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Publication Date

1962-04-06

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Berkeley, California

UCLA Conf. on Sector-Focused Cyclotrons,
Nuclear Instr. and Methods.

UCRL-10020

UNIVERSITY OF CALIFORNIA
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Berkeley, California

Contract No. W-7405-eng-48

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ABSTRACT

Silicon diodes have proved to be convenient and reliable in all types of power supplies throughout the Berkeley 88-inch cyclotron. The paper includes practical design data, a discussion of the pitfalls our experience has taught us to avoid, and comparison of silicon power supplies with other types of power supplies with respect to cost and performance.

APPLICATION OF SILICON DIODES TO CYCLOTRON OSCILLATOR AND MAGNET POWER SUPPLIES[†]

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1. Introduction

At the beginning of the 88-inch cyclotron program, it was decided to review the state of the art of power-conversion equipment, for new concepts or components to provide improved performance or economies. Silicon diodes, appeared not to have the limitations of selenium and germanium rectifiers. The major power-conversion features associated with a cyclotron are low voltage, high current for the magnet, and high voltage, low current for the oscillator. Commercially manufactured silicon rectifiers fit both of these applications by suitably arranging diodes in parallel or in series.

The outstanding characteristics of such rectifiers are high efficiencies, relatively modest cost, and low maintenance.

Silicon diodes are compatible with all of the normal rectifier circuits and voltage controls and, because of their high back resistance, are particularly good in magnetic-amplifier power supplies.

2. Costs

One of the major factors in determining the practicability of any new device is a comparison of its cost with that of a conventional device.

In the case of the magnet power supply, normally a motor-generator set would be used, therefore a comparison of motor-generator-set and

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[†] Work done under the auspices of the U.S. Atomic Energy Commission.

magnet-amplifier cost was made, as shown in fig. 1. It can be seen that, except for the smallest size, there is a substantial cost advantage to the magnetic amplifier. Most of this saving is the result of the lower initial cost of the fabricated power supply, and the modest foundation requirements. There are several other cost advantages, not shown in fig. 1, of magnetic amplifiers: (a) they can be made outdoor construction, which eliminates expensive building space; (b) maintenance costs are reduced because of the static nature of the units; and (c) operating expense is lowered because of their high degree of efficiency.

In the case of oscillator power supplies, the cost saving is not as dramatic, but again savings can be made by outdoor construction, and–compared to vacuum tube supplies–there are no tube-replacement costs.

3. High-Voltage Silicon Power Supplies

In this application, silicon diodes are primarily in competition with mercury-vapor tubes. The advantages of silicon diodes are flexibility, reliability, and in many applications, lower cost. In regard to flexibility: by stocking a few bins of diodes–for example, bins of 0.5-A, 2-A, 6-A and 20-A diodes–a very large variety of power supplies can be assembled as the need arises. Almost any voltage and current combination can be assembled within a few hours and procurement delays are avoided. Silicon diodes lend themselves well to pulsed power supplies because of their very high surge-current capabilities. High-current rectifier assemblies can be built with small heat sinks for low-duty-factor applications. Silicon diodes lend themselves very nicely to outdoor installation, also, thus saving valuable building and rack space.

In regard to reliability, they have the advantage that there is no deterioration such as occurs in the cathodes of the mercury-vapor tubes. The failures that do occur seem to be a result of either excessive voltage transients or defective manufacture.

Cost comparisons between silicon diodes and mercury-vapor tubes favor the silicon diodes in supplies where the current is high for the voltage class of the power supply. Mercury-vapor tubes are usually less expensive than silicon diodes in rectifiers where the voltage and current requirements match the tubes well.

In high-voltage rectifiers many diodes have to be connected in series. To distribute the inverse voltage across the diodes evenly, each diode is shunted by a resistor and a capacitor. The resistance masks the differences in the back resistances of the diodes, and prevents any charge accumulation on the capacitors. The capacitance, although it masks the junction capacitance of the diodes, has the primary function of preventing excess inverse voltage across the diode. If the capacitance across the diode is too small the displacement current flowing in the stray capacitance to ground can produce an excessive inverse voltage across the diode. Even though the stray capacitance is very small, the voltage across it, for those diodes near the A. C. end of a rectifier leg, is often very high—so that the displacement current is appreciable.

The values shown in the chart in fig. 2 have proved to be satisfactory in supplies used in Berkeley. The heat sinks shown in fig. 2 are made of 16-gauge copper. The diode assemblies are mounted on a fiberglass board, and are connected in the circuit shown in fig. 3. We use a sufficient number of diode assemblies in each leg of the bridge so that the diodes' peak inverse voltage is only a third of its rating. This margin is used as protection against transients.

The heat sinks shown in fig. 2 are intended for convection cooling, with the exception of the last one (20-A diodes), where a linear air speed of 10 ft/sec is used.

Because of the high surge-current capacity of silicon diodes, they are excellent rectifiers for power supplies employing crowbars (fault diverters). A typical circuit is shown in fig. 4. Experience indicates that it is best to use an ignitron with a holding anode so that the crowbar is not quenched if the circuit rings but, instead, once established, the arc is maintained until the vacuum switches open the primary circuit. The 2-Ω resistor shown in the positive bus insures that under fault conditions there is still sufficient anode voltage to maintain the arc in the crowbar. The 50-Ω resistor and 0.1-μF capacitor serve as a surge network for the D. C. circuit. The surge network in the primary circuit is designed to protect the rectifier transformer and the silicon diodes from transients when the vacuum switches open. We have found that it is better to put the vacuum switches in the primary of the rectifier transformer, rather than the secondary, so that the limited frequency response of the rectifier transformer can be used to suppress the transmission of very high frequency components associated with the snap-out currents of the arcs in the vacuum switches. Ahead of the vacuum switches, lightning arrestors are used to protect the incoming transmission line from transients.

We have measured the speed of typical A.C. overload relays, and found that it is about 10 msec, and also that vacuum relays operate within about 10 msec. In addition, the arc in the vacuum switch may persist for as long as 8 msec following mechanical openings. Thus, following a crowbar operation there may be a period as long as 28 msec before the power supply is de-energized. It is not difficult to design a silicon rectifier to withstand the short-circuit current for this period.

In designing the surge network, it is necessary to know the snap-out current of the vacuum switch. This was measured to be about 16 A in a typical case vacuum-switch operation in a 13 kV line). The surge network can be designed to provide critical damping. However, by permitting the first overshoot to be the same height as the initial surge voltage (see fig. 5) the power dissipation in the network is reduced by approximately a factor of 10. For this case the design equations are:

$$R = \frac{V_{max}}{I_0} \qquad (1)$$

$$Z_0 = 1.67\ R, \qquad (2)$$

where

$Z_0 = \sqrt{\frac{L}{C}}$,

V_{max} is the maximum permissible surge voltage per phase,

I_0 is the snap-out current of the vacuum switch,

C is the surge network capacitance in F, and

L is the total transformer leakage inductance per phase referred to the transformer primary winding.

It is reasonable to limit the surge voltage to 20% of the line-to-neutral voltage in most cases.

4. Magnet Power-Supply Design

Figure 6 shows a typical diagram of a magnetic-amplifier power supply suitable for magnets. In the design of low-voltage, high-current power supplies, the following items must be considered:

(a) Diode rating,

(b) Diode cooling,

(c) Type of voltage control,

(d) Transformer and reactor design,

(e) Filtering,

(f) Magnet protection against cell failure.

5. Diode Rating and Cooling

The current rating that can be assigned to a particular silicon diode is for the most part determined by the cell cooling. For accelerator applications, water cooling is most efficient and offers no particular problems, because satisfactory cooling water is always available as it is required for other parts of the accelerator. The most common method of providing water cooling is to screw the diode into a copper block that is provided with water-cooling passages. Figure 7 shows the temperature rise versus current of an International Rectifier Corporation Series 70 diode when screwed into a copper block 3 by 1-3/4 by 1-3/4 in., through which water was passed at a flow rate of 4 gpm. All tested rectifiers failed in the vicinity of 550 to 600 A average D.C. The temperature of the base at failure was about 240 to 250° C. This coincides with the manufacturer's failure rating of 200° C for the wafer. International Rectifier Corporation rates their diodes full current when the base temperature is 100° C. Based on this criterion, we recommend a rating of 350 A D.C. per diode with water cooling as described.

High-current silicon diodes are available with peak inverse rating of 50 to 500 V. They are particularly sensitive to puncture by peak inverse voltages that exceed their rating. Complete destruction will occur even if this voltage is exceeded momentarily. Consequently, care should be excercised to rate them conservatively, and to provide means for preventing transients, etc. from appearing across the diode. De-rating the peak inverse voltage by dividing by three is recommended unless all transients are known to be filtered or suppressed.

To protect the power supply and system from silicon diode internal faults, a high-speed current-limiting fuse is used in series with each diode. Fuses of this type have been developed specifically for the protection of semiconductor diodes with special attention given to the problem of limiting recovery voltages.

For a power supply having an output current larger than can be achieved by one set of diodes, parallel diodes are required. With no provision made for forced-current balancing, de-rating factors for parallel diodes must be applied, even with carefully matched diodes. A far better means is the use of current-balancing reactors. If they are well designed the first cost of these reactors will be offset by the reduction in the number of diodes required to carry a given current, and they will not add appreciably to the diode and copper losses. Additional savings are realized in the longer life of the diodes.

To achieve proper voltage division where multiple series-connected cells are required, voltage-divider resistors and capacitors are connected across the individual cells. This method works well and is practical for the lower-current diodes. However, for high-current applications it produced excessively high losses and seriously impairs the efficiency of an inherently efficient device.

In the early days, it was common practice to use a separate transformer, or at least a separate transformer winding, for each of the series diodes, so that accurate voltage division was insured. To a large extent this nullified the economy of large units since, in effect, there were two or more complete rectifiers of individually low-voltage ratings connected in series to obtain the required output.

A method using relatively small and inexpensive voltage-divider transformers has been developed and is in successful operation. The economy of full-voltage transformer windings is retained, and absolutely rigid voltage division is obtained without added losses. This system has the additional advantage of providing ready means for detecting a single bad diode in a series string and taking corrective action without first inducing failure in the remaining series diodes.

6. Voltage Control

For most accelerator applications, a continuous wide-voltage adjustment is usually required and can be achieved by three methods:

(a) Step-voltage regulators,
(b) Induction regulators,
(c) Self-saturating reactors.

(a). Step voltage regulators. For applications where a number of voltage steps can be tolerated, a standard step-voltage regulator will do an excellent job with little effect on overall efficiency and power factor. Because of the size of the tap-changing mechanism and the speed with which they can be moved, step-voltage regulators are not applicable to jobs requiring fast response.

(b). Induction regulators. By their nature, induction regulators provide stepless control over wide ranges and affect efficiency and power factor only slightly more than step regulators. Because of the absence of contact mechanisms, their speed of response can be made faster than that of step regulators.

(c). Self-saturating reactors. Self-saturating magnetic amplifiers provide excellent stepless control and can be made extremely fast. The losses incurred in these reactors will generally be somewhat greater than in step or induction regulators of the same range. The power factor of the system decreases directly as the direct voltage is reduced.

For magnet power-supply applications, the dc type of magnetic amplifier is most economical–i. e., placing the self-saturating reactors in the secondary of the rectifier-transformer circuit. For high-voltage application, where the reactor insulation might be a problem or expensive, the self-saturating reactors may be located in the A.C. line along with a separate set of silicon diodes.

7. Transformer and Reactor Design

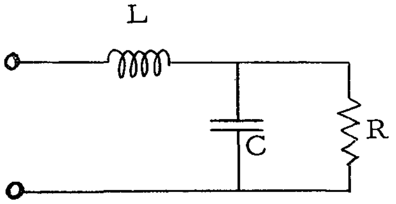
Where space is at a premium, water cooling of the rectifier transformer, magnetic-amplifier reactors, and filter choke can be done. This is easily accomplished by using hollow copper conductors on the high-current windings, and passing water through the coil. The low-current winding, usually the primary winding, can be interleaved with the water-cooled coil. By water cooling in this manner, the current density can be increased and the physical size of the unit is substantially reduced.

8. Filtering

The high percentage of ripple produced by a magnetic amplifier can be reduced by simple LC filters; or, in the case of loads such as magnets, a combination with an eddy-current or skin-effect shield is a very good solution. In this case, a high percentage of ripple current is permitted in the magnet coils, but a sufficiently thick nonmagnetic shield is placed between the coil and the fringing field. The ripple in the magnet gap itself is ordinarily negligible, because of the small skin depth in steel. Figure 8 shows how the ripple decreases with shield thickness for the usual ripple frequencies.

With very large electrolytic capacitors now available from several manufacturers at low cost, it is feasible to build filters for low-voltage, high-current magnets. The following paragraphs give design information for these filters, most of which require a few hundred microhenries of inductance and about 0.1 F of capacitance.

The recommended filter circuit is shown below. It is an "L" half-section low-pass filter.



where

$$L = \frac{\sqrt{2}\, R}{\omega_c}, \tag{3}$$

$$C = \frac{1}{\sqrt{2}\, \omega_c R}, \tag{4}$$

$$K = \frac{1}{\sqrt{1+\left(\frac{f}{f_c}\right)^4}}. \tag{5}$$

If the inductance and capacitance are chosen according to eqs. (3) and (4), and the filter is terminated in a pure resistance R, the transfer function [eq. (5)] rolls off smoothly. Figure 9 shows the cutoff frequency required to produce a given amount of ripple reduction. In a design problem, one simply decides how much attenuation is needed for each ripple component. The, the required cutoff frequency can be chosen from fig. 9, and the inductance and capacitance from fig. 10. The latter is simply a plot of eqs. (3) and (4).

There are a number of practical factors to be considered when designing a filter. First, the transfer function of the filter rolls off smoothly only when terminated in the proper resistive load. Usually a magnet is not a resistive load. Fortunately, the deviations in the transfer function when the filter is not properly terminated occur only in the vicinity of the cutoff frequency. If the filter is too lightly loaded, the transfer function peaks at the cutoff frequency; if too heavily loaded, the function sags at f_c. Since the ripple components to be attenuated by the filter necessarily occur at several times f_c, they are not appreciably affected by the mismatch of the load to the filter. However, care should be taken to see that a lower ripple frequency, which otherwise might be negligible, is not increased by the filter. For example, if a filter is designed to attenuate a 360-cycle component, and the load does not properly match the filter, f_c should be chosen to be either above or below 60 cycles by at least 40% in order to avoid increasing the 60-cycle component which might be present.

A second practical consideration concerns the ripple current in the capacitor bank. The number and size of the individual capacitors that make up the bank must be so chosen that the ripple current per unit is within the manufacturer's rating. Figure 11 shows the maximum ripple current in the capacitor bank when the filter is connected to a magnetic amplifier that modulates the input voltage 100%. This is the most severe case.

A third consideration concerns the voltage rating of the capacitors. Electrolytic capacitors form themselves, over a period of time, to the voltage at which they are operating. One, therefore, does not gain a safety factor by de-rating their voltage in the initial design.

9. Sample Problem

A 90-V 2000-A magnet is to be excited by a magnetic amplifier. Suppose that after consideration of the natural eddy-current shielding provided by the coil tanks, vacuum tank, and magnet pole (see fig. 8), it is decided that the 360-cycle ripple component should be attenuated by a factor of 100. Figure 9 shows that the cutoff frequency of the filter should be 36 cps.

For this cutoff frequency and a magnet resistance of 0.045 ohms, fig. 10 shows that the inductance should be 280 μH, and the capacitance 0.070 F. Figure 11 shows the maximum ripple current in the capacitor bank to be 52 A rms. We must select enough capacitors to carry the 52 A and total 70 000 μF of capacitance.

A suitable capacitor is the Sangamo type 15C/100–a1500-μF, 100-V unit. The catalog gives a maximum ripple current of 1.2 A per unit for this capacitor. In order to make up the 70 000 μF bank, 47 such capacitors would be required. The total permissible ripple current would be 56 A.

Table 1 shows the ripple currents of various capacitors manufactured by the Sangamo Company.

10. Magnet Protection Against Cell Failure

In the event that all rectifier diodes fail, some provision must be made to provide a current path for the inductive load current of the magnet–to prevent having an excessively high voltage puncture the magnet dielectric.

A group of unfused silicon diodes are connected across the output of the power supply, or, better yet, across the terminals of the magnet, to carry this inductive current. These "free-wheeling" diodes must carry the full magnet current for the time constant of the magnet. Water cooling may be used on the diodes, but a better way of cooling the diodes is to mount them on a copper block of sufficient size so that all of the energy of the magnet would not raise the temperature of the block above the danger point of diode failure.

11. Conclusion

On the 88-inch cyclotron, we have found that using silicon diodes in magnetic-amplifier power supplies and in high-voltage application is a very reliable and economical way of providing power conversion, compared to the motor-generator sets and mercury-vapor tubes used in the past.

Table 1

Sangamo type DCM electrolytic capacitors

Catalog No.	Capacitance (μF)	Working voltage (D.C.)	Ripple current (A rms)
45M/5	45 000	5	6.75
6M/15	6 000	15	2
13M500/15	13 500	15	3.7
20M/15	20 000	15	5
35M/15	35 000	15	7
8M/25	8 000	25	2.5
10M/25	10 000	25	3
15M/25	15 000	25	4
25M/25	25 000	25	6
20M/30	20 000	30	5
1M/50	1 000	50	0.85
1M750/50	1 750	50	1.1
5M/50	5 000	50	1.9
8M/50	8 000	50	2.7
10M/50	10 000	50	3.2
8M/55	8 000	55	2.7
10M/55	10 000	55	3.2
2M/60	2 000	60	1.1

Table 1 (Cont.)

Sangamo type DCM electrolytic capacitors

Catalog No.	Capacitance (μF)	Working voltage (D. C.)	Ripple Current (A rms)
1M/75	1 000	75	0.95
5M/75	5 000	75	1.9
6M/75	6 000	75	2.2
15C/80	1 500	80	1.1
5M/90	5 000	90	1.9
15C/100	1 500	100	1.2
4M/100	4 000	100	1.6
25C/150	2 500	150	1.32
3M/150	3 000	150	1.4
35C/150	3 500	150	1.52
1M250/180	1 250	180	1.24
7C50/200	750	200	1.05
15C/200	1 500	200	1.3
12C/250	1 200	250	1.34
11C/300	1 100	300	1.4
8C/450	800	450	1.35

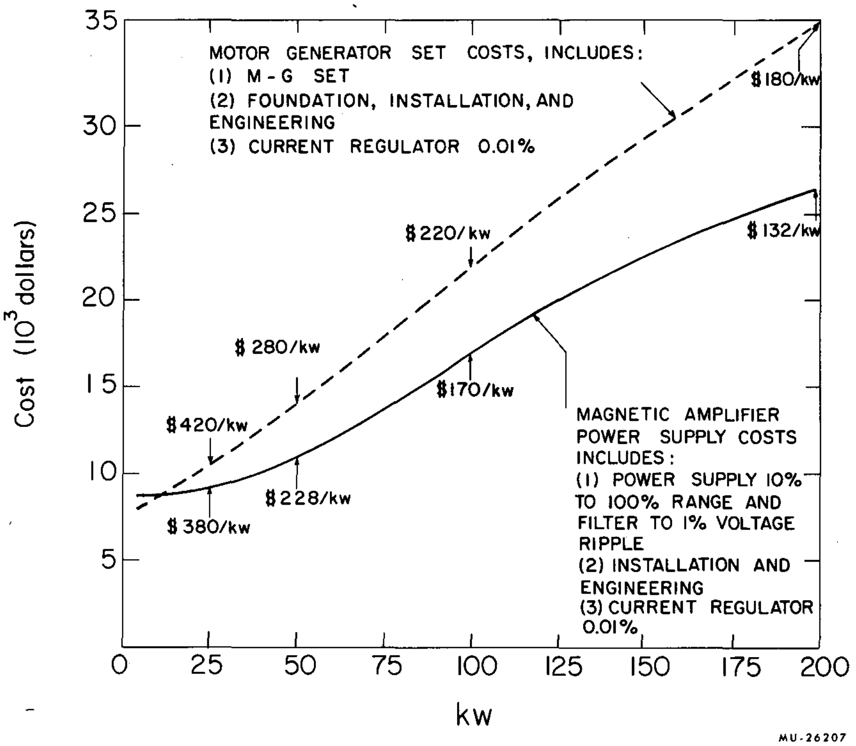
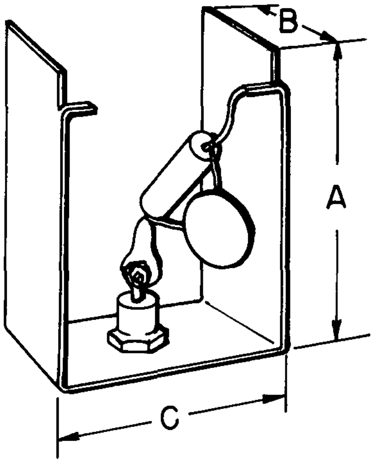


Fig. 1. Cost comparison of motor-generator sets and magnetic-amplifier power supplies.



Diode	Res.	Cap.	A (in.)	B (in.)	C (in.)
0.5 A	510k, $\frac{1}{2}$W	0.001 μF	0	0	0
2 A	120k, $\frac{1}{2}$W	0.01 μF	1"	$\frac{7}{8}$"	$\frac{5}{8}$"
6 A	47k, 1 W	0.05 μF	$2\frac{1}{4}$"	$1\frac{1}{2}$"	$1\frac{1}{2}$"
20 A	15k, 5 W	0.22 μF	3"	$1\frac{1}{2}$"	$1\frac{1}{2}$"

MU-26326

Fig. 2. Diode assembly data.

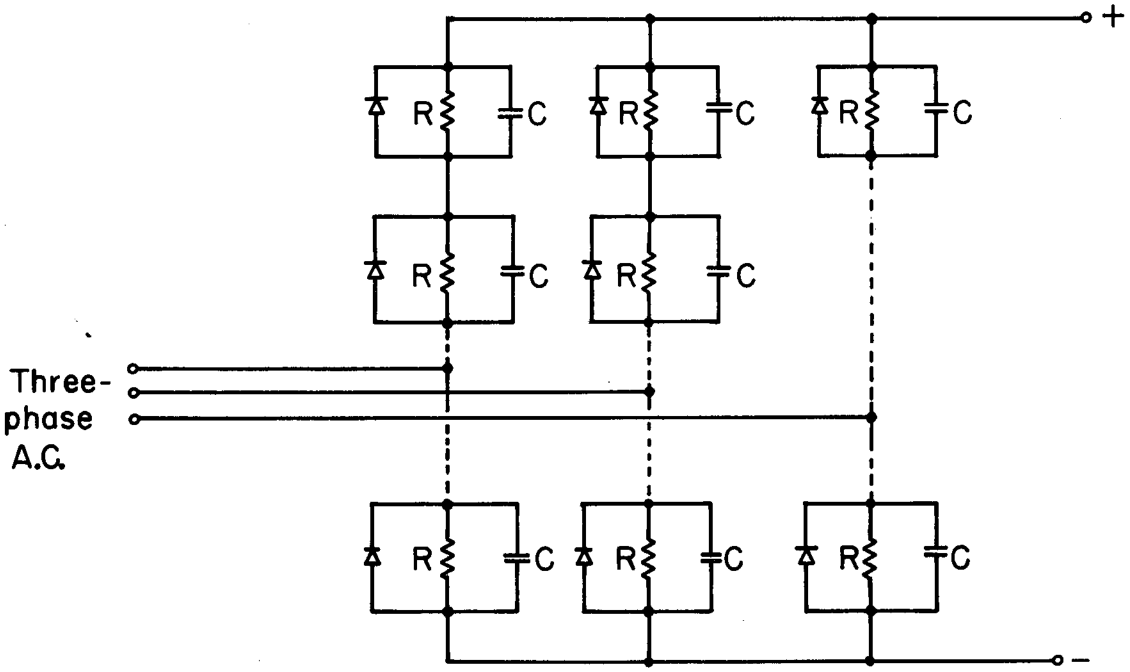


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Fig. 3. Typical high-voltage silicon rectifier circuit.

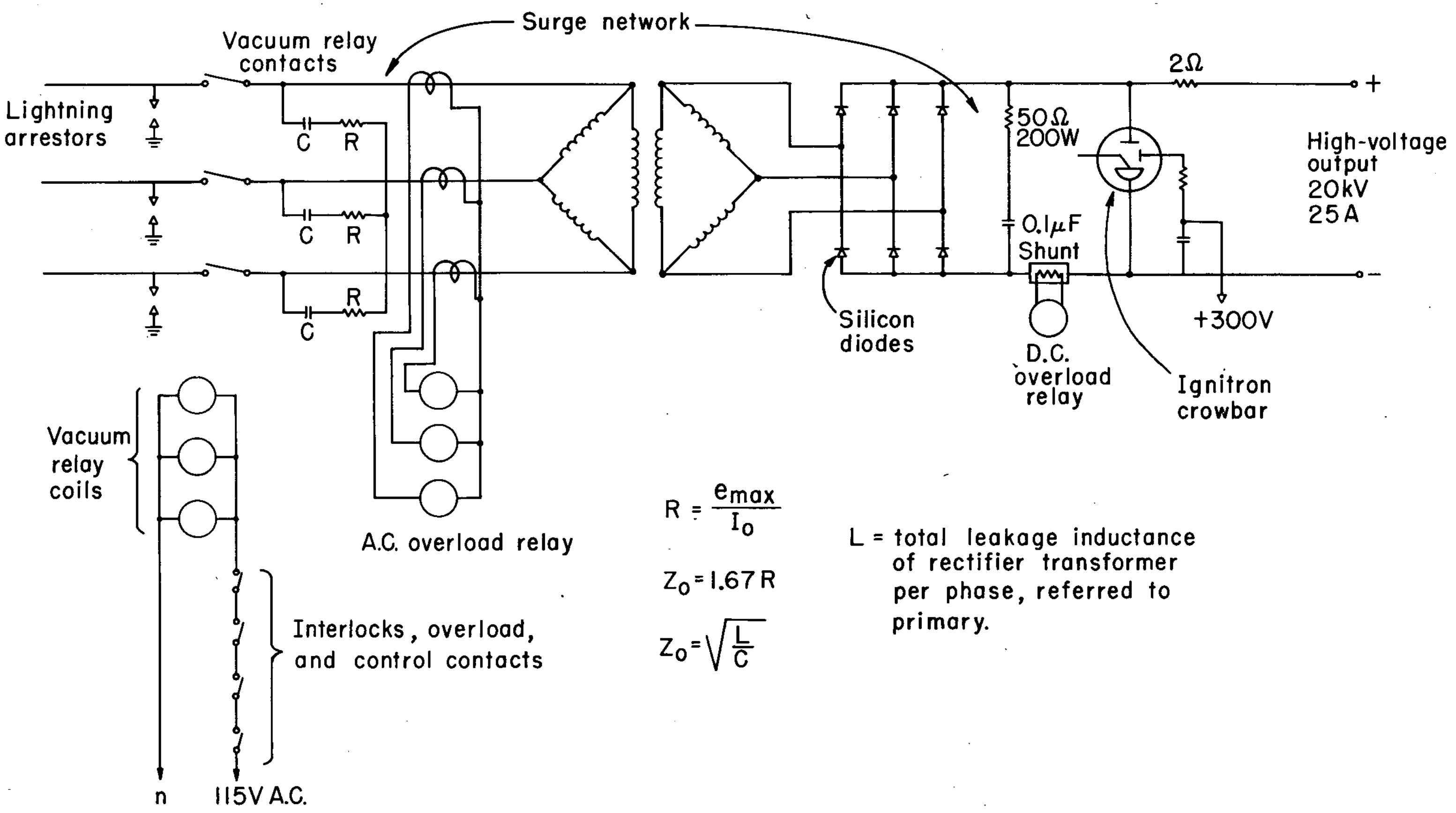

Fig. 4. Circuit diagram of typical oscillator power supply with a crowbar.

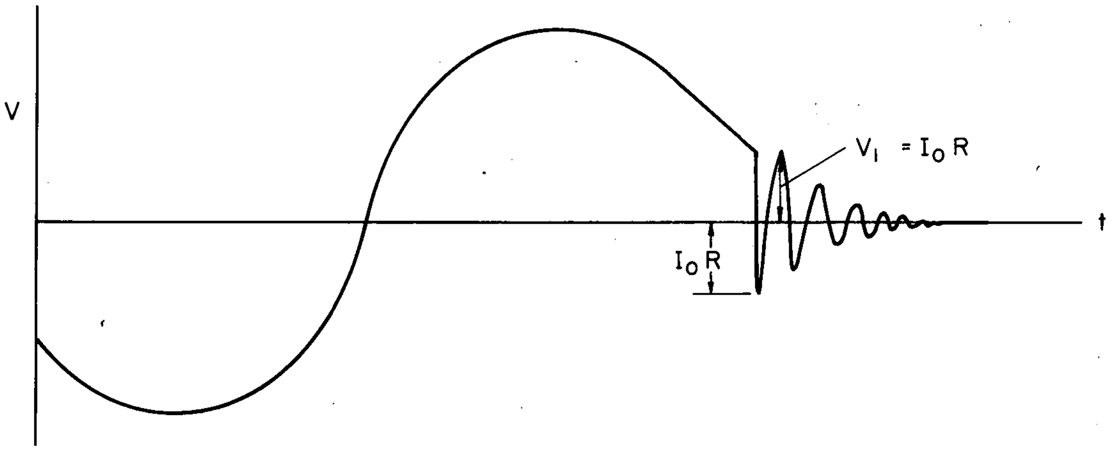

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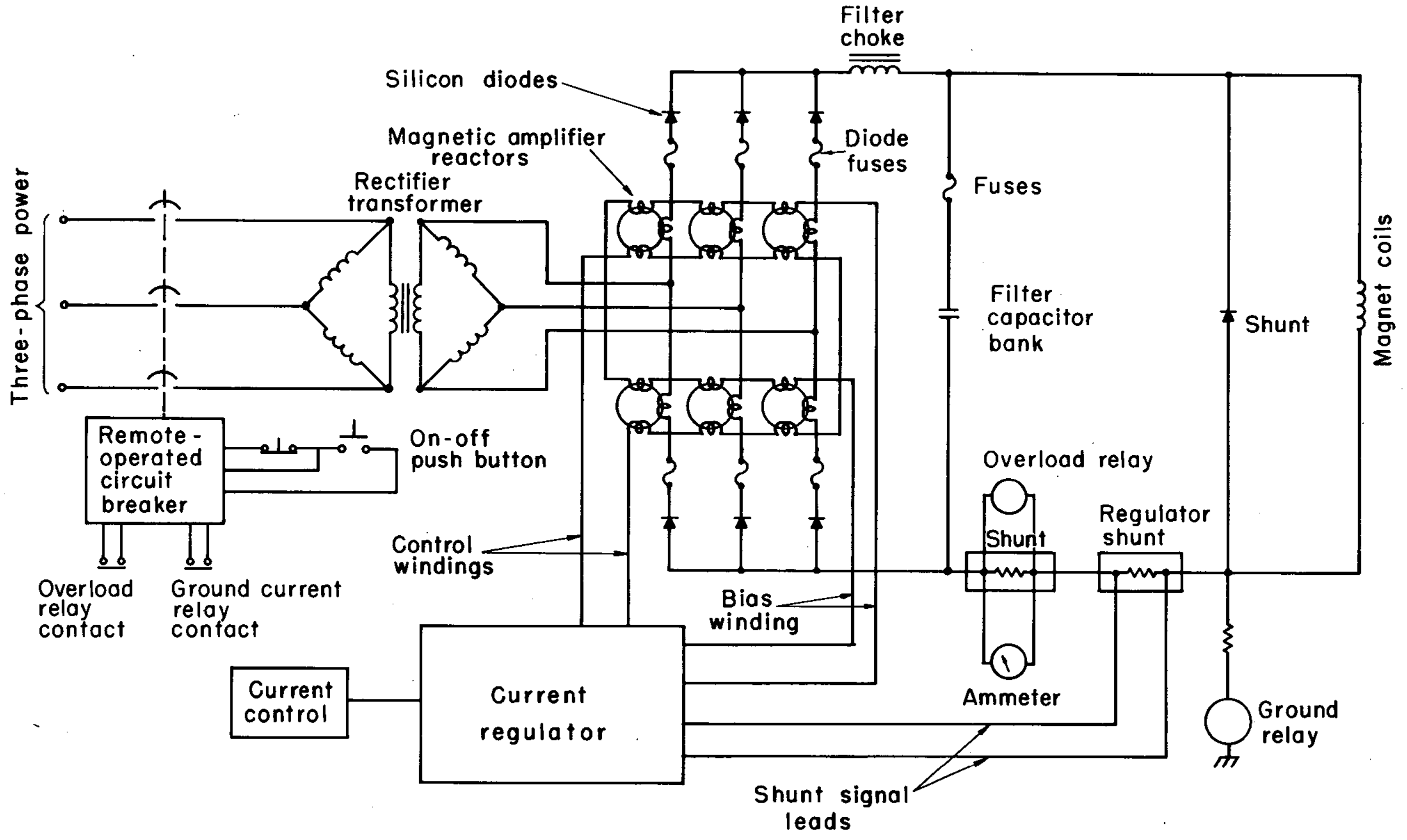
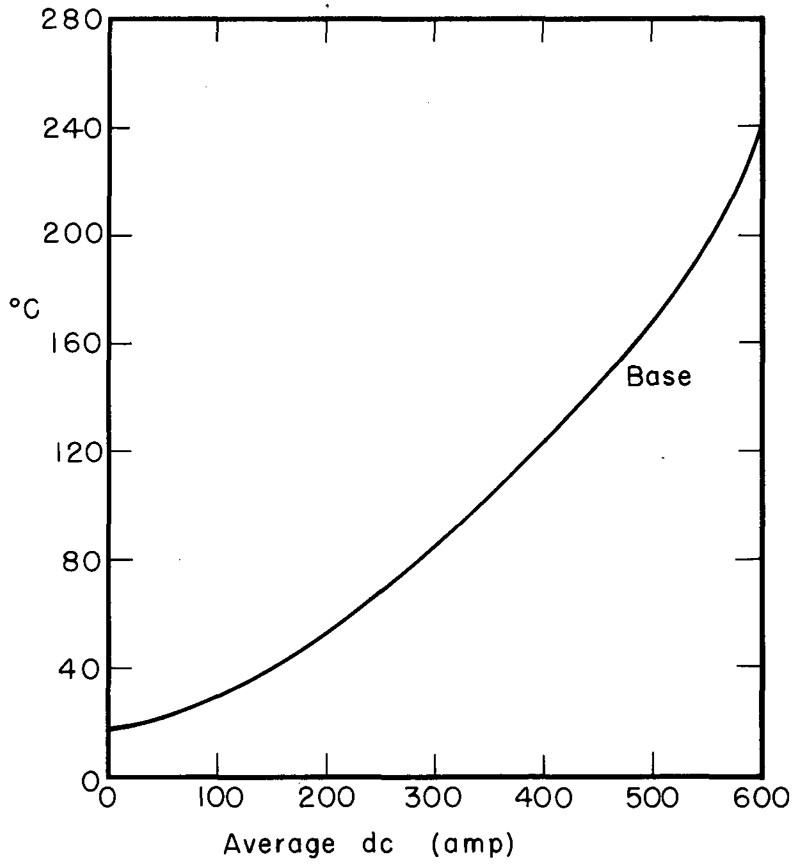
Fig. 5. Line-to-neutral voltage for power supply of Fig. 4. during current interruption by vacuum switch. Surge network limits transient in such a way that first overshoot, V_1, equals initial surge.



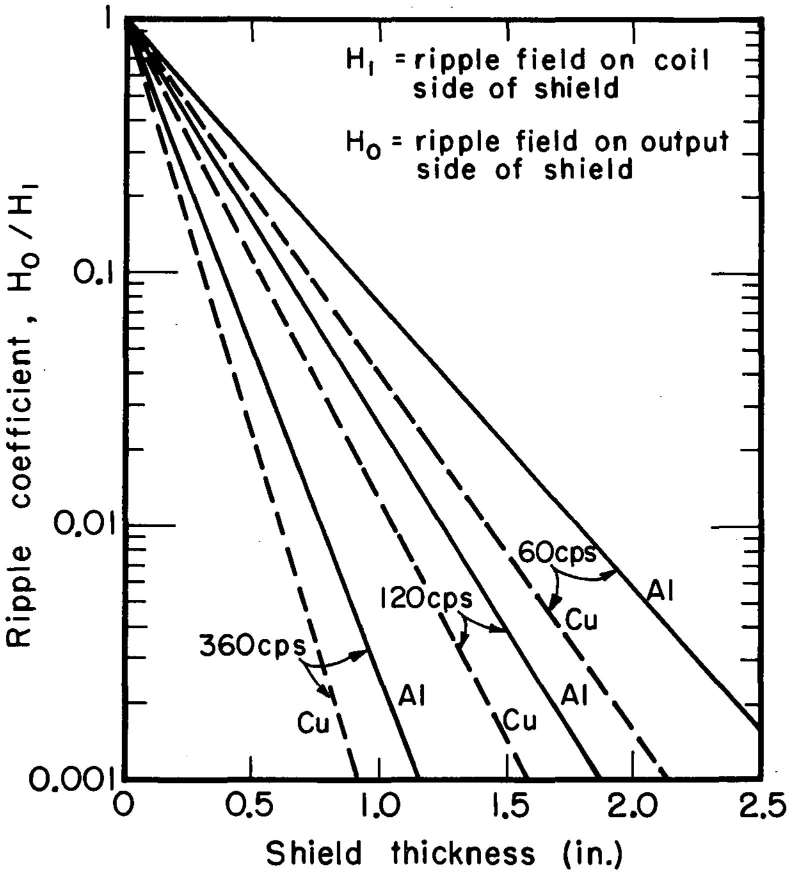
MUB-1001

Fig. 6. Circuit Diagram of a typical magnetic-amplifier power supply.



MU-26320

Fig. 7. Base temperatures vs current for IRC type 70 diodes.



MU-26319

Fig. 8. Eddy-current shield ripple coefficient vs thickness.

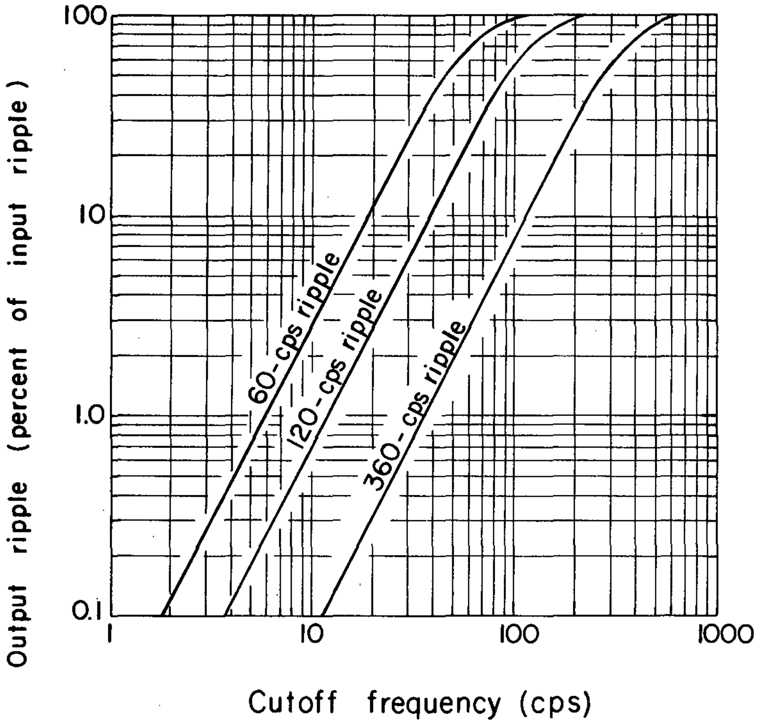


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Fig. 9. Output ripple vs filter cutoff frequency.

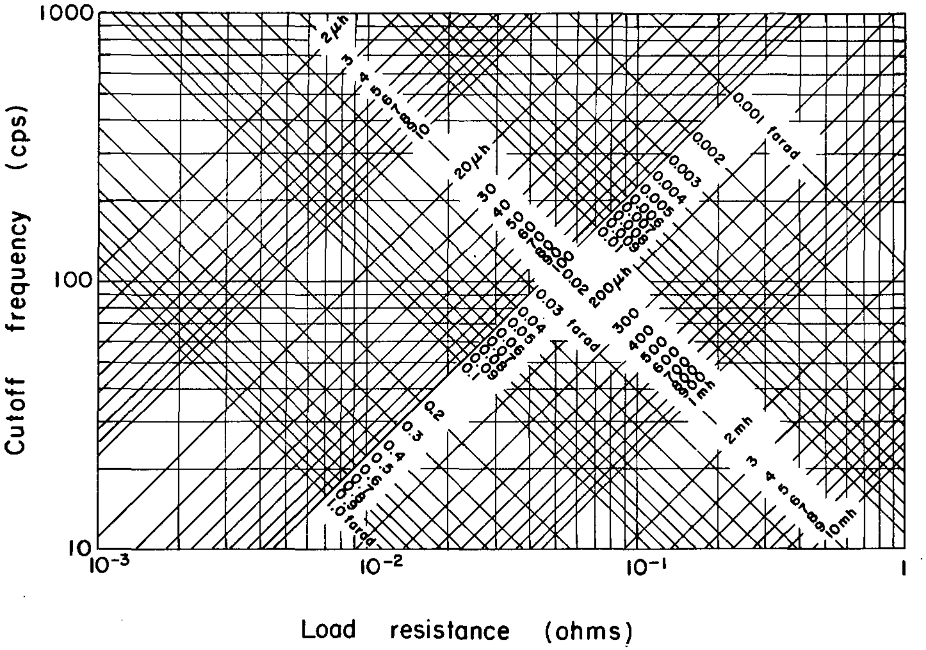

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Fig. 10. Magnet filter design chart.

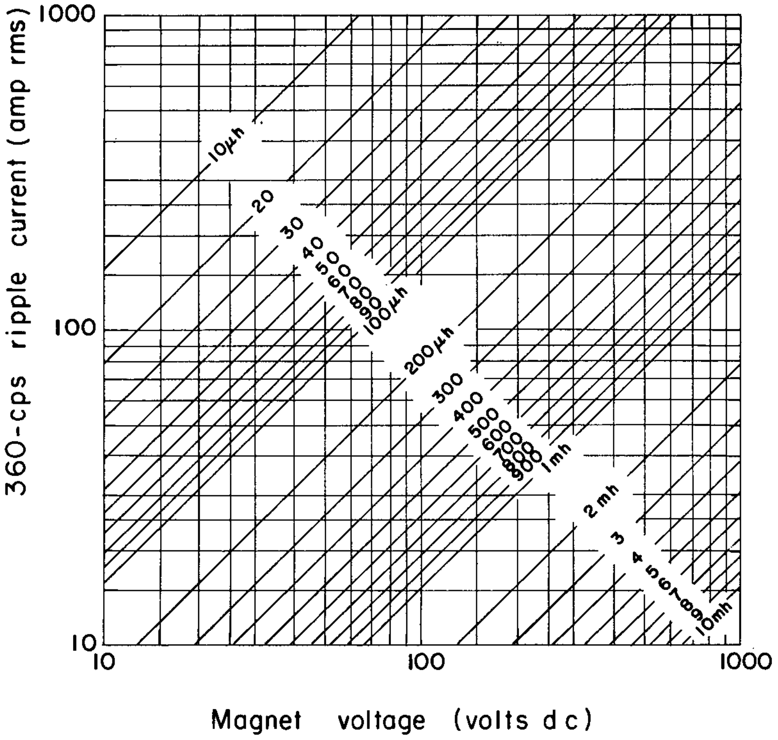

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Fig. 11. Filter capacitance ripple current vs output voltage (this assumes 100% modulation of input voltage by magnetic amplifier).