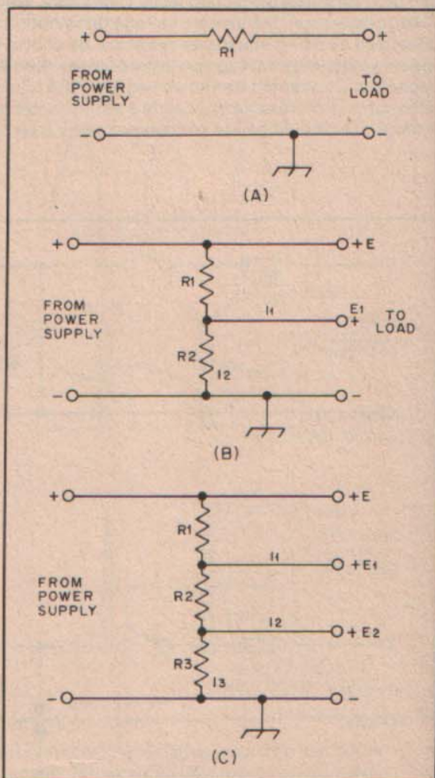


Fig. 23 — A — Series voltage-dropping resistor. B — Simple voltage divider.

$$R2 = \frac{E1}{I2};\quad R1 = \frac{E - E1}{I1 + I2}$$

I2 must be assumed.

C — Multiple divider circuit.

$$R3 = \frac{E2}{I3};\quad R2 = \frac{E1 - E2}{I1 + I3}$$

$$R1 = \frac{E - E1}{I1 + I2 + I3}$$

I3 must be assumed.

voltage divider. The second resistor, R2, acts as a constant load for the first, R1, so that any variation in current from the tap becomes a smaller percentage of the total current through R1. The heavier the current drawn by the resistors when they alone are connected across the supply, the better will be the voltage regulation at the tap.

Such a voltage divider may have more than a single tap for the purpose of obtaining more than one value of voltage. A typical arrangement is shown in Fig. 23C. The terminal voltage is E, and two taps are provided to give lower voltages, E1 and E2, at currents I1 and I2 respectively. The smaller the resistance between taps in proportion to the total resistance, the lower is the voltage between the taps. The voltage divider in the figure is made up of separate resistances, R1, R2 and R3. R3 carries only the bleeder current, I3; R2 carries I2 in addition to I3; R1 carries I1, I2 and I3. To calculate the resistances required, a bleeder current, I3, must be assumed; generally it is low compared with the total load current (10 percent or so). Then the required values can be calculated as shown in the caption of Fig. 23, I being in decimal parts of an ampere.

The method may be extended to any desired number of taps, each resistance section being calculated by Ohm's Law using the needed voltage drop across it and the total current through it. The power dissipated by each section may be calculated by multiplying I and E or I^2 and R.

Voltage Stabilization: Gaseous Regulator Tubes

There is frequent need for maintaining the voltage applied to a low-voltage low-current circuit at a practically constant value, regardless of the voltage regulation of the power supply or variations in load current. In such applications, gaseous regulator tubes (0B2/VR105, 0A2/VR150, etc.) can be used to good advantage. The voltage drop across such tubes is constant over a moderately wide current range. Tubes are available for regulated voltages near 150, 105, 90 and 75 volts.

The fundamental circuit for a gaseous regulator is shown in Fig. 24. The tube is connected in series with a *limiting resistor*, R1, across a source of voltage that must be higher than the starting voltage. The starting voltage is about 30 to 40 percent higher than the operating voltage. The load is connected in parallel with the tube. For stable operation, a minimum tube current of 5 to 10 mA is required. The maximum permissible current with most types is 40 mA; consequently, the load current cannot exceed 30 to 35 mA if the voltage is to be stabilized over a range from zero to maximum load. A single VR tube may also be used to regulate the voltage to a load current of almost any value as long as the *variation* in the current does not exceed 30 to 35 mA. If, for example, the average load current is 100 mA, a VR tube may be used to hold the voltage constant provided the current does not fall below 85 mA or rise above 115 mA.

The value of the limiting resistor must lie between that which just permits minimum tube current to flow and that which just passes the maximum permissible tube current when there is no load current. The latter value is generally used. It is given by the equation:

$$R = \frac{(E_s - E_r)}{I}$$

where

R = limiting resistance in ohms

E_s = voltage of the source across which the tube and resistor are connected.

E_r = rated voltage drop across the regulator tube.

I = maximum tube current in amperes (usually 40 mA, or 0.04 A)

Two tubes may be used in series to give a higher regulated voltage than is obtainable with one, and also to give two values of regulated voltage. Regulation of the order of one percent can be obtained with these regulator tubes when they are operated within their proper current range. The capacitance in shunt with a VR tube should be limited to 0.1 μF or less. Larger values may cause the tube drop to oscillate between the operating and starting voltages.

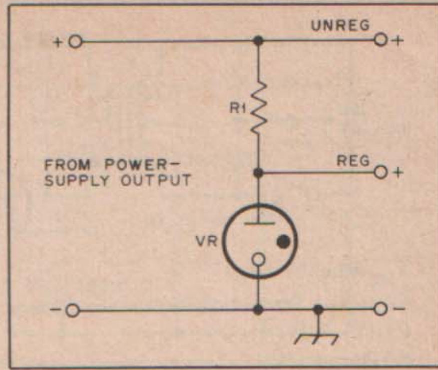


Fig. 24 — Voltage stabilization circuit using a VR tube. A negative-supply output may be regulated by reversing the polarity of the power-supply connections and the VR-tube connections from those shown here.

Zener Diode Regulation

A Zener diode (named after Dr. Carl Zener) can be used to stabilize a voltage source in much the same way as when the gaseous regulator tube is used. The typical circuit is shown in Fig. 25A. Note that the cathode side of the diode is connected to the positive side of the supply. The electrical characteristics of a Zener diode under conditions of forward and reverse voltage are given in chapter 4.

Zener diodes are available in a wide variety of voltages and power ratings. The voltages range from less than two to a few hundred, while the power ratings (power the diode can dissipate) run from less than 0.25 watt to 50 watts. The ability of the Zener diode to stabilize a voltage is dependent upon the conducting impedance of the diode, which can be as low as one ohm or less in a low-voltage, high-power diode to as high as a thousand ohms in a low-power, high-voltage diode.

Diode Power Dissipation

Unlike gaseous regulator tubes, Zener diodes of a particular voltage rating have varied maximum current capabilities, depending upon the power ratings of each of the diodes. The power dissipated in a diode is the product of the voltage across

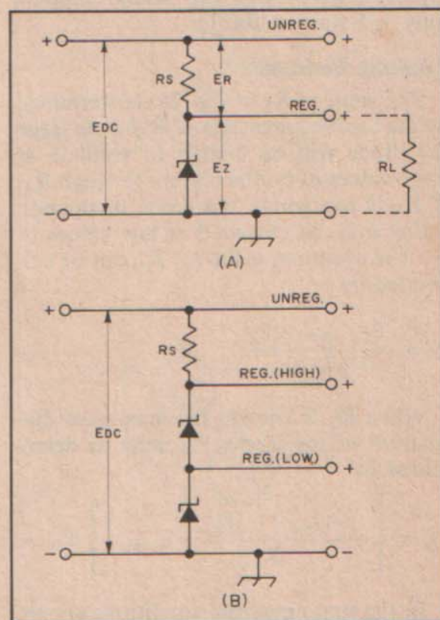
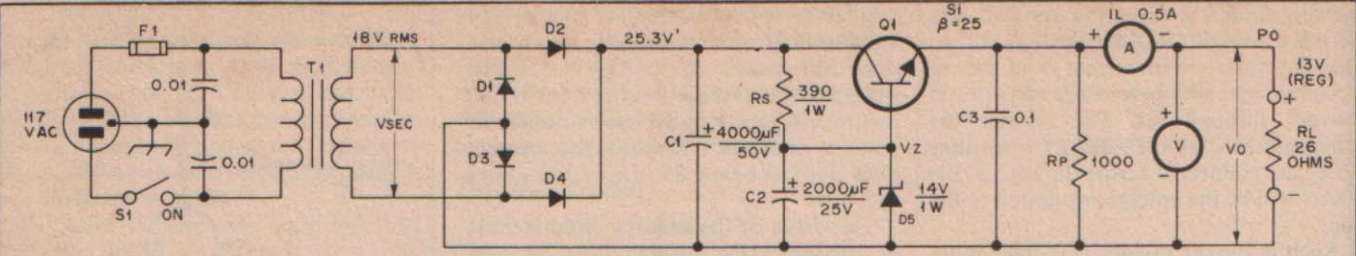


Fig. 25 — Zener-diode voltage regulation. The voltage from a negative supply may be regulated by reversing the power-supply connections and the diode polarities.

it and the current through it. Conversely, the maximum current a particular diode may safely conduct equals its power rating divided by its voltage rating. Thus, a 10-V, 50-W Zener diode, if operated at its maximum dissipation rating, would conduct 5 amperes of current. A 10-V 1-W diode, on the other hand, could safely conduct no more than 0.1A, or 100 mA. The conducting impedance of a diode is its voltage rating divided by the current flowing through it, and in the above examples would be 2 ohms for the 50-W diode, and 100 ohms for the 1-W diode. Disregarding small voltage changes which may occur, the conducting impedance of a given diode is a function of the current flowing through it, varying in inverse proportion.

The power-handling capability of most Zener diodes is rated at 25°C, or approximately room temperature. If the diode is



V_{sec} (rms) $\cong 1.4\ V_o$
C1 (μF) – See section on capacitive filters
C1 (V) $\cong 2V'$
C2 (V_{min}) $> V_Z$

C2 (μF) $\cong 0.5$ C1 (μF)
$R_p \cong V_o \times 80$
$VR1 = V_o + 0.7$
$V_o \cong V_o - V_z - 0.7$

$V' = V_{sec}(\text{rms}) \times 1.41$
$P_o = V_o \times I_L$
$R_L = V_o + I_L$
$F1 = I_L \times 2$

Fig. 26 — Illustration of a power supply with regulation. A pass transistor, Q1, is used to extend the range of the Zener-diode regulator.

operated in a higher ambient temperature, its power capability must be derated. A typical 1-watt diode can safely dissipate only 1/2 watt at 100°C.

Limiting Resistance

The value of R_S in Fig. 25 is determined by the load requirements. If R_S is too large the diode will be unable to regulate at large values of I_L, the current through R_L. If R_S is too small, the diode dissipation rating may be exceeded at low values of I_L. The optimum value for R_S can be calculated by:

$$R_S = \frac{E_{DC\,(min)} - E_Z}{1.1\ I_{L\,(max)}}$$

When R_S is known, the maximum dissipation of the diode, P_D, may be determined by

$$P_D = \left[\frac{E_{DC(max)} - E_Z}{R_S} - I_{L\,(min)}\right] E_Z$$

In the first equation, conditions are set up for the Zener diode to draw 1/10 the maximum load current. This assures diode regulation under maximum load.

Example: A 12-volt source is to supply a circuit requiring 9 volts. The load current varies between 200 and 350 mA.

$E_Z = 9.1$ V (nearest available value)

$$R_S = \frac{12 - 9.1}{1.1 \times 0.35} = \frac{2.9}{0.385} = 7.5 \text{ ohms}$$

$$P_D = \left[\frac{12 - 9.1}{7.5} - 0.2\right] 9.1$$

$$= 0.185 \times 9.1 = 1.7 \text{ W}$$

The nearest available dissipation rating above 1.7 W is 5; therefore, a 9.1-V 5-W Zener diode should be used. Such a rating, it may be noted, will cause the diode to be in the safe dissipation range even though the load is completely disconnected [I_L (min) = 0].

Obtaining Other Voltages

Fig. 25B shows how two Zener diodes may be used in series to obtain regulated voltages not normally obtainable from a single Zener diode, and also to give two values of regulated voltage. The diodes need not have equal breakdown voltages, because the arrangement is self equalizing. However, the current-handling capability of each diode should be taken into account. The limiting resistor may be calculated as above, taking the sum of the diode voltages as E_Z, and the sum of the load currents as I_L.

Electronic Voltage Regulation

Several circuits have been developed for regulating the voltage output of a power supply electronically. While more complicated than the VR-tube and Zener-diode circuits, they will handle higher voltage and current variations, and the output voltage may be varied continuously over a wide range.

Voltage regulators fall into two basic types. In the type most commonly used by amateurs, the dc supply delivers a voltage higher than that which is available at the output of the regulator, and the regulated voltage is obtained by dropping the voltage down to a lower value through a dropping "resistor." Regulation is accomplished by varying either the current through a fixed dropping resistance as changes in input voltage or load currents occur (as in the VR-tube and Zener-diode regulator circuits), or by varying the equivalent resistive value of the dropping element with such changes. This latter technique is used in electronic regulators where the voltage-dropping element is a vacuum tube or a transistor, rather than an actual resistor. By varying the dc voltage at the grid or current at the base of these elements, the conductivity of the device may be varied as necessary to hold the output voltage constant. In solid-state regulators the series-dropping element is called a pass transistor. Power transistors are available which will handle several amperes of current at several hundred volts, but solid-state regulators of this type are usually operated at potentials below 100 volts.

The second type of regulator is a switching type, where the voltage from the dc source is rapidly switched on and off (electronically). The average dc voltage available from the regulator is proportional to the duty cycle of the switching wave form, or the ratio of the on time to the total period of the switching cycle. Switching frequencies of several kilohertz are normally used to avoid the need for extensive filtering to smooth the switching frequency from the dc output.

The above information pertains essentially to voltage regulators. A circuit can also be constructed to provide current regulation. Such regulation is usually obtained in the form of current limitation — to a maximum value which is either preset or adjustable, depending on the circuit. Relatively simple circuits, such as described later, can be used to provide current limiting only. Current limiting circuitry may also be used in conjunction with voltage regulators.

Discrete Component Regulators

The previous section outlines some of the limitations when using Zener diodes as regulators. Greater current amounts can be accommodated if the Zener diode is used as a reference at low current, permitting the bulk of the load current to flow through a series pass transistor (Q1 of Fig. 26). An added benefit in using a pass transistor is that of reduced ripple on the output waveform. This technique is commonly referred to as "electronic filtering."

Q1 of Fig. 26 can be thought of as a simple emitter-follower dc amplifier. It increases the load resistance seen by the Zener diode by a factor of beta (β). In this circuit arrangement D5 is required to supply only the base current for Q1. The net result is that the load regulation and ripple characteristics are improved by a factor of beta. Addition of C2 reduces the ripple even more, although many simple supplies such as this do not make use of a capacitor in that part of the circuit.

The primary limitation of this circuit is that Q1 can be destroyed almost immediately if a severe overload occurs at R_L. The fuse cannot blow fast enough to protect Q1. In order to protect Q1 in case of an accidental short at the output, a current limiting circuit is required. An example of a suitable circuit is shown in Fig. 27.

It should be mentioned that the greater the value of transformer secondary voltage, the higher the power dissipation in Q1. This not only reduces the overall efficiency of the power supply, but requires stringent heat sinking at Q1.

Design Example

Example: Design a regulated, well-filtered, 13-volt dc supply capable of delivering 0.5 A, using the circuit of Fig. 26. Calculate the ratings for all components. A standard 18-volt secondary transformer is to be used.

Information on calculating the transformer, diode and input capacitor ratings were given earlier in this chapter and will not be repeated here. In order to calculate the value or R_S in Fig. 26 the base current of Q1 must be known. The base current is approximately equal to the emitter current of Q1 in amperes divided by beta. The transistor beta can be found in the manufacturer's data sheet, or measured with simple test equipment (beta = I_c/I_b). Since the beta spread for a particular type of transistor — 2N3055 for example, where it is specified as 25 to 70 — is a fairly unknown quantity, more precise calculations for Fig. 26 will result if the transistor beta is tested before the calculations are done. A conservative approach is to design for beta minimum of the transistor used. Calculating I_b:

$$I_b = \frac{0.5}{25} = 0.02\text{A} = 20\text{ mA}$$

As pointed out earlier, in order for D5 to regulate properly it is necessary that a fair portion of the current flowing through R_S should be drawn by D5. The resistor will have 0.02 A flowing through it as calculated above (base current of Q1). A conservative amount of 10 mA will be used for the Zener diode current bringing the total current through R_S to 0.03 A or 30 mA. From this, the value of R_S can be calculated as follows:

$$R_S = \frac{(V'-V_Z)}{I_{R_S}} = \frac{(25.3-14)}{0.03} = 376\text{ ohms}$$

The nearest standard ohmic value for R_S is 390. The wattage ratings for R_S and D5 can be obtained with the aid of the formulas given earlier for Zener-diode regulators.

The power rating for Q1 will be calculated next. The power dissipation of Q1 is equal to the emitter current times the collector-to-emitter voltage. Calculate as follows:

$$P_{Q1} = I_E \times V_{CE}$$

where

V_{CE} = the desired $V' - (V_Z - V_{BE})$, and V_{BE} is approximately 0.7 V for a silicon transistor.

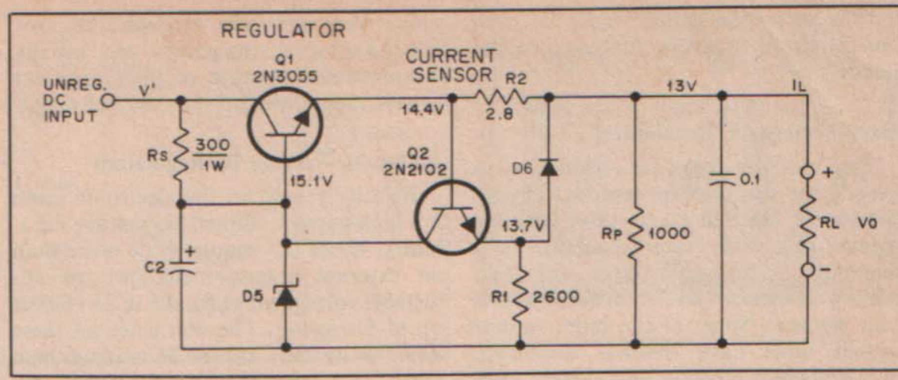


Fig. 27 — Overload protection for a regulated supply can be effected by addition of a current-overload protective circuit.

Therefore:

$$P_{Q1} = 0.5\text{A} \times 12\text{V} = 6\text{ watts}$$

It is a good idea to choose a transistor for Q1 that has at least twice the rating calculated. In this example a transistor with a power dissipation rating 12 watts or more would be used.

The 0.01-μF capacitors at the primary of T1 serve two functions. They act as transient suppressors and help prevent rf energy from entering the power-supply regulator.

Current Limiting for Discrete-Component Regulators

Damage to Q1 of Fig. 26 can occur when the load current exceeds the safe amount. Fig. 27 illustrates a simple current-limiter circuit that will protect Q1. All of the load current is routed through R2. A voltage difference will exist across R2, the amount being dependent upon the exact load current at a given time. When the load current exceeds a predetermined safe value, the voltage drop across R2 will forward bias Q2 and cause it to conduct. Since D6 is a silicon diode, and because Q2 is a silicon transistor, the combined voltage drops through them (roughly 0.7 V each) will be 1.4 V. Therefore the voltage drop across R2 must exceed 1.4 V before Q2 can turn on. This being the case, R2 is chosen for a value that provides a drop of 1.4 V when the maximum safe load current is drawn. In this instance 1.4 volts will be seen when I_L reaches 0.5A.

When Q2 turns on, some of the current through R_S flows through Q2, thereby depriving Q1 of some of its base current. This action, depending upon the amount of Q1 base current at a precise moment, cuts off Q1 conduction to some degree, thus limiting the flow of current through it.

High-Current-Output Regulators

When a single pass transistor is not available to handle the current which may be required from a regulator, the current-handling capability may be increased by connecting two or more pass transistors in parallel. The circuits at B and C of Fig. 28 show the method of connection. The

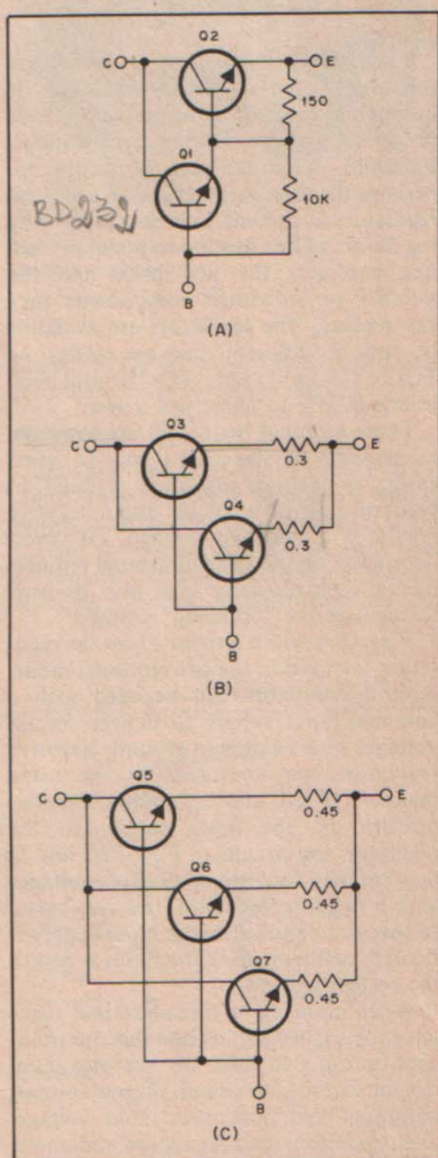


Fig. 28 — At A, a Darlington-connected pair for use as the pass element in a series-regulating circuit. At B and C, the method of connecting two or more transistors in parallel for high current output. Resistances are in ohms. The circuit at A may be used for load currents from 100 mA to 5 A, at B for currents from 6 to 10 A, and at C for currents from 9 to 15 A.
Q1 — Motorola MJE 340 or equivalent.
Q2-Q7, incl. — Power transistor such as 2N3055 or 2N3772.

resistances in the emitter leads of each transistor are necessary to equalize the currents.

Fixed-Voltage IC Regulators

The modern trend in regulators is toward the use of three-terminal devices commonly referred to as three-terminal regulators. Inside each regulator is a reference, a high-gain error amplifier, sensing resistors and transistors, and a pass element. Some of the more sophisticated units have thermal shutdown, over-voltage protection and current foldback. Many of the regulators currently on the market are virtually destruction-proof. Several supplies using these ICs are featured in the construction section of this chapter.

Three-terminal regulators (a connection for unregulated dc input, regulated dc output and ground) are available in a wide range of voltage and current ratings. Fairchild, National and Motorola are perhaps the three largest suppliers of these regulators at present. It is easy to see why regulators of this sort are so popular when one considers the low price and the number of individual components they can replace. The regulators are available in several different package styles — TO-3, TO-39, TO-66, TO-220 and dual in-line (DIP), to name just a few.

Three-terminal regulators are available as positive or negative types. In most cases, a positive regulator is used to regulate a positive voltage and a negative regulator a negative voltage. However, depending on the systems ground requirements, each regulator type may be used to regulate the "opposite" voltage.

Figs. 29A and B illustrate how the regulators are used in the conventional mode. Several regulators can be used with a common-input supply to deliver several voltages with a common ground. Negative regulators may be used in the same manner. If no other common supplies operate off the input supply to the regulator, the circuits of Figs. 29C and D may be used to regulate positive voltages with a negative regulator and vice versa. In these configurations the input supply is floated; neither side of the input is tied to the system ground.

When choosing a three-terminal regulator for a given application the important specifications to look for are maximum output current, maximum output voltage, minimum and maximum input voltage, line regulation, load regulation and power dissipation.

In use, these regulators require an adequate heat sink since they may be called on to dissipate a fair amount of power. Also, since the chip contains a high-gain error amplifier, bypassing of the input and output leads is essential to stable operation (See Fig. 30). Most manufacturers recommend bypassing the input and output directly at the leads where they protrude through the heat sink. Tantalum capacitors are usually recommended because of their excellent bypass capabilities up into the vhf range.

Adjustable-Voltage IC Regulators

Relatively new on the electronic scene are high-current, adjustable voltage regulators. These ICs require little more than an external potentiometer for an adjustable voltage range from 5 to 24 volts at up to 5 amperes. The unit price on these items is currently around $6 making them ideal for a test bench power supply. An adjustable-voltage power supply using the Fairchild 78HG series of regulator is described in the construction section of this chapter. The same precautions should be taken with these types of regulators as with the fixed-voltage units. Proper heat sinking and lead bypassing is essential for proper circuit operation.

A 12-Volt 3-Ampere Power Supply

Shown in Fig. 31 is a no-frills 12-volt supply capable of continuous operation at the 3-ampere level. Many low-power hf transceivers and most vhf-fm transceivers require voltages and currents on this order. Power supplies of this type purchased from the manufacturers can be quite costly. Described here is a very simple to build and relatively inexpensive (around $20 using all new components) alternative.

The schematic diagram for the power supply is shown in Fig. 32. As can be seen, the circuit is simplicity itself. A transformer, two diodes, three capacitors and a regulator form the heart of the supply. Binding posts, a pilot light, fuse and on-off switch complete the design.

Ac from the mains is supplied to the transformer-primary winding through the fuse in one leg, and the on-off switch in the other. The secondary circuit feeds a full-wave rectifier circuit which is filtered by C1. This unregulated voltage is routed to the input terminal of the regulator IC which is bypassed directly at the case with a 2-μF tantalum capacitor. The case of the IC is connected to ground. A 2-μF tantalum capacitor is also used at the output terminal of the regulator to prevent unwanted oscillation of the error amplifier inside the IC. A pilot light attached to the regulated output indicates when the supply is in use.

The regulator has built-in thermal shut down and over-current protection. Short circuiting the output of the supply will cause no damage. A wide margin of conservative component rating was used in the design of this supply. It should be possible to run the supply for hours on end at its maximum rating.

Construction

Rather than using an expensive cabinet, the power supply is housed on an aluminum chassis measuring 5 × 9-1/2 × 3

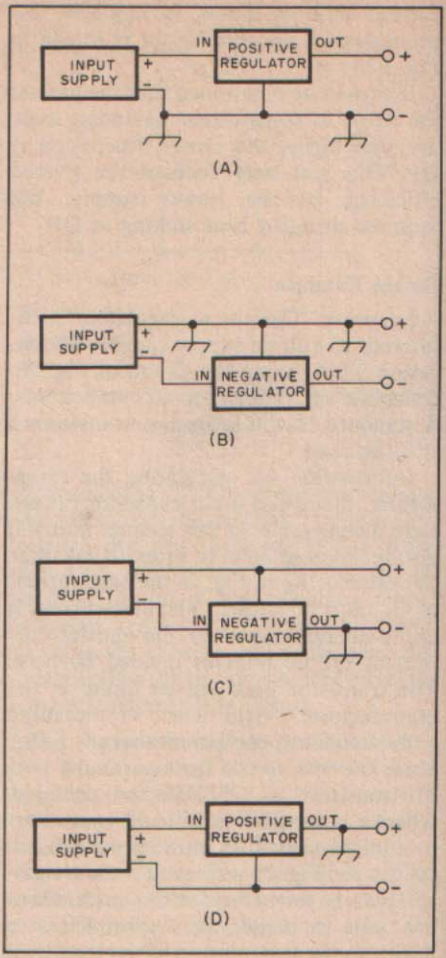


Fig. 29 — A and B illustrate the conventional manner in which three-terminal regulators are used. C and D show how one polarity regulator can be used to regulate the opposite polarity voltage.

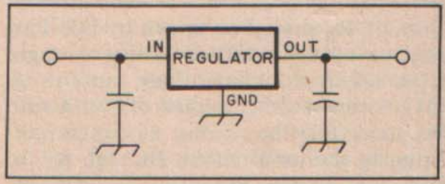


Fig. 30 — Three-terminal regulators require careful bypassing directly at the case. Here, both the input and output leads are bypassed.

Fig. 31 — Exterior view of the 12-volt, 3-ampere, no-frills power supply.

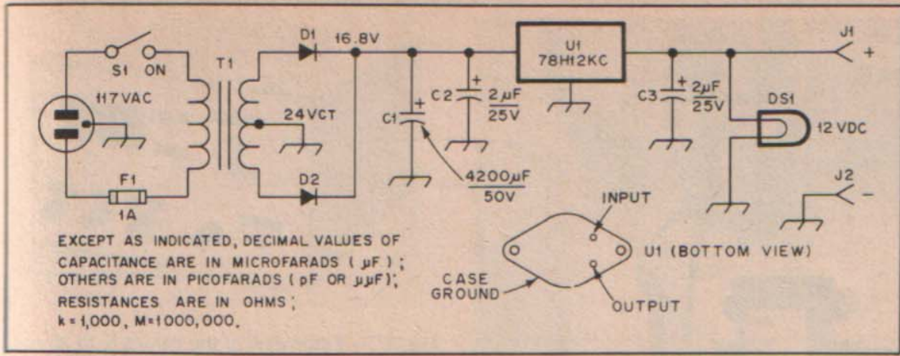


Fig. 32 — Schematic diagram of the 12-volt, 3- ampere power supply.

C1 — Electrolytic capacitor, 4200 µF, 50 V, General Electric 86F166M or equiv.
C2, C3 — Tantalum capacitor, 2 µF, 50 V.
D1, D2 — Silicon diode, 50 V, 6 A, HEP RO100 or equiv.
DS1 — Pilot light assembly, 12 V.
F1 — Fuse, 1 A.
J1, J2 — Binding post.
S1 — Spst toggle.
T1 — Power transformer; primary 117 V, secondary 24 V ct, Stancor P-8663 or equiv.
U1 — Voltage regulator, Fairchild 78H12KC or equiv.

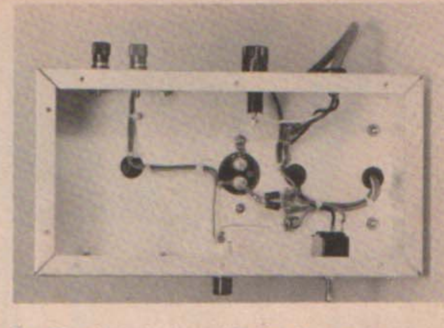


Fig. 33 — Interior view of the power supply.

Fig. 34 — Front view of the 0- to 25-volt, 1.5-ampere power supply.

inches (127 × 241 × 76 mm). Mounted atop the chassis is the power transformer, filter capacitor and regulator. The regulator is attached to a heat sink that measures 3 × 4 1/2 × inches (76 × 114 × 25 mm). Two tantalum capacitors, not visible in the pictures, are mounted at the IC terminals on the underside of the sink. Since good ground connections are required to prevent IC oscillations, remove the anodizing from the heat sink where it will contact the chassis.

The layout of the underside of the chassis can be seen in Fig. 33. Two binding posts (one red and one black) and the fuse holder are mounted on the rear apron. The on-off switch and pilot light occupy a portion of the front panel. Dymo tape labels complete the front panel.

A 0-25 Volt Adjustable Power Supply

The power supply shown in Fig. 34 can supply 1.5 amperes at voltages from near zero to +25. This range will cover most voltages needed for low power amateur work. Digital circuits require 5- and 12-volt supplies and most rf designs require 12 or 13 volts as a power source. On occassion there is need for voltages somewhat higher than this. The unit described here can supply voltages up to 30 volts. Voltage metering is provided to the 25-volt level since using the next larger meter (50 volts) would give rather poor resolution of low-voltage readings.

The Circuit

Power from the ac line is applied to the transformer primary through a single-pole, on-off switch and a fuse, F1. C1 and C2 are spike-eliminating capacitors used to protect the diodes in the transformer secondary circuit. These capacitors also

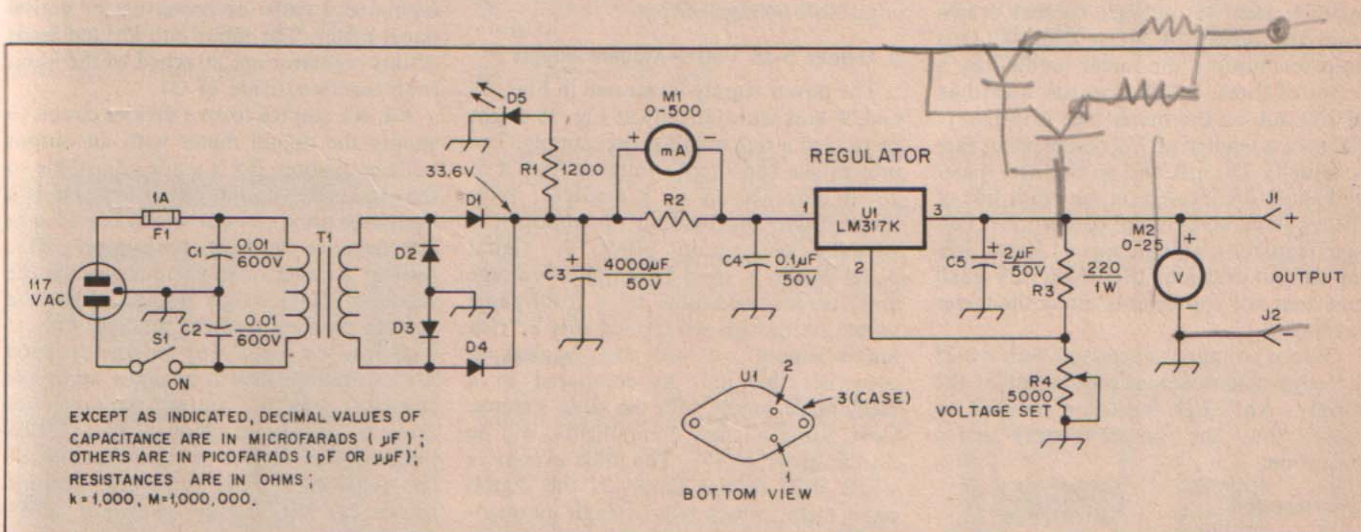


Fig. 35 — Schematic diagram of the 0- to 25-volt, 1.5-ampere power supply. Components not listed in the parts list are for text reference only. R1 is a 1/2-watt carbon type and R3 is a 1-watt carbon resistor. Capacitors are disk ceramic types unless noted otherwise.

C3 — Electrolytic capacitor, 4000 µF, 50 V.
C5 — Tantalum capacitor, 2 µF, 50 V.
D1 - D4, incl. — Silicon diode, 100 V, 6 ampere. HEP RO101 or equiv. Full-wave bridge assembly may be used.
D5 — LED, general purpose.
F1 — Fuse, 1 A.
J1, J2 — Binding post.
M1 — Panel meter, 0-500 mA. Calectro DI-916 or equiv.
M2 — Panel meter, 0-25 V. Calectro DI-923 or equiv.
R2 — Meter shunt. Wind 16- 1/2 inches (419 mm) no. 26 enam. wire on a large-value 2-watt carbon resistor.
R4 — Potentiometer, 5000 ohm, 2 watt.
S1 — Toggle switch, spst.
T1 — Power transformer; primary 117 V, secondary 24 V, 3 A, Hammond HG2G or equiv.
U1 — Regulator, LM317K or equiv.

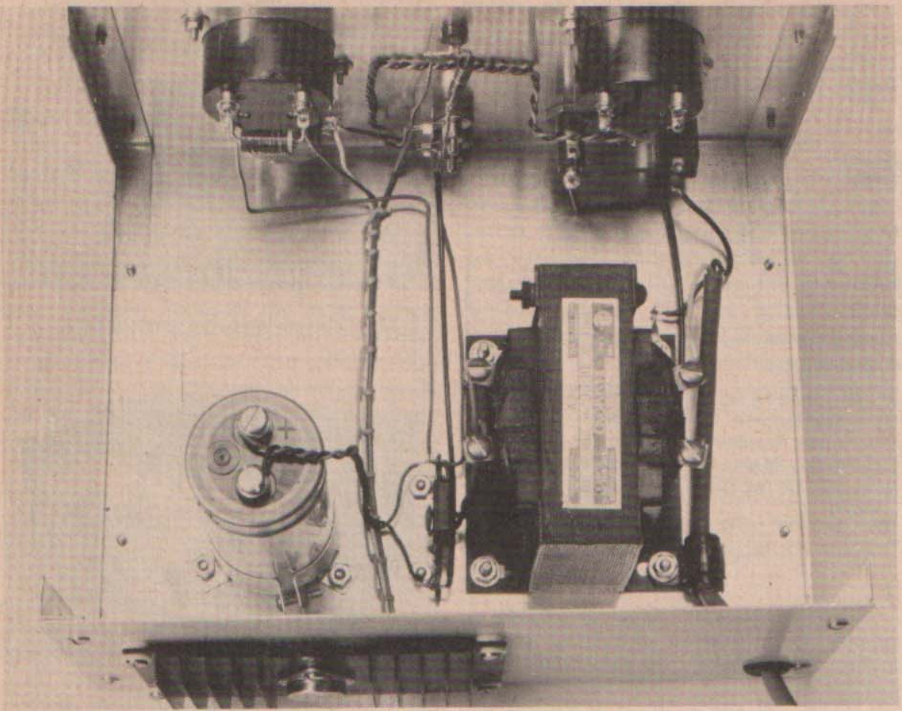


Fig. 36 — Interior view of the regulated power supply capable of delivering 1.5 amperes.

bypass the supply for rf energy entering or leaving the supply through the ac cord.

The voltage from the secondary of the transformer is applied to a full-wave bridge rectifier. C3, a large-value electrolytic, filters the dc. The heart of the supply is an LM317K adjustable-voltage regulator. Voltage adjustment is facilitated by the potentiometer in the voltage-divider circuit at the output of the chip. Bypass capacitors are used directly at the chip for stable operation. A 0-500 mA meter is used to measure current drawn from the supply. A meter shunt (R1) is used to multiply the meter reading by a factor of three. In other words, a reading of 100 mA on the meter face is in reality 300 mA, a reading of 250 on the meter face is actually 750 mA and so on. The meter and shunt are located on the input side of the regulator so as not to compromise voltage regulation at high current levels. Idling current drawn by the '317 is very small and does not appreciably affect the meter reading.

Output voltage is measured with a 0-25 V meter placed across the output of the supply. An LED indicator is used to signal that the power supply is in operation.

Construction

The supply is housed in a homemade enclosure made from sheet aluminum. Overall dimensions of the cabinet are 8 × 10 × 5 inches (203 × 254 × 127 mm). Mounted to the front panel are the two meters, voltage-adjust potentiometer, binding posts, fuse holder, on-off switch and LED indicator. The general layout of the inside of the supply is shown in Fig. 36. Point to point wiring is used throughout. The shunt resistor for M1 is located directly across the meter terminals.

Should the builder wish to limit the maximum voltage available at the output terminals of the supply, a resistor can be placed in parallel with the potentiometer. Additionally, the experimenter may wish to add a switchable voltage divider circuit for "programming" certain standard voltages into the supply. A rotary switch could be used to select, say, 5, 9, 12.6 and adjustable-voltage output.

A Deluxe 5-25 Volt 5-Ampere Supply

The power supply illustrated in Figs. 37 and 39 and schematically at Fig. 38 might be termed a rich man's power supply. The unit shown can supply voltages from 5 to 25 at currents up to 5 amperes. With thermal and short-circuit protection it is virtually destruction proof. A digital panel meter is used to monitor voltage and current, selectable by a front-panel switch. Although we termed this a "rich man's supply", it will cost far less to construct this unit as compared to a ready-made supply with the same features. Cost, using all new components, will be on the order of $75. The most expensive single item in the supply is the digital panel meter, which sells in single lot quantities for around $40 at present. As more companies start manufacturing these items the prices should drop significantly.

The digital readout, however, is not much more expensive than two high-quality meters. The prospective builder should consider this when choosing between the digital panel meters and two analog panel meters. Voltage measure-

Fig. 37.— Front view of the deluxe 5- to 25-volt, 5-ampere power supply.

ments are read directly off the panel meter in volts. Current is measured in amperes with a reading of 0.05 equal to 50 mA.

Circuit Details

The circuit diagram of the power supply is shown in Fig. 38. T1 is a 36-volt, center-tapped transformer rated at 6 amperes. D1 and D2 are used in a full-wave rectifier providing dc output to the filter capacitor, C3, a 34,000-μF, 50-volt electrolytic of the computer-grade variety. The unregulated voltage is fed to U1, a Fairchild 78HGKC regulator, the heart of the supply. This chip is rated for 5-A continuous duty when used with an adequate heat sink. R1 and R2 form a voltage divider which sets the output voltage of the supply. R1 is a ten-turn potentiometer. U1 is bypassed with 2-μF tantalum capacitors directly at the input and output pins.

Z1, as outlined earlier, is a digital panel meter. Connections to the meter are made through a special edge connector supplied with the readout. U2 is used to supply a regulated 5 volts for powering the digital panel meter. The input and ground leads of this regulator are attached to the input (non-regulated) side of U1.

R4, R5 and R6 form a divider circuit to supply the digital meter with an output voltage reading. R5 is made adjustable so that the meter can be calibrated. R3 is a current-sensing resistor which is placed in the negative lead of the supply. This resistor is used on the input side of the regulator (U1) so as not to affect the voltage regulation of the power supply at high load currents. Any voltage dropped across the resistor will be made up by the regulator, so the output voltage will remain unchanged. Notice that U2 is placed to the left or at the input side of the regulator. This is so the current drawn by the readout will not affect current readings taken at the load. Sections A and B of S2 are used to switch the meter between the voltage and current sensors. S3C is used to switch the decimal point in the digital panel meter to read correctly for both voltage and current.

As shown in the schematic, a single-point ground is used for the supply. Used in many commercial supplies, this tech-

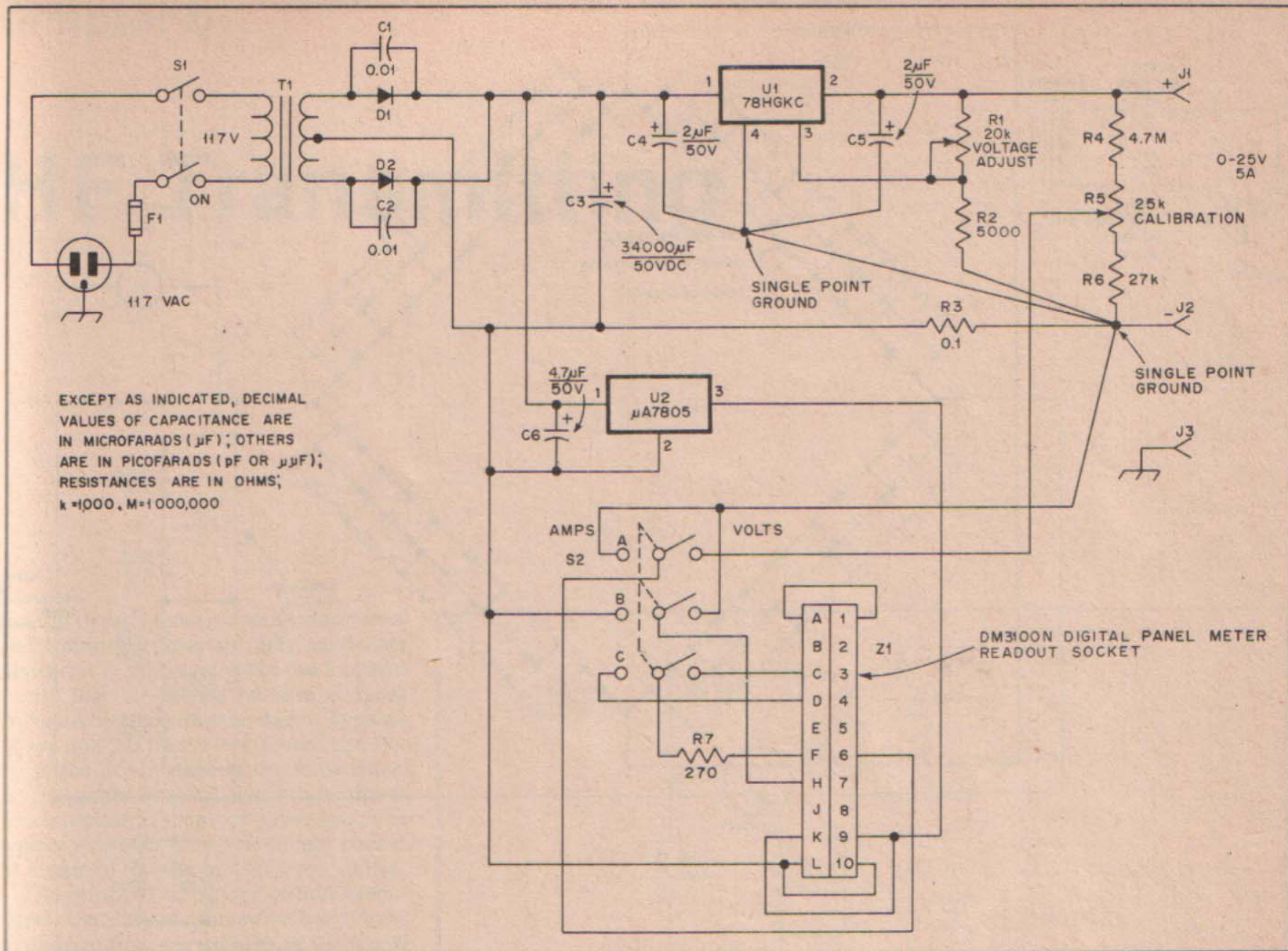


Fig. 38 — Schematic diagram of the deluxe power supply. All resistors are half-watt carbon types unless noted otherwise. Capacitors are disk ceramic unless noted otherwise. Numbered components not appearing in the parts list are for text reference only.

C3 — Electrolytic capacitor, 34,000 µF, 50 V. Sprague 36D343G050DF2A or equiv.
C4, C5 — Tantalum capacitor, 2 µF, 50 V.
C6 — Tantalum capacitor, 4.7 µF, 50 V.
D1, D2 — Silicon rectifier, 100 V, 12 A
F1 — Fuse, 2 A.
J1-J3, incl. — Binding post.
R1 — Potentiometer, 20-kΩ, linear, 10 turn. Clarostat type 731A or equiv.
R3 — Resistor, 0.1Ω, 10 W. Dale R-10 or equiv.
R5 — Potentiometer, 25-kΩ, circuit board mount.
S1 — Toggle switch, dpst.
S2 — Toggle switch, 3pdt.
T1 — Power transformer; primary 117 V, secondary 36 V ct, 6 A. Stancor P-8674 or equiv.
U1 — Regulator, Fairchild 78HGKC or equiv.
U2 — Regulator, µ A7805 or equiv.
Z1 — Digital panel meter. Datel DM3100N or equiv.

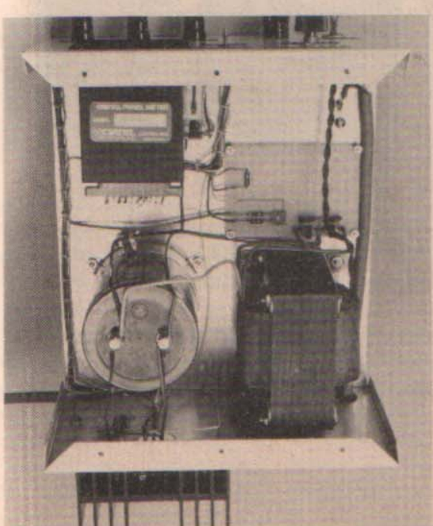


Fig. 39 — Interior view of the deluxe power supply.

nique provides better voltage regulation and stabilization than the "ground it anywhere" attitude. In this supply, the single-ground point is at the front panel binding post labelled MINUS. All leads that are to be connected to ground should go only to that point.

Construction

The deluxe power supply is housed in a homemade enclosure that measures 9 × 11 × 5 1/4 inches (229 × 279 × 133 mm). U1 is mounted to a large heat sink (3 × 5 × 2 inches; 76 × 127 × 51 mm) which is attached to the rear apron of the supply. The front panel sports the digital-panel meter, power switch, binding posts, fuse holder, voltage-adjust potentiometer and meter-selector switch. Although a circuit board is shown in the photograph as supporting R4, R5, R6, R7, D1, D2, C1 and C2 these items could just as well be mounted on terminal strips. For this reason a board pattern is not supplied.

The front and rear panels are spray painted white and the cover is blue. Dymo labels are used on the front panel to identify each of the controls. Cable lacing

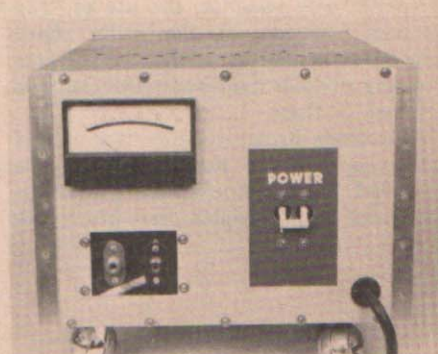


Fig. 40 — Front view of the heavy-duty, 3400-volt power supply.

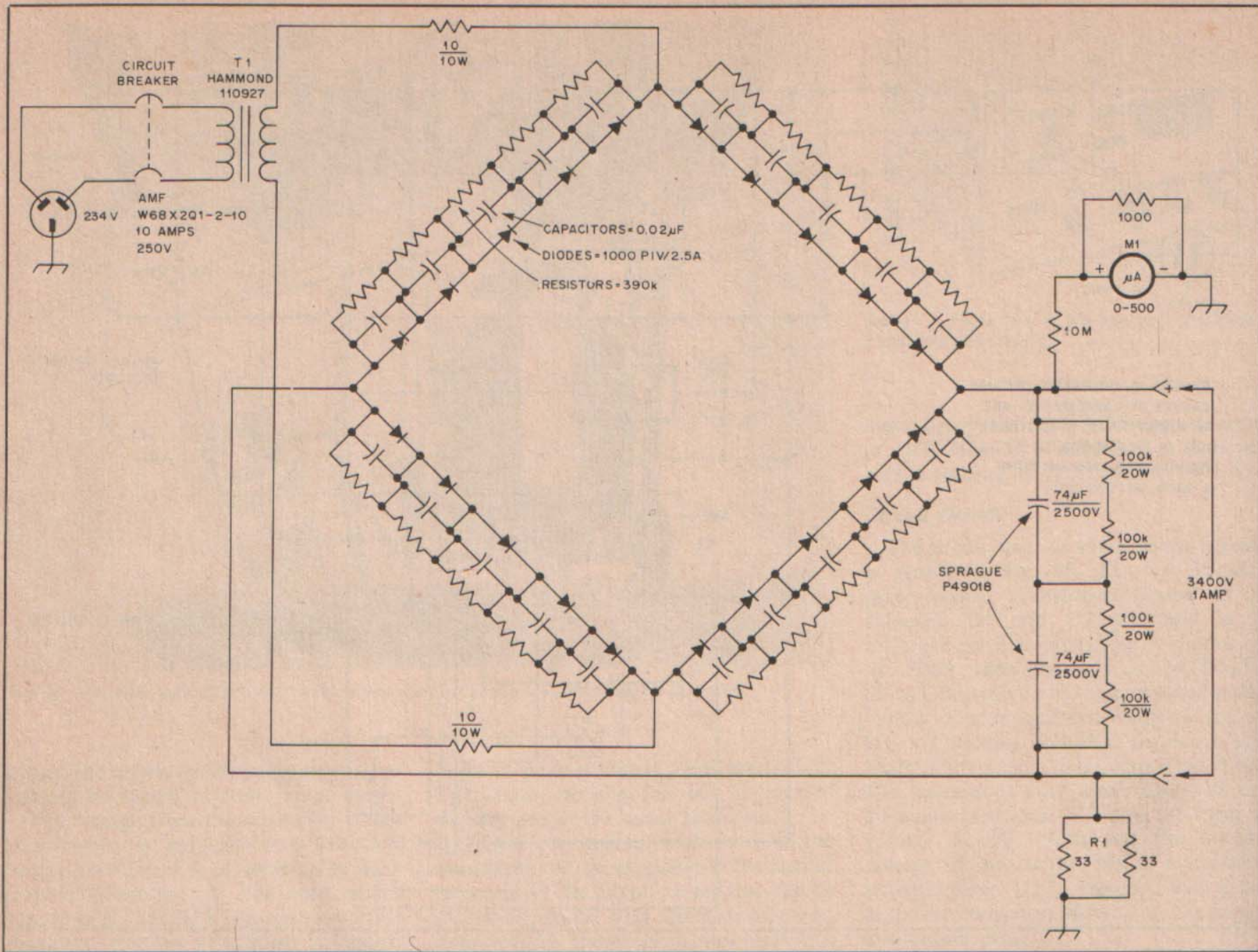


Fig. 41 — Schematic diagram of the high-voltage power supply.

of the various leads adds to the clean appearance of the supply.

A "Sanitary" High-Voltage Supply

Power supplies aren't usually noted for adding decor to the shack. Most hams would rather hide them so that nonham visitors won't ask, "What's that ugly looking thing?" However, an attempt was made to improve the appearance of this model along with the function of providing high voltage for general amplifier purposes. Not all the additions are frivolous. For instance, the use of "rug runners" instead of the usual sharp corners on the bottom of the unit prevents gouging an easily damaged surface such as a bench or floor.

The diode bridge rectifier is mounted on a separate pc board that can be removed easily. Accidental contact is prevented by a Plexiglas sheet which also permits viewing of the circuit board while it is still in the power supply. Although a sheet-metal cutter and bender were used to fabricate the sides, a "cut-and-file" method could result in a similar job if the builder was willing to spend the time. Either that or angle brackets (such as those on the front of the unit) could be used inside of the top and bottom covers in order to form an overlap surface for the covers.

Circuit Details

The power supply employs a full-wave bridge rectifier and is capable of 1-A output at 3400 V dc. Primary-circuit and surge considerations are simplified by the use of 234 V ac instead of 117 V. While the addition of a 234-V line might seem like an unjustified inconvenience, experience has proven this approach to be the most acceptable method. However, additional surge protection is afforded by the use of the 10-ohm, 10-watt resistors in the secondary of T1.

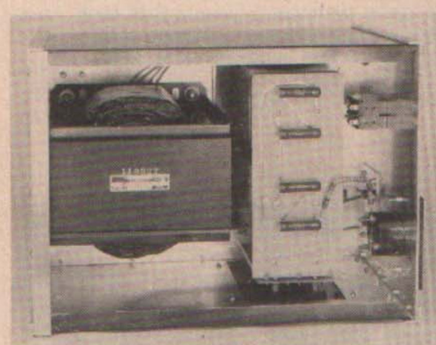


Fig. 42 — Interior view of the heavy-duty power supply. The bleeder/equalizing resistors are mounted to the circuit board which is in turn mounted directly to the capacitor terminals.