Technical Bulletin BULLETIN SR-4

Mag Amp Cores and Materials

The best choice for tightly regulated outputs in switching power supplies

Proper regulation is an important consideration in specifying and designing switching power supplies. In multiple output power supplies where individual outputs must be tightly controlled, the design can be complicated by such things as additional circuits, heat sinks, larger size, etc.

The continuing need for more compact and reliable switching power supplies has aroused a renewed interest in a well founded control technique — the Magnetic Amplifier. Mag amps mean higher power density, simple control circuitry, very good regulation, high running frequency and rugged performance.

This bulletin describes mag amp regulation in switching power supplies. Three core materials are recommended for this application: 1 mil Permalloy 80, 1/2 mil Permalloy 80, and cobalt based amorphous material.

Many popular physical core sizes are listed here. These cores have been derived from our extensive selection of tape wound cores to encourage informal standardization, allowing unprecedented economy of manufacture. They are suitable for controlling individual outputs ranging from a few watts to well over 100 watts in power converters whose frequencies range from 20 kHz to several hundred kHz. Contact the factory for other sizes which may be needed for this application.

Advantages of Mag Amp Control:

- **Smaller size**
- **High reliability**
- **Generally less E.M.I**
- **Higher efficiency**
- **Simpler circuits**
- **Fewer components**
- **Less costly for outputs over 2 amperes**

Permalloy 80

Permalloy 80 is an ideal material for SMPS applications. Squareness is almost as high as the cobalt based amorphous material and flux density is higher. Core losses for 1 mil are much higher, and the 1/2 mil core losses are slightly higher than the cobalt based amorphous material. Permalloy cores offer the advantage of lower core cost.

Cobalt Based Amorphous Material

A cobalt based alloy, has low losses, very high permeability, high squareness and low coercive force. These characteristics make the alloy most ideal for SMPS applications such as magnetic amplifiers, semi-conductor noise suppressors and high frequency transformers. It also finds use in high sensitivity matching transformers and ultra-sensitive current transformers.

The cobalt based alloy has near-zero magnetostriction, high corrosion resistance and a high insensitivity to mechanical stress. These properties make it also useful in magnetometer applications.

Which Material To Choose?

The selection of any of these materials depends on desired characteristics and/or economic trade-offs.

The final choice of material for a particular application depends on several considerations, but in general, the 1 mil permalloy is chosen for the lower-frequency applications (under 50 kHz) because of its lower cost. At higher frequencies, the 1/2 mil material is chosen for its lower core loss and higher squareness.

Amorphous material is intended for higher frequency applications which demand the lowest loss and highest squareness.

HIGH FREQUENCY MAG AMP CORES

These cores specifically designed for this application. (5D = ½ mil permalloy, 1D = 1 mil permalloy, 1E =1 mil Cobalt based amorphous material).

		DIMENSIONS											
		I.D.		O.D.		Ht.		Core loss (w)@50KHz, 2000 gauss (Max.)	m*l* cm	Ac cm2	Wa See Note 1	Core wt. grams	Wa Ac See Note 2
Part Number		Core	case (Min.)	core	case (Max)	core	case (Max)						
50B10-5D	in.	.650	.580	.900	.970	.125	.200	.118	6.18	.051	348,000	2.7	.0177
	mm	16.5	14.7	22.9	24.6	3.18	5.08				1.76		.0897
50B10-1D								.220		.076		4.0	.0264
													.1340
50B10-1E								.092		.076		3.5	.0264
													.1340
50B11-5D	in	.500	.430	.625	.695	.125	.200	.044	4.49	.025	194,000	1.0	.0048
	mm.	12.7	10.9	15.9	17.6	3.18	5.08				.984		.0243
50B11-1D								.083		.038		1.5	.0074
													.0375
50B11-1E								0.34		0.38		1.3	.0074
													.0375
50B12-5D	in	.375	.305	.500	.570	.125	.200	.035	3.49	0.25	99,000	.8	.0025
	mm.	9.53	7.75	12.7	14.5	3.18	5.08				.50		.0127
50B12-1D								.066		.38		1.2	.0038
													.0193
50B12-1E								.027		.038		1.04	.0038
													.0193
20B45-5D	in.	.500	.430	.750	.820	.250	.325	.194	4.99	.101	194,000	4.4	.0143
	mm.	12.7	10.9	19.1	20.8	6.35	8.26				.984		.0725
50B45-1D								.363		.151		6.6	.0214
													.1080
50B45-1E								.149		.151		5.7	.0214
													1080
50B66-5D	in.	.500	.430	.750	.820	.125	.200	.097	4.99	.050	194,000	2.2	.0071
	mm.	12.7	10.9	19.1	20.8	3.18	5.08				.984		.0360
50B66-1D								.182		.076		3.3	.0108
													.0548
50B66-1E								.075		.076		2.9	.0108
													.0548

(1) Top no.= circ. Mils.
Bottom no =cm^2

(2) Top no.= circ. Mils. X^{2}cm x10^6
Bottom no= cm^4

Above “50000” series cores are provided in nylon boxes. “1E” cores can be supplied in “54000” series (encapsulated, no box). Dimensions of the 54000 series cores are as shown at right.

Additional “1E” encapsulated cores are listed on p. 3

CORE		I.D (min.)	O.D (max.)	Ht (max.)	Wa (seeNote 1
54B10	in.	.0610	.0940	.175	372,000
	mm.	15.5	23.9	4.45	1.89
54B11	in.	.460	.665	.175	211,600
	mm.	11.7	.16.9	4.45	1.07
54B12	In	.335	.540	.175	112,225
	mm.	8.51	13.7	4.45	.569
54B45	in.	.460	.790	.300	211,600
	Mm	11.7	20.1	7.62	1.07
54B66	in.	.460	.790	.175	211,600
	mm.	11.7	20.1	4.45	1.07

(1) (Top no.= circ. Mils
Bottom no. = cm2

COBALT BASED AMORPHOUS MATERIAL
HIGH FREQUENCY MAG AMP CORES (encapsulated

Part Number		DIMENSIONS I.D. Core	I.D. coated (Min.)	O.D. core	O.D. coated (Max)	Ht. core	Ht. coated (Max)	Core loss (w)@50KHz, 2000 gauss (Max.)	m*l* cm	Ac cm2	Wa See Note 1	Core wt. grams	Wa Ac See Note 2
54C90-1E	in.	.312	.272	.500	.540	.188	238	.055	3.24	.085	97,000	2.09	.008
	mm	7.92	6.91	12.7	13.7	4.77	6.05				.491		.041
54C70-1E	in.	.375	.335	.500	.540	.188	.238	.040	3.49	.057	141,000	1.51	.008
	mm.	9.53	8.51	12.7	13.7	4.77	6.05				.715		.041
54D26-1E	in.	.375	.335	.547	.587	.188	.238	.061	3.67	.083	141,000	2.31	012
	mm.	9.53	8.51	13.9	14.9	4.77	6.05				.715		.061
54D27-1E	in.	.375	.335	.594	.634	.188	.238	.085	3.87	.110	141,000	3.23	.016
	mm.	9.53	8.51	15.1	16.1	4.77	6.05				.715		.081
54C91-1E	in.	.375	.335	.625	.665	.188	.238	.090	3.99	.113	141,000	3.42	.016
	mm.	9.53	8.51	15.9	16.9	4.77	6.05				.715		.081
54319-1E	in.	.375	.335	.625	.665	.250	.300	.119	3.99	.150	141,000	4.52	.021
	mm.	9.53	8.51	15.9	16.9	6.35	7.62				.715		1.06
54C88-1E	in.	.500	.460	.590	.630	.188	.238	.034	4.35	.040	250,000	1.32	.010
	mm.	12.7	11.7	15	16	4.77	6.05				1.27		.117
54942-1E	in.	.500	.460	.700	.740	.188	.238	.873	4.79	.091	250,000	3.30	.23
	mm	12.7	11.7	17.8	18.8	4.77	6.05				1.27		.117
54632-1E	in.	.500	.460	.750	.790	.188	.238	.113	4.99	.113	250,000	4.27	.28
	mm.	12.7	11.7	19.1	20	4.77	6.05				1.27		.142
54904-1E	in.	.500	.460	.750	.790	.312	.362	.188	4.99	.188	250,000	7.11	.047
	mm.	12.7	11.7	19.1	20	9.19	7.92				1.27		.239
54C89-1E	in.	.550	.510	.825	.865	.188	.238	.137	.5.48	1.25	.303,000	5.19	.038
	mm.	13.97	12.9	21	22	4.77	6.05				1.54		1.93
54094-1E	in.	.625	.585	1,000	1,040	.375	.425	.440	6.48	.339	391,000	16.64	.133
	mm.	15.88	14.9	25.4	26.4	9.53	10.8				1.98		.674
54C92-1E	in.	.688	.648	.875	.915	.188	.238	.106	6.23	.085	473,000	4.01	.040
	mm.	17.48	16.5	22.23	23.2	4.77	6.05				2.40		.203
54168-1E	in.	.750	.710	1,000	1,040	375	.425	.316	6.98	.226	563,000	11.95	.127
	mm.	19.05	.18	25.4	26.4	9.53	10.8				2.85		.644
54C17-1E	in	.800	.760	1.205	1.245	.375	.425	.586	8.00	.366	640,000	22.18	.234
	mm.	20.32	19.3	30.61	31.6	9.53	10.8				3.24		1.19
54029-1E	in.	1,000	.960	1,375	1,415	.250	.300	.505	9.47	.226	1,000,000	19.08	.226
	mm.	25.4	24.4	34.93	35.9	6.35	7.62				5.07		1.15
54932-1E	in.	1,000	.960	1,625	1,665	.625	.675	1.98	10.47	.942	1,000,000	74.71	.942
	mm.	25.4	24.4	41.28	42.3	15.88	17.1				5.07		4.78

(1) Top no.= circ. Mils.
Bottom no =cm^2

(2) Top no.= circ. Mils. X cm^2 x10^6
Bottom no= cm^4

PART NUMBER CODE

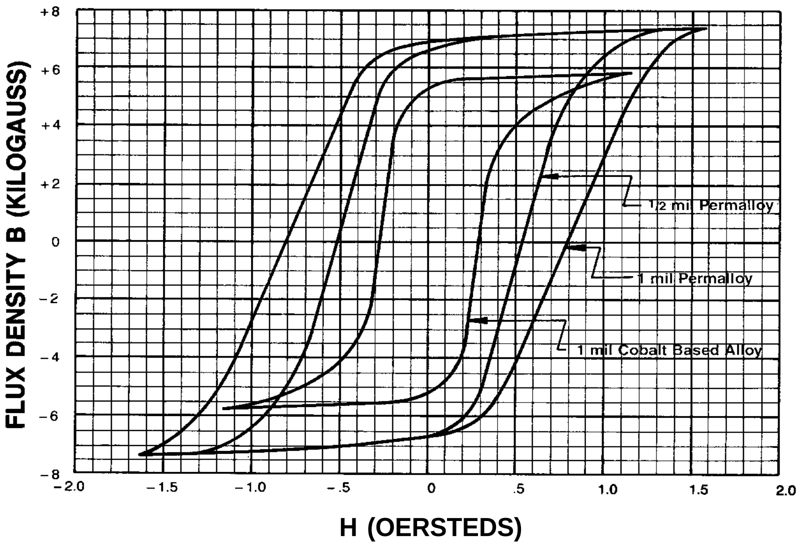
50	B10	1 or 5	D or E
Nylon Case	size Code	1-1mil 5-1/2 mil	D-permalloy E-Cobalt Based Amorphous Material

MATERIAL CHARACTERISTICS

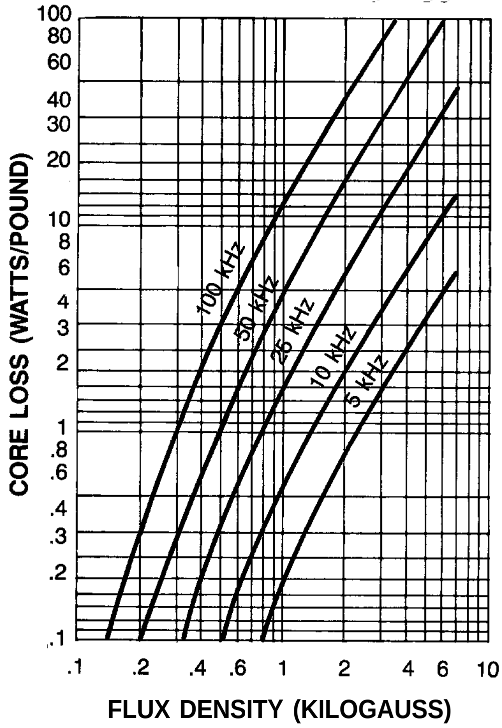
	Alloy 2714A	½ mil Permalloy	1 mil Permalloy
Bm (gauss min.)	5000	7000	7000
Br/Bm (min)**	.9	.83	.80
H1 (oersted max.)**	.025	.045	.045
Core loss(w/lb. Max. @ 50 kHz, 2000 gauss)	12	20	25

** Measured @ 400 Hz, CCFR Test

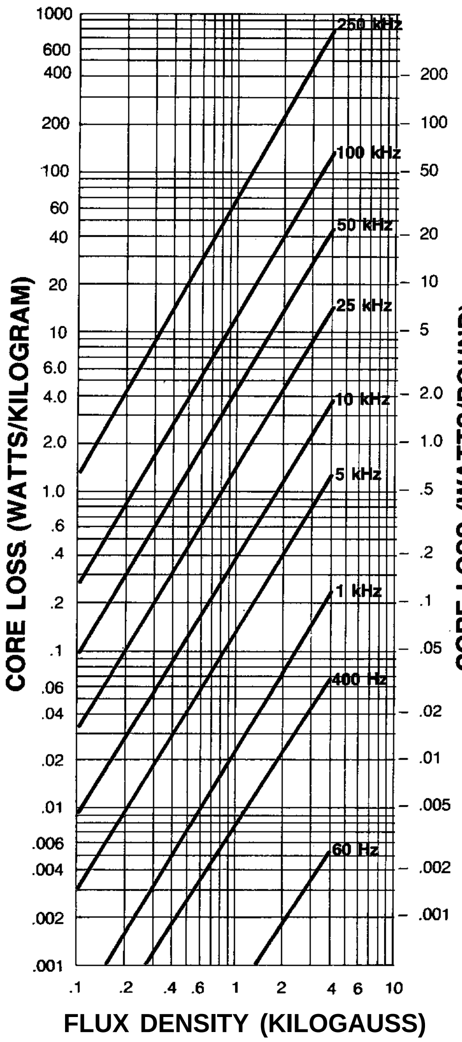
100 kHz B—H LOOPS PERMALLOY AND COBALT BASED AMORPHOUS MATERIAL



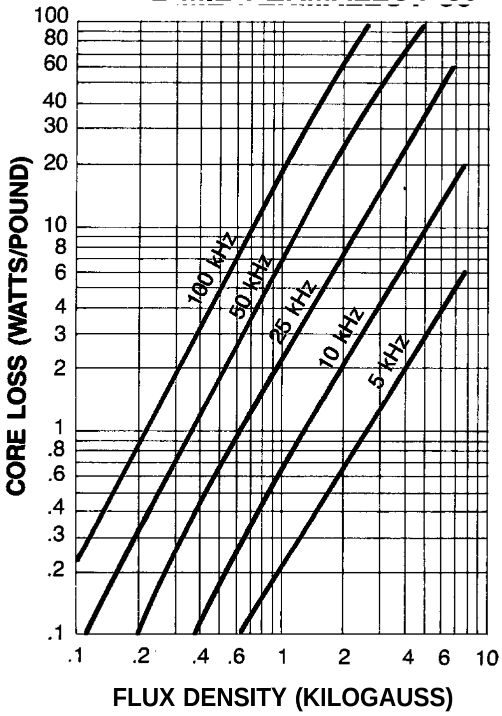
TYPICAL CORE LOSS — ½ MIL PERMALLOY 80



TYPICAL CORE LOSS — AMORPHOUS MATERIAL-COBALT BASED



TYPICAL CORE LOSS — 1 MIL PERMALLOY 80



Output Regulators for Switch-Mode Power Converters

A popular and effective application of the Square Permalloy 80 tape wound cores occurs in multiple-output switched-mode power supplies. By using such a square-loop core to provide a controllable delay at the leading edge of the pulses at the secondary of the transformer, one or more outputs can be independently and precisely regulated without the losses inherent in linear regulators or the complexity of conventional switching regulators. In cases where the load currents of the subordinate outputs are high (in excess of one or two amps), the advantages of the saturable-core regulators become more and more significant. Figure 1 shows the block diagram of a typical multi-output supply of this type, while Figure 2 illustrates the regulation scheme.

For simplicity of this example, a forward converter topology is shown, but the technique is equally useful in flyback and push-pull converters. Typical waveforms are shown in Figure 2. In the pulse width modulator (PWM), the primary pulse width is controlled by sensing the 5V output, comparing it to a reference, and using the error signal to adjust the pulse duration. If there were no saturable core (SC) in the circuit, the 15V output would be "semi-regulated," since the primary control loop would provide line regulation. But the output would vary with load and temperature.

To produce a 15V dc at the output, the average value of the rectified waveform applied to the input inductor L must be 15V. Given the pulse height of 50V and a repetition period of 10 µs, the required width of the positive pulse at e_2 must be:

$$PW = (15V/50V) \bullet 10\mu s = 3\mu s$$

Because the input pulse (e_1) is 4 µs wide, the saturable core must delay the leading edge by 1 µs. Since the amplitude of the pulse is 50V, we can say that the core must "withstand" 50V • 1 µs, or 50 volt-microseconds. To accomplish this, the core is reset by this amount during each alternate half cycle. The waveform at e_2 illustrates this. As the input to the core swings negative, diode CR1 conducts and allows the error amplifier A1 to "clamp" the output side of the core at -37.5V. The result is that the core is subjected to a reverse voltage of 50 - 37.5V for a duration of 4 µs, producing a reset of:

$$\Lambda = 12.5V \bullet 4\mu s = 50V \bullet \mu s$$
$$(\Lambda = \text{Withstand})$$

As the output varies, the error amplifier will alter this value to ensure that the output is regulated at 15V dc in spite of changes in the rectifier voltage drops, etc.

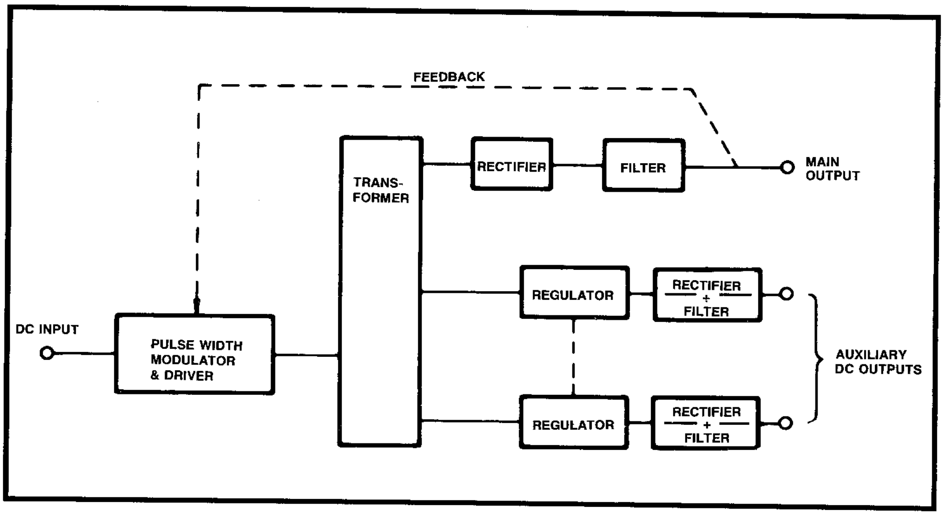


Figure 1 *Multiple-output switched-mode power supply*

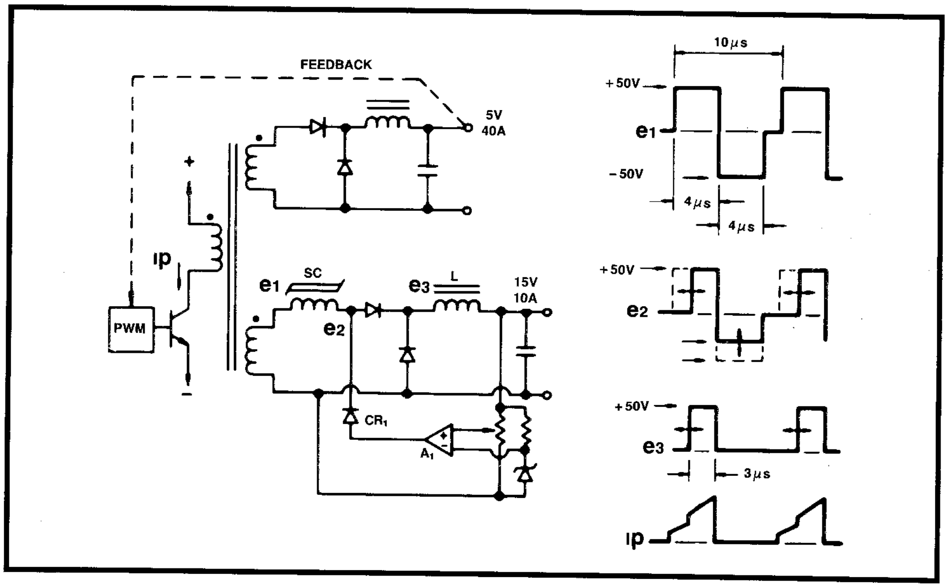


Figure 2 *Regulation Scheme.*

The waveform of the primary current, i_p, shows the increase in current when the core saturates and begins to deliver current to the output inductor. This has an incidental bonus: the primary switching transistor has already turned on and saturated and hence the 15V output does not contribute to turn-on switching losses in the transistor.

The design of the saturable reactor requires three steps:

1. Determine Λ,the withstand volt-seconds to delay the leading edge of the pulse and achieve the required output voltage. Here, the designer must decide whether the output must be capable of independent "shutdown" (for short-circuit protection or turn-off from and external logic signal), or simply regulated at a fixed value.

Withstand = Excluded Pulse Area = Λ

$$\Lambda = V \bullet t$$

Where V = pulse amplitude, and

t = delay at leading edge.

Case 1 — Shutdown. The required withstand is simply the area under the entire positive input pulse. In the circuit of Figure 2, it would be 50V • 4µs = 200 Volt-microseconds.

Case 2 — Regulation only. Assuming that the output inductor has been designed for continuous conduction, the reactor must only reduce the input pulse width enough to furnish the required average value (equal to the dc output voltage) at the input of the filter inductor.

In both cases, one must allow "headroom" to accommodate load transients. This comment relates to the choice of turns on the secondary winding of the transformer which feeds the regulator, which must precede the calculation of the volt-seconds which the reactor must support. For example, one might design for control range of ±20% to allow the pulse width to increase or decrease by 20% when the load current steps up or down. To allow the pulse width to increase, the input pulse width must be 20% greater than the nominal pulse at the output of the reactor. Depending on the operating frequency and core used, one must allow an additional margin due to the risetime of current in the core after it saturates. This is typically on the order of one microsecond. This implies that the secondary voltage be at least 20% higher than it would be to produce the desired output voltage if the saturable reactor were not present. To allow the pulse width to decrease, the reactor must withstand additional volt-seconds to reduce the pulse width 20% below the nominal value.

In the circuit of Figure 2, a "regulation only" design would require a withstand of Λ = 50V • 1µs + 20%, or 60V • µs.

2. Choose the core. There are two popular methods of determining the size of the required core. Each results in a minimum area product, WaAc, to provide the necessary withstand and accommodate the wire size (which determines the temperature rise). One method [1] begins with the desired temperature rise and power to be handled (withstood), the core geometry, and the fill factor. The other requires an initial choice of the wire size, which must be estimated based on intuition about the ultimate temperature rise. Although the latter is admittedly pragmatic, it is popular because of its simplicity.

In the latter method, the steps are as follows:

A. Pick the wire size, based on the current. A reasonable value is 500 circular mils per amp of current (rms) for a temperature rise of 30 to 40 degrees C in core sizes of .5 to 1 inch o.d. This yields Aw, the cross-sectional area of one conductor.

B. Choose a core material to determine the saturation flux density, Bm. In this application, Square Permalloy 80 is a good choice, since it has low coercive forces and a very square BH loop. Its Bm is approximately 7000 gauss.

C. Choose the fill factor, K, using .3 to .5, with the lower values for power applications.

D. Calculate WaAc as follows:

$$WaAc = \frac{Aw \bullet \Lambda \bullet 10^8}{2 \bullet Bm \bullet K}\text{, in circ. mils.} \bullet cm^2$$

E. Select a core from the selection tables here or on pp. 58-61 of Catalog TWC-400 with at least this area product. In doing so, the tape thickness must be chosen, and the values in the WaAc column (pp. 58-61) must be modified according to Note 3 at the bottom. Tape thicknesses of .0005 and .001 inch are recommended for frequencies up to 100 kHz, with the thinner tapes found in the bobbin-wound core catalog preferred at higher frequencies.

In the circuit of Figure 2, the current during conduction of the core is 10A, and the duty ratio is 15/50, or .3. Thus the current is $(10^2 \bullet .3)^{1/2}$, or 5.5A. An appropriate wire size is 16 gauge, since its cross-sectional area, Aw, is 2581 c.m. Again, using the "regulation only" case, WaAc is as follows:

$$WaAc = \frac{2581 \bullet 60 \bullet 10^{-6} \bullet 10^8}{2 \bullet 7000 \bullet .1}$$

$$= .011 \bullet 10^6 \text{ c.m. } cm^2$$

Note that fill factor of .1 has been used, since the wire size is relatively large.

Since the converter frequency is 100 kHz, the tape thickness of .0005" is perhaps a wise choice. In consulting the table on page 58, the WaAc figures must be altered by a factor of approximately .013/.022 (the typical ratio of the cross sectional areas of cores with .0005" and .002" tape thickness), ac-

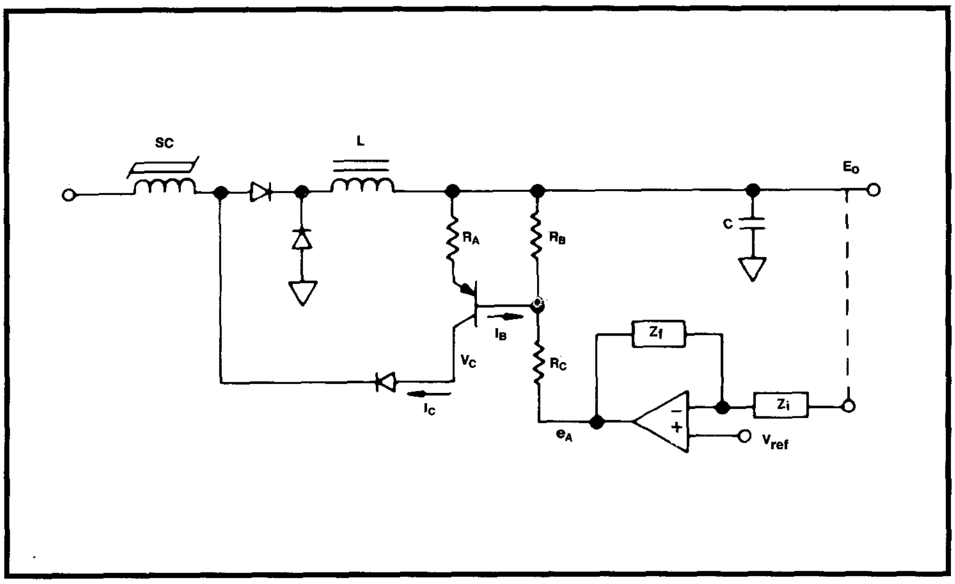


Figure 3 *Alternative control circuit*

cording to Note 3 at the bottom of the page. The most convenient way to do this is to alter the value of the desired WaAc, and then find the appropriate core in the table. Using this approach, the listed value must be at least .011• (.022/.013) • 10^6, or .019 • 10^6. Two logical candidates are the 5_374 and 5__063 cores, whose WaAc (x 10^6) values are .028 and .026, respectively.

For the purpose of this example the 5_063 core is chosen. Its effective core cross sectional area, Ac, is ,050 cm²; its mean length of magnetic path, Ml, is 5.98 cm. These values are noted for future use.

3. Determine the number of turns. The number of turns is determined by the withstand, Λ, to produce the desired output of the regulator:

$$N = \frac{\Lambda \bullet 10^8}{2 \bullet Bm \bullet Ac} \quad \text{turns}$$

Where: Λ = withstand, in volt-seconds

Bm = Saturation flux density in gausses

Ac = Core cross sectional area in cm².

The control circuit can now be designed. In doing so, it is helpful to estimate the current required to reset the core and thus calculate the average control current based on the duty ratio of the resetting (negative portion) of the input pulse. The current is related to the magnetizing force as follows:

$$Im = \frac{.794 \bullet H \bullet Ml}{N} \quad \text{amps}$$

Where: H = Magnetizing force in Oersteds

Ml = Magnetic path length in cm.

H is not simply the dc coercive force, but rather the value corresponding to the flux swing and frequency, as given in curves on pages 38 through 49 of Catalog TWC-400. Note the "loop widening effect" — the force increases with frequency.

Again, using the circuit of Figure 2 and the chosen core, the required number of turns is:

$$N = \frac{60 \bullet 10^{-6} \bullet 10^8}{2 \bullet 7000 \bullet .050} = 8.57 \text{ turns}$$

= (round off to 9 turns)

Completing the example, the magnetizing current is calculated as follows: Since the regulator will be required to swing across the entire BH loop during transcience, the curve on page 48 of Catalog TWC-400 will give a typical estimate of the magnetizing force. At a frequency of 100 kHz, the 1/2 mil curve has the value of H = .215 Oersteds. Thus, the magnetizing current will have a typical value of:

$$Im = \frac{.794 \bullet .215 \bullet 5.98}{9} = .11A$$

(See page 10 for additional information on calculating reset current)

An alternative control circuit is given in Figure 3. It has two notable features:

1. The resetting control circuit is derived from the output, providing a "preload" — a means of preventing the magnetizing current of the reactor from raising the output voltage at zero load.

2. The core is reset from a current source, rather than a voltage source. This has been shown by Middlebrook[2] to minimize the phase shift of the control transfer function. In this circuit, Ra degenerates the transconductance of the transistor, making the transfer function more independent of the transistor. Rb and Rc simply shift the level of the amplifier's output, which is unnecessary if the amplifier is powered from a voltage higher than the output.

3. The compensation networks, Zf and Zi, can be designed using techniques for conventional buck-derived regulators.

Note, however, that this circuit actually has two feedback loops—one through the error amplifier, and one directly from the output through Ra and the transistor.

Full-wave outputs can be handled in the same manner as the forward converter discussed earlier. The Circuit of Figure 4 illustrates this application.

Finally, it is sometimes useful to be able to translate the voltage required to reset the core, change its level, or trade voltage for current. In these cases, a second winding can be placed on the core, with a larger or smaller number of turns than the power-handling winding, and with its end opposite the control transistor being returned to a convenient bias voltage. For example, a control winding with fewer turns will exhibit less voltage swing but will require more control current than the main winding.

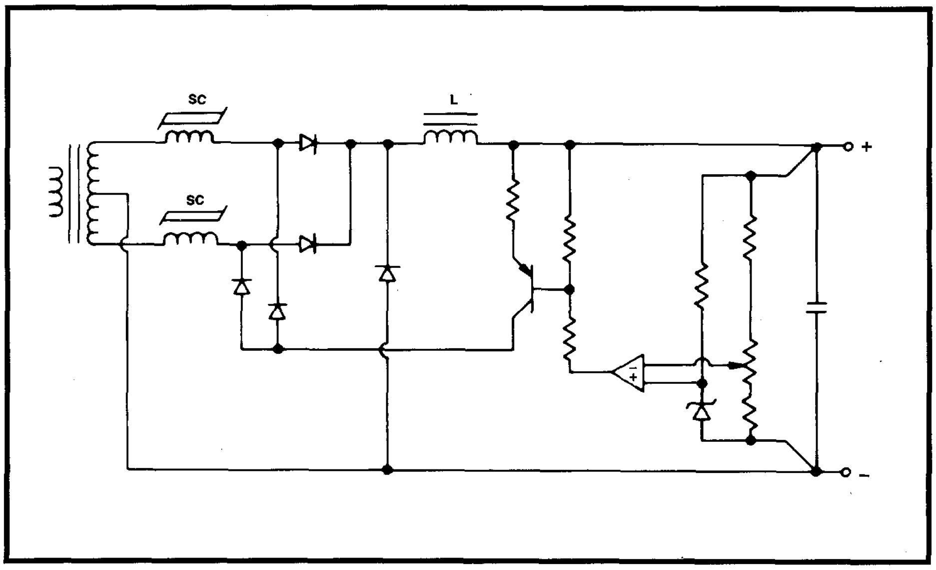


Figure 4 *Full-wave saturable core regulator.*

Current Mode Control

Another configuration of control circuitry is shown in Figure 5. It is equally useful for half-wave and full-wave applications, but is shown here in the half-wave case for simplicity. This circuit is particularly advantageous when independent current limiting (of the mag amp output) is desired. Unlike the current-limiting methods of the past, where the output of an overcurrent detector op amp or comparator was “ORed” with the error amplifier output, this method “embeds” the current-monitoring function in the feedback loop. Thus, it is always active and provides exceptionally smooth transitions as the output is loaded beyond the current limit and then returned to normal load conditions.

Amplifier U1B is the current error amplifier, whose input circuit is comprised of resistors R6, R7, an R12. The output of U1A is the most important input to this network, since it is a result of the output voltage error. The output voltage, Vo, is also introduced (via R12) for bias and to shape the current foldback characteristics, if desired. Resistor R8 samples the regulator’s output current, and resultant voltage is applied to the error amplifier through resistor R7. Resistors R4 and R5 determine the gain of the amplifier; gain = (R4 + R5)/R5.

To visualize the operation of this circuit, first assume that the output of U1A is stationary during a change in the current. An increase in current causes an increase in the voltage drop across R8. Since the regulator output is treated as the arbitrary ground reference, this increase in current is evidenced by a downward voltage swing at the junction of R7 and R8. This is amplified without inversion by U1B and applied to the mag amp reset transistor, Q1, through R3. The increase in reset current decreases the pulse width at the output of the mag amp and thus opposes the increase in current which was sensed by R8.

The voltage feedback loop begins with R9, the input resistor for the voltage error amplifier, U1A. Biasing resistor R10 is not part of the transient response analysis, since its voltage doesn’t change (the inverting input of U1A is a virtual ground and remains stationary). Resistor R11 and capacitor C1 form the feedback network of U1A, making it an integrator with a zero at the frequency where C1’s reactance equals R11. The output of U1A is then applied through R6 to the other amplifier, U1B, which amplifies it and applies it to the reset transistor. An increase in the output voltage, Vo, is inverted by amplifier U1A and ultimately increases the mag amp’s reset current supplied by Q1. This corrects the perturbation.

Diode CR4 limits the positive voltage swing at the output of U1A. Since U1’s output voltage is the “reference” for the current limiter, the clamping action of CR4 determines the maximum output current.

The design philosophy is to have the current-mode feedback determine the

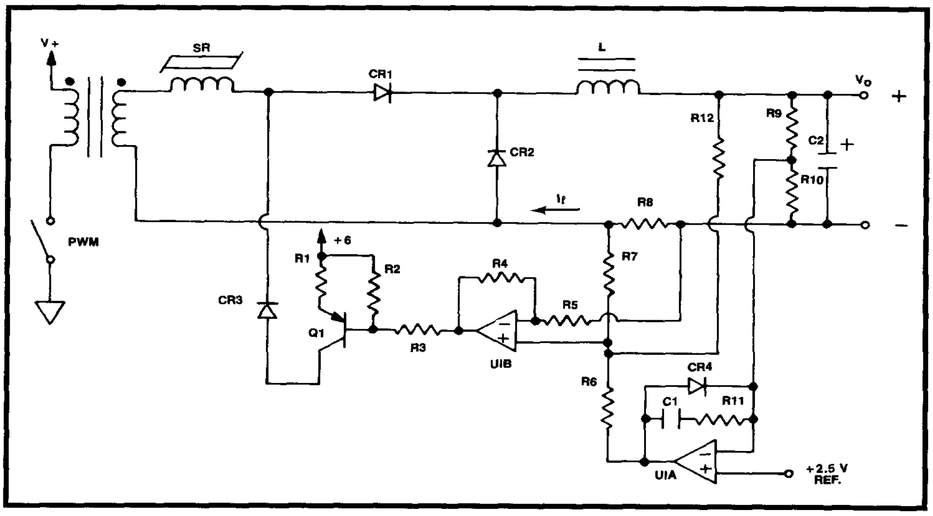


Figure 5 *Mag amp regulator with current mode control.*

phase shift at the unity-gain crossover frequency, since its maximum is 90 degrees. The two feedback paths combine to a vector sum, and thus the dominant one will determine the phase shift. It is recommended that the current-mode loop cross the unity-gain axis at around one-tenth the switching frequency, and the voltage-mode feedback cross over at least two or three octaves below the current-mode loop.

References 4 and 5 describe the design of mag amp output regulators in more detail.

Circuits using these square-loop cores have appeared in power converters operating at frequencies up to 1 MHz[3]. Not only can they perform output regulation, but also they can be used in the primary circuits to control the frequency of the converter. Applications are practically limitless in the hands of the designer with imagination and a firm concept of these interesting “volt-second” components.

Calculation of Coercive Force H_R Needed for Reset of Saturable Reactor Core

One of the questions frequently asked is how to calculate the coercive force H_R needed to reset the saturable core. The use of Figure 38 in MAGNETICS tape wound core catalog TWC-400 can give an estimate as suggested, but Clifford Jamerson in his presentation, "Calculation of Magnetic Amplifier Post Regulator Voltage Control Loop Parameters" (see #5 listed under "References" on page 11) discusses a more accurate method. Excerpts from that paper are presented here.

The use of Figure 38 above gives the values for H needed to saturate the core via a square wave. At 50 to 100kHz, the flux swing is limited by core loss. The coercive force required for reset is usually considerably less than Figure 38 suggests. In an attempt to find a more accurate procedure for estimating H_R, the author discovered a simple, yet reliable, method for calculating H_R from the loss curves supplied. The derivation of the method is shown below, together with the assumptions. The results are shown here.

To calculate H_R for any flux swing at any frequency, the formula for **square permalloy 80** is

$$H_R = \frac{1.2 \bullet 10^6 \bullet (\text{watts/lb from the loss curves})}{\Delta B \times \text{Switch Frequency}}$$

For **cobalt based amorphous material** the formula is:

$$H_R = \frac{1.05 \bullet 10^6 \bullet (\text{watt/lb from the loss curves})}{\Delta B \bullet \text{Switch Frequency}}$$

(In the above formulas,
B is in gauss
Frequency is in Hertz
H is in Oersteds)

Calculation of Reset Current

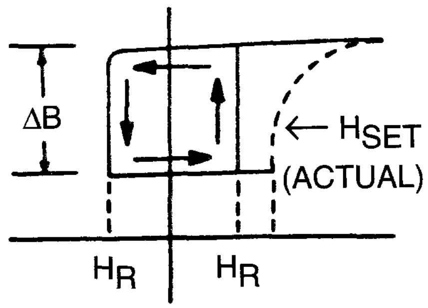
Once H_{reset} is found, the reset current can be calculated via Ampere's Law.

$$I_{reset} = \frac{H_{reset} \bullet l_e}{.4\, \pi\, N}$$

where le is the mean magnetic path length and N is the number of turns on the saturable reactor core.

Derivation of Formula for H_{reset}

Assume the ideal square BH loop shown below. Here H_{reset} is shown to be artificially equal to H_{set}, just to make the hysteresis loop symmetrical. However, validity of the formula does not depend upon Hset being equal to H_{reset}.



For the ideal square loop, the core loss per volume is

$\Delta B \bullet 2H_R \bullet \text{Frequency} = \text{watts/meter}^3$, ΔB in Tesla, H_R in AT/M. Solve for H_R.

$$H_R = \frac{\text{Watts/Meter}^3}{2\,\Delta B \bullet \text{Frequency}}$$

But since the loss curves are in watts per pound, this must be converted from watts/lb to watts/meter³.

$$H_R = \frac{\text{No. of Watts/lb} \bullet \text{lb/454 grams} \bullet 8.7\ \text{grams/cm}^3 \bullet 10^6\ \text{cm}^3\text{/meter}^3}{2\,\Delta B \text{ in gauss} \bullet 1\ \text{tesla}/10^4\ \text{gauss} \bullet \text{frequency}}$$

$$= \frac{95.8 \bullet \text{No. of watts/lb}}{\Delta B \bullet \text{Frequency}}$$

This gives H_R in Ampere-Turns/Meter. To convert to Oersteds, divide by 79.6.

Thus

$$H_R = \frac{1.2 \bullet 10^6 \bullet \text{No. of watts/lb}}{\Delta B \bullet \text{Switch Frequency}}$$

(for square Permalloy 80)

To find the H_R for another square loop material, the same derivation applies, but the density for the new material should be used in the above formula.

For example the density for cobalt based amorphous is 7.59 grams/cm³. The formula is thus adjusted to become:

$$H_R = \frac{1.05 \bullet 10^6 \bullet \text{No. of Watts/lb}}{\Delta B \bullet \text{Switch Frequency}}$$

(cobalt based amorphous material)

Assumptions:

1. The loop is square.
2. The H_R for reset of magamps with ΔB flux swing is the same as for a symmetrical ΔB swing about the origin.
3. H_R is the same for a 30 to 40% duty cycle constant-current reset as for 1/2 of a sine wave.

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