

# CHAPTER 14

## MAGNETIC AMPLIFIERS

### INTRODUCTION

The magnetic amplifier enjoyed its maximum prominence in power control and low-frequency signal-processing electronics from about 1947 to 1957. By 1957 junction transistors were readily available. Development rapidly shifted away from magnetic amplifiers toward transistor and semiconductor switch equivalents, combined magnetic/transistor amplifiers, and the host of new devices made possible by the joint use of square-loop cores and transistors. Complex magnetic-amplifier designs developed in this period must be looked on as an interim technology. Many magnetic amplifiers in present use and manufacture are the result of early design commitments which have not yet been replaced.

There are a few areas, however, in which magnetic amplifiers continue to excel. In power control, they tolerate extreme environmental and overload conditions that would be fatal to semiconductors. They may also generate less noise because of the slower switching saturable reactors. Perhaps most important, they permit the summing of a number of input signals that must remain electrically isolated. In instrumentation amplifiers, magnetic-amplifier circuits still offer high, drift-free gain with this summing feature. The development of magnetic-core transistor oscillators makes it possible to supply ac power of practically any desired frequency for these amplifiers, making them much smaller than they would be with power-frequency excitation. Similar circuits have come into increasing use in magnetometry where the unique direct transducing capability of the magnetic amplifier puts it in a class by itself.

The magnetic-core transistor oscillator, which is capable of inverting dc to ac up to about 100 kilohertz and by rectification can convert a single primary dc power source at high efficiency to several independent conductively isolated dc voltage supplies, today has all of the engineering prominence that magnetic amplifiers once enjoyed. Many of these newer circuits are, in fact, regulated and timed by magnetic-amplifier principles.

The history and present state of the art in magnetic amplifiers is documented in the proceedings of the Conference on Nonlinear Magnetics and Magnetic Amplifiers [1], and the more recent *IEEE Transactions on Magnetics* [2]. Several books ranging between texts and advanced treatments of design principles are also available [3-7]. The development of soft magnetic materials used in magnetic-amplifier circuits can be traced in the proceedings of the Annual Conferences on Magnetism and Magnetic Materials [8]. Most of the material to be found in these references deals with basic principles and the physical properties of materials. For the circuit engineer who wishes to capitalize quickly and effectively on the design rules that have emerged from this effort, the design manuals available from the manufacturers of magnetic-amplifier core materials tabulate the available material types and related design information such as wire holding capacity, insulation, and temperature characteristics. Having made the basic selection of the core windings and circuit configuration, the designer should expect to spend some time tailoring the circuit to meet design specifications.

### PRINCIPLE OF THE MAGNETIC AMPLIFIER

The elementary principle of magnetic amplification can be conveniently represented in terms of a flux-actuated switch in series with a load. The magnetic material is characterized by the nearly rectangular hysteresis loop of Fig. 1. In this figure, the narrow hysteresis loop corresponds to the loop measured at dc and the wider loop is that measured at the power supply frequency. It should be noted that this dynamic loop widens with increasing frequency. Figure 2 shows the dc hysteresis loops corresponding to three of the materials listed in Table 1.

Figure 3 shows a winding on a core in series with a resistor representing the load. At the beginning of a positive half-cycle of the supply voltage, the

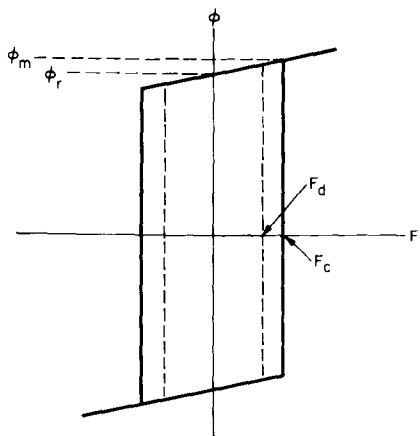


Fig. 1—Schematic representation of dynamic and dc (dotted lines) hysteresis loops.

core is in some initial flux state  $\phi_0$ . Essentially all of the supply voltage is impressed on the core winding until the flux in the core reaches saturation. In saturation, the core becomes a very low impedance and practically all of the supply voltage appears across the resistor. This situation is diagrammed in Fig. 4A for a sinusoidal supply voltage and in Fig. 4B for a square-wave supply voltage. The lower portions of the diagrams show that the switching of the supply voltage from the core to the resistor becomes perfect when the width of the hysteresis loop is zero. In practical designs, the choice of power supply voltage and frequency, core size, winding, and load impedance is subject to the constraints of the problem. To approximate the above-mentioned ideal conditions is often the main object of the design.

Figure 5 shows integrals of portions of the supply-voltage integral in analytical form. It is clear from this figure that the average voltage applied to the load is a function of the switching angle  $\alpha$ , which in turn depends on the initial flux  $\phi_0$ . The half-cycle average of the load voltage is expressed in terms of  $\phi_0$  for the sine-wave and square-wave cases as follows.

Sine Wave	Square Wave
$\bar{v}_r = (2/T) (E_s/\omega) \times (1 + \cos \alpha)$	$\bar{v}_r = E_s [1 - (\alpha/\pi)]$
$(E_s/\omega) (1 - \cos \alpha) = (E_s/\omega) - N\phi_0$	$(T/2) E_s (\alpha/\pi) = N(\phi_m - \phi_0)$
$\cos \alpha = \phi_0/\phi_m$	$\alpha/\pi = \frac{1}{2} [1 - (\phi_0/\phi_m)]$
$\bar{v}_r = (2/T) (E_s/\omega) \times [1 + (\phi_0/\phi_m)]$	$\bar{v}_r = E_s/2 [1 + (\phi_0/\phi_m)]$

It is further obvious that in order for there to be no output, and no excess flux capacity in the core (normal excitation), the flux linkage capacity of the core must be set equal to the volt-second capacity of the power supply. This results in the simple equations

$$E_s = B_m A \omega N \quad (\text{sine wave})$$

$$E_s = (2/\pi) B_m A \omega N \quad (\text{square wave})$$

relating the peak value of the supply voltage, the maximum flux density of the core (in webers/meter<sup>2</sup>), the material cross-section (in meters<sup>2</sup>), the angular frequency, and the number of turns.

Correspondingly, the exciting current for a given coercive force  $F_c$  in ampere-turns is  $i_c = F_c/N = H_c l/N$ , where  $H_c$  is in ampere-turns/meter and  $l$  is the mean length of the magnetic path in the core in meters. These equations in CGS units become

$$E_s = B_m A \omega N \times 10^{-8} \quad (\text{sine wave})$$

$$E_s = (2/\pi) B_m A \omega N \times 10^{-8} \quad (\text{square wave})$$

where  $B_m$  = gauss,  $A$  = cm<sup>2</sup>,  $E_s$  = volts (peak), and  $N$  = turns.

$$I_m = 0.794 H_c l / N$$

where  $H_c$  = oersteds,  $l$  = cm, and  $I$  = amperes.

Although the above discussion contains most of the ideas basic to magnetic amplifiers, no mention has been made of how  $\phi_0$  is related to the control

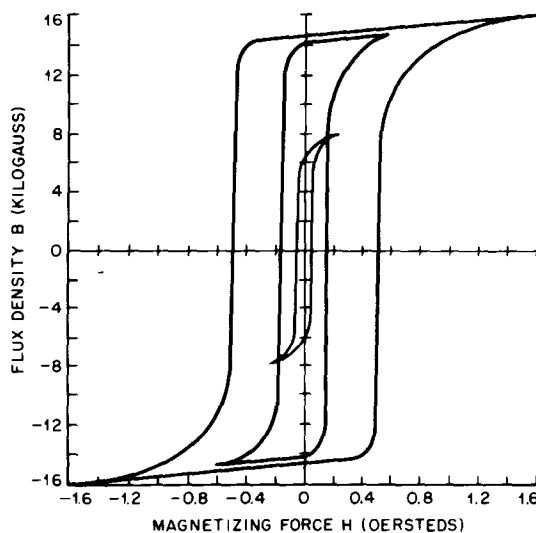


Fig. 2—Comparison of dc loops of three materials listed in Table 1.

TABLE 1—MATERIALS AND APPLICATIONS.

Material Letter Code	Trade Name	Cost/Core* Relative to Material A	Principal Use
A	Square Orthonol, Hipernik V, Orthonik, 49 Square Mu, Deltamax	1.00	High-gain amplifiers, oscillators, integrators, timers, memory devices.
N	Round Orthonol	1.00	Amplifier applications with slightly less gain, but less liability to triggering instability, than material A.
H	48 Alloy, Carpenter 49	1.00	Material A and N applications with lower sensitivity, lower losses, and less triggering instability. High-quality transformers.
D	Square Permalloy 80, Square Mu 79, Super Square Mu 79, Hy Ra 80, 4-79 Permalloy, Square Permalloy	1.14	High-gain amplifiers at low signal levels and low losses, low-power-consumption inverters and converters.
R	Round Permalloy 80, Hy Mu 80	1.14	High-quality low-loss inductors and transformers.
F	Supermalloy	1.63	Material D and R applications in which the lowest possible exciting currents and losses are required.
K	Magnesil, Selectron, Microsil, Hypersil, Supersil	0.70	Power amplifiers requiring lower gain and lower cost than material A applications. High-quality power transformers.
S	Supermendur	3.4	Material A and K applications where minimum size and weight and maximum operating temperatures are required.

\* Based on 2-mil tape-wound core of about 3-inch diameter.

signal. Further, note that at the end of the half-cycle,  $\phi$  is at saturation. Thus the core must be reset to  $\phi_0$  in the second half-cycle. If a similar voltage is to be applied to the load in the second half-cycle, a second core must be included in the

circuit. Such resetting and output problems are responsible for the variety of amplifier circuit configurations that have been used.

### AMPLIFIER CONFIGURATIONS

The amplifier configuration is arranged with two considerations in mind. One is the method of control, and the other is the type of output desired. As seen from the discussion above, a core that is brought to saturation and is gating power to a load on a positive half-cycle must be reset to its initial state if it is to repeat this function on the next positive half-cycle. On the other hand it is usually desirable to deliver power to the load on both half-cycles. Thus a second core will be gating power to the load during the half-cycle in which

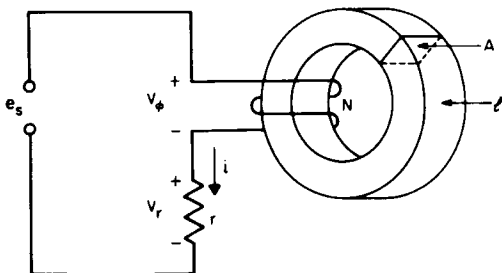


Fig. 3—Saturable reactor in series with a resistor.

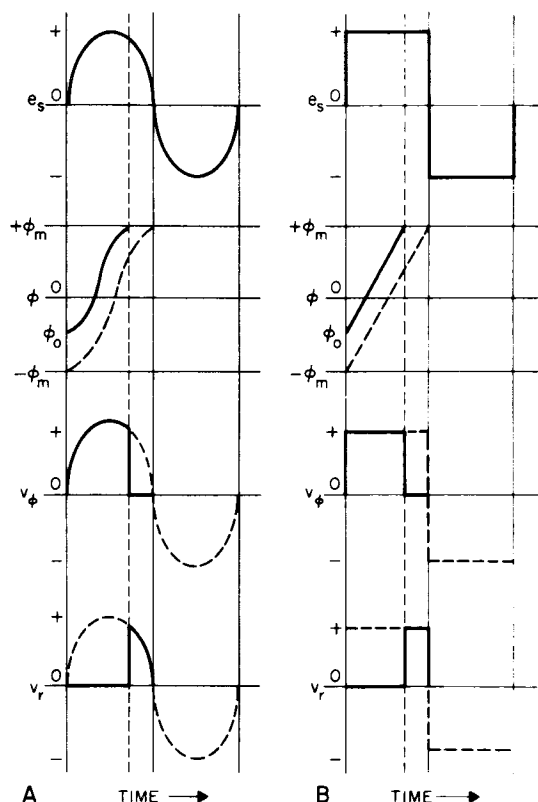


Fig. 4—Voltage and flux waveforms for the circuit of Fig. 3 for (A) sine-wave and (B) square-wave excitation.

the first core is being reset. Since the increment in flux linkage in the two cores is equal in the steady state, the core driven from the power supply can be used to reset the second core by transformer action through the control circuit. In single-phase circuits the roles of the cores interchange during alternate half-cycles. The use of one core to reset the other is fundamental to most amplifier configurations.

Several of the most common configurations are shown in Fig. 6. Figure 6A is the series-connected amplifier, sometimes called the transductor. It has been extensively analyzed [3, 9, 10] and is commonly used to measure large direct currents in electrochemical and power applications [11]. The details of the circuit operation are complicated but, in essence, at most one core is saturated at a time, gating power to the load. During this interval, the second (unsaturated) core, which has a very small coercive force, must therefore maintain the condition  $N_L i_L - N_c i_c = F_c$ , the dynamic coercive force of the unsaturated core. The control current  $i_c$  has the same waveform as the load current  $i_L$ .

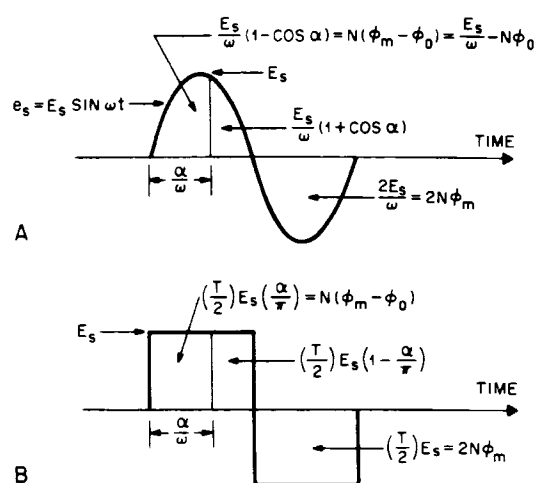


Fig. 5—Components of integrals of (A) sine-wave and (B) square-wave half-cycles.

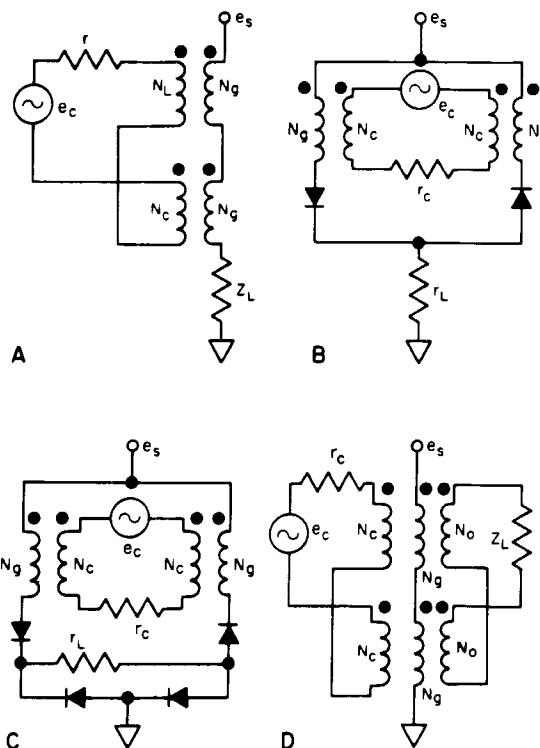


Fig. 6—Circuits for (A) the series-connected amplifier or transductor, (B) self-saturating amplifier with ac output, (C) self-saturating amplifier with dc output, and (D) second-harmonic modulator.  $N_c$  = control turns,  $N_g$  = gating turns, and  $N_o$  = output turns.

but has the same polarity on each half-cycle. Thus, the rectified average values of the load current and the control current,  $I_L$  and  $I_c$ , are related by the turns ratio.

$$I_L = (N_c/N_L)I_c + (1/N_L)F_c$$

a linear function with a constant offset as shown in Fig. 7A. In practice, the linearity of this function can be kept within about 0.1 percent, which makes it very useful for instrumentation.

The circuit in Fig. 6B is characterized by parallel-connected saturable reactors, so that the load current does not have to flow through an unsaturated core. Thus, since the resetting core is primarily transformer driven through the control circuit by the power-gating core, only the exciting current for the resetting core must be carried in the control circuit. In the steady state, the flux linkages coming from the winding on the gating core must equal the flux linkages delivered to the resetting core. With zero control voltage, the two flux linkages will differ by the integral of  $i_{cr}$  over the half-cycle. The function of the control voltage is to offset this voltage drop to make the two flux linkages equal at the desired output level. The diodes decouple the cores from the power supply during their resetting half-cycles.

When the amplifier is delivering full output, there is essentially no flux excursion in the gating core. It therefore does not drive the resetting core. Since the resetting core must not be reset under these conditions, the control current must be just below the coercive direct current for the core. At zero output, the gating core drives the resetting core at normal power voltage, resulting in a control current equal to the normal power-frequency coercive current. Full control of the amplifier is obtained over a control-current range equal to the widening of the hysteresis loop from dc to the power-frequency loop, divided of course by the control-winding turns. The resulting control characteristic is shown in Fig. 7B, with reference to Fig. 1. Again, there is a minimum output corresponding to the exciting current for the gating core. Also, the control current and voltage are automatically rectified because of the half-cycle symmetry of the circuit as seen from the control terminals. The modification shown in Fig. 6C delivers dc to the load.

A fourth configuration (Fig. 6D) used for very-small-signal amplifiers and magnetometers [12, 13] takes advantage of transformer coupling of the output to eliminate the residual exciting-current component found in the other circuits. The fundamental component of the induced voltage is canceled out and, at input currents other than zero, there is a second-harmonic component in the output proportional to the input current. The

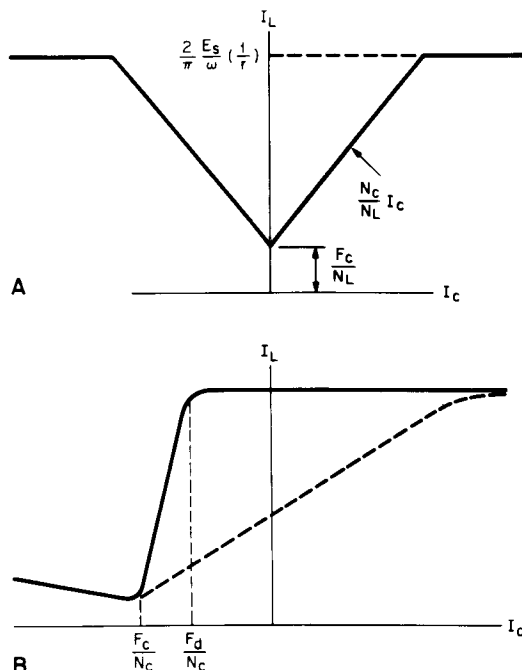


Fig. 7—Control characteristics for (A) the amplifier of Fig. 6A; and (B) for the amplifier of Fig. 6B and 6C. The dotted characteristic corresponds to nonsquare hysteresis loop, resistively shunted rectifiers, or negative feedback.

phase of this output reverses with the input-current polarity. Other examples of high-sensitivity amplifiers can be found in the literature [14]. For many power applications, 3-phase amplifiers are preferred for the usual reasons. The output is much easier to filter when dc is required [15].

Examples of parametric amplifiers and microwave magnetic amplifiers can be found in the literature.

## MULTISTAGE AMPLIFIERS

Because of the bilateral characteristics of magnetic amplifiers, multistage amplifiers are difficult to analyze. In addition, their properties as measured experimentally are technically unattractive and they are difficult to design even by empirical techniques except when the coupling-circuit impedance isolates the circuits. In this case, gain and frequency response are usually sacrificed. With the advent of a highly developed transistor technology, it is rarely necessary to design multistage amplifiers. Where it is necessary, the problem is treated as though the first stage were driving any other  $R$ - $L$  or  $R$ - $L$ - $C$  isolated dc load. If this

approach does not yield adequate performance, the designer must resort to less well documented techniques or initiate a new approach to solving the problem.

## BIAS AND FEEDBACK

Bias and feedback windings look and act exactly like the control windings shown in Fig. 6. Problems in design arise because these windings couple induced voltages from the cores into the bias and feedback circuits. Thus, the circuits are not simply unilateral and passive [16]. If these circuits are isolated by suitable  $R$ - $L$  networks, then the currents can be treated as though they were additive. In that case, the bias winding simply translates the origin of the control current in the direction of the bias.

The output current or voltage can be rectified and passed through simple  $R$ ,  $R$ - $L$ , or  $R$ - $L$ - $C$  networks and the derived current put into a feedback winding. In such a case, feedback is treated as it is in any other amplifier. It can linearize the amplifier successfully if the range of output is lowered. If the amplifier is used as a power device and its range of output is fixed by performance specifications, feedback is of no help unless compensating non-linearity can be inserted in other amplifying stages. Feedback in such cases is useful primarily to lower output or impedance of the amplifier.

## FREQUENCY RESPONSE

Analysis has shown that the voltage-gain bandwidth product [17] is

$$G_v \times (bw) = 4f/N \quad (\text{series-connected amplifier, Fig. 6A})$$

$$G_v \times (bw) = 2f/N \quad (\text{self-saturating amplifier, Fig. 6B})$$

where  $f$  is the frequency and  $N$  is the turns ratio  $N_c/N_g$ . These relations show again the advantage of high-frequency power supplies. They also suggest that if a design is adjusted to increase gain, the bandwidth (frequency response) will be reduced unless the turns ratios are adjusted to raise the gain-bandwidth product. High-gain amplifiers can be expected to have a bandwidth of about one-tenth the carrier frequency, which is frequently sufficient in instrumentation applications. In signal-processing applications, it is preferable to use several stages of low-gain wide-band amplification since the gains multiply and the bandwidths

go down more or less additively. It is in this area that magnetic amplifiers have been largely replaced by semiconductor circuits.

## CHOICE OF CORE MATERIALS

A variety of core materials can be chosen for magnetic amplifiers. They can be obtained in the form of tape-wound cores, laminations, and encapsulated tape cores cut into two mating  $C$ -shaped pieces. The latter configurations permit the use of simple inexpensive lathe-wound windings which can be assembled on the core. The cut- $C$  core configuration maintains the rolling direction of the tape along the primary magnetic path. The laminations are stamped out of continuous strip such that part of the magnetic path is along the direction of rolling and part is perpendicular to it. Since most good-quality strip is anisotropic, the resulting characteristics of the cores are not as good as they would be in the tape-wound configuration, which has the best magnetic qualities. As a result, only the lower-cost lower-quality magnetic materials are widely used in other than the tape-wound configuration. In addition to differences in processing and in the cost of raw materials, the above manufacturing considerations significantly contribute to the economic basis for choosing core materials. Cost is the dominant consideration in most amplifier designs. There are, however, extreme cases where only the highest-quality material can meet the technical specifications.

In general, the large, heavy, power-control applications make use of the least expensive materials in the least expensive configuration. In one important sense, they are sometimes technically superior as well. First, because of low remanence of the core, using these materials in a self-saturating configuration (Fig. 6B, 6C) results in a control characteristic which crosses the control-current axis as shown dotted in Fig. 7B. This automatically biases the amplifier near the desired operating point. In addition, the lack of squareness also causes fairly slow switching at the firing time of the power-gating core. The result is much less noise than found in the more objectionable gas tubes, semiconductors, and square-loop core circuits.

Table 1 lists the core materials available in tape-wound cores from most of the core manufacturers, plus a guide to their principal applications. An approximate cost ratio is given for 2-mil tape in a core of about 3 inches in diameter. This indicates the economic advantage of using materials no better than necessary. In lamination form, the cost per pound of the material is lower by about a factor of 5.

TABLE 2—TECHNICAL PROPERTIES OF MATERIALS.

Material Letter Code	Flux Density (kilogauss)	Squareness ( $B_r/B_m$ ) (400 Hz) ccfr*	Coercive Force (oersted)		Gain (kG/oe) (400 Hz) ccfr*	Curie Temp (°C)	Core Loss (mW/lb) (at +1 kG, 400 Hz)
			(dc)	(400 Hz) ccfr*			
A	14.2–15.8	0.94 up	0.1–0.2	0.45–0.65	310–715	500	56
N	14.0–15.6	0.85–0.95	0.07–0.17	0.10–0.20	260–500	500	42
H	11.5–14.0	0.80–0.92	0.05–0.15	0.08–0.15	280–550	500	19
D	6.6–8.2	0.80 up	0.02–0.04	0.022–0.044	550–1650	460	6.5
R	6.6–8.2	0.45–0.75	0.008–0.02	0.008–0.026	250–715	460	4.5
F	6.5–8.2	0.40–0.70	0.003–0.008	0.004–0.015	250–715	460	3.7
K	15.0–18.0	0.85 up	0.40–0.60	0.45–0.65	130–220	750	42
S	19.0–22.0	0.90 up	0.15–0.35	0.50–0.70	85–135	940	230

\* The values are typical of cores with ID/OD ratio of about 0.80 and tape thickness of 0.002 in. Tests made are in accordance with AIEE Standards paper II-432. (ccfr = constant-current-flux reset.)

Table 2 summarizes the technical properties of these materials. Note that many of their properties are given in terms of the IEEE standard referenced in AIEE Standards paper II-432 [18]. The use of these standards in circuit design and component specification is highly recommended.

The control-current range for self-saturating amplifiers, as indicated in Fig. 7B, can be estimated by referring to the difference between the dc and 400-hertz coercive-force columns in Table 2. This value must be multiplied by about 0.8 times the mean magnetic path length of the core in centimeters to obtain control ampere-turns. The values are for 400 hertz and must be corrected experimentally for other frequencies. Studies of the properties of materials and their influence on circuits covering a wide range of frequencies, temperatures, and materials can be found in the published literature [19, 20].

For more-specific and detailed design information, the designer should use the referenced literature. Also, several core-materials manufacturers have prepared excellent booklets containing all the essential tables and nomograms for designing magnetic-core circuits.

linearly with cross-section. There is also a linear relation between the exciting current and the mean magnetic path length for a fixed  $H$ . Thus, the exciting current is proportional to the volume of the core, as indicated by the energy dissipated in the material.

As the frequency rises, it is possible to use a smaller core for a fixed voltage. Comparing a 400-hertz design with a 60-hertz design, for example, the cores in the 400-hertz unit would be smaller by about a factor of 7. Since this is true for transformers and inductors as well, high-frequency power supplies are commonly found on aircraft where space and weight are important. The higher supply frequency also puts the carrier farther above the modulation-signal frequency spectrum, making it easier to recover the signal.

Since in many small-signal applications it is not necessary to have a large supply voltage, it is common to change available dc signals to square-wave ac voltages in the range from 5 to 25 kHz and higher. This means very small cores and very compact, sensitive amplifiers, a combination that often yields better performance in low-noise low-signal applications than semiconductor circuits.

## OTHER DESIGN CONSIDERATIONS

The basic design calculations, as discussed above, pick the core size and number of turns to fit the frequency and voltage. For a given magnetic material, a larger core requires fewer turns to support a given voltage at a given frequency. The number of turns varies as the inverse of the cross-section. From this fact alone, exciting current rises

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