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A TRANSFORMER OF CLOSELY SPACED PULSED WAVEFORMS HAVING A NONZERO AVERAGE VALUE

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16. Abstract

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SUMMARY

With the use of diodes, transistors, and magnetic cores, a passive circuit was constructed that transformed the voltage, either up or down, of repeated positive or negative pulses. This circuit combined a pulse transformer with switching devices to effect a resonant flux reset. It could transform various pulsed waveforms that had a nonzero average value and were relatively closely spaced in time. The circuit provided direct-current isolation between multiple loads and between a load and the source, similar to a conventional transformer. Several experimental circuits were tested.

INTRODUCTION

On occasion a need arises to drive a load, or multiple loads such as transistor switches in an inverter circuit configuration, from a source of repetitive positive or negative voltage pulses with the requirement of direct-current isolation from load to source and load to load. This isolation is usually provided by an ordinary pulse transformer. In such a transformer, the pulse spacing is limited by the ability of the flux in the core to return to its initial value before the arrival of the next pulse. Especially with low impedance sources and loads, the needed flux reset time is normally significantly greater than the pulse width. In the case of relatively closely spaced repetitive positive or negative pulses, an ordinary transformer will be driven toward magnetic saturation and, therefore, cannot be used.

A new voltage-transforming circuit was constructed to handle repetitive positive or negative pulses that may be relatively closely spaced in time and have a nonzero average value. The only power source of the circuit is the pulse source; its output has no direct-current voltage offset, and it provides the isolation of a transformer. The principles of operation and limitations of this circuit are given in the following sections, along with a discussion of several experimental versions that were tested.

BASIC OPERATION OF REPETITIVE PULSE TRANSFORMER

The transformer circuit to be described is in its simplest form a passive four-terminal network The circuit is passive in the sense that all of its output energy comes from the input voltage source. Its input voltage v_i and output voltage v_o are functions of time ideally related by $v_o = nv_i$, where the constant n is the turns ratio of the transformer. (All symbols are defined in appendix A.) Further, the circuit transforms the average value of the input. That is, if the average value of the input is $\overline{v}_i$, then the average value of the output is $\overline{v}_o = n\overline{v}_i$. As in a normal transformer, departures from the ideal case exist because of the effects of leakage inductances, winding capacitances, and resistive and core losses. In addition, there is a restriction on the permissible time variation of v_i for this circuit.

The operation of the repetitive pulse transformer can be understood by considering the simplified circuit shown in figure 1. This is a partial circuit that will react to posi-

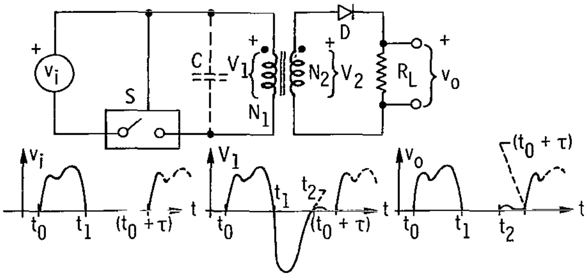


Figure 1. - Waveforms in a transformer of positive pulses.

tive pulses only; however, two such circuits can be combined, as shown later in the section Low-Frequency Bipolar Transformer, to form a bipolar transformer. The components are a switch S operated by the input v_i, a small capacitance C that may be the winding capacitance, a suitable transformer having N_1 primary and N_2 secondary turns, a diode D, and a load resistor R_L. By sensing v_i, the switch S remains closed if $v_i > 0$ and opens whenever $v_i \rightarrow 0$. When S is open, the v_i is isolated from the circuit, and the primary winding is open circuited at S.

At time t_0, the magnetic core is resting at the point marked by t_0 on its characteristic of flux φ against the magnetizing field H shown in figure 2. As v_i rises, the switch S closes, and the flux increases along the path marked by the arrows. Since the diode D is forward biased, a transformed voltage $v_o = (N_2/N_1)v_i$ appears across the

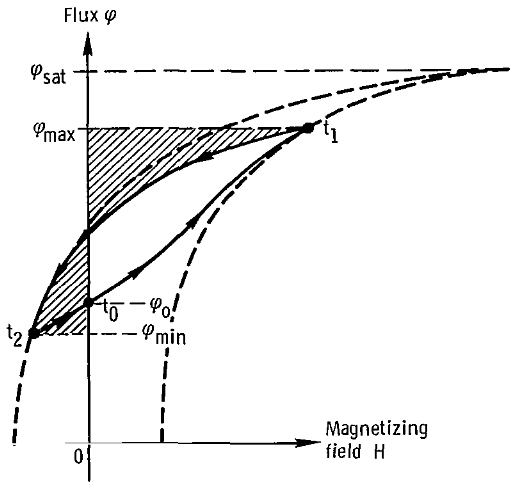


Figure 2. - Flux as function of H characteristic for core and cycled minor loop.

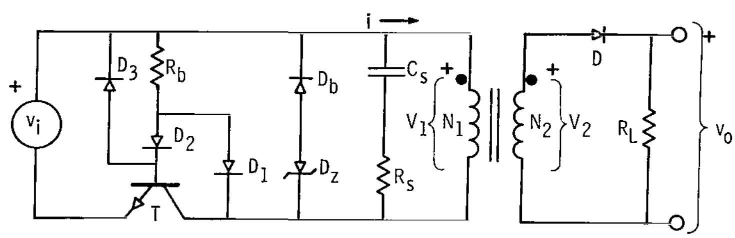
load R_L. The φ reaches a maximum value φ_{max} at time t_1, when v_i falls to zero and S opens. At this time, the terminal voltage v_1 reverses polarity, which reverse biases D. Now the core sees a high impedance through the open switch, and v_1 performs a half cycle of an LC-type oscillation, which quickly resets the flux to the point marked t_2. The capacitance is provided by the distributed winding capacitance together with any external capacitance added across the primary winding, and the inductance is provided by the nonlinear path of φ against H from point t_1 to point t_2. After time t_2, the diode D becomes forward biased, but v_1 ramains quite low because of the damping provided by R_L. The v_o pulse appearing just after t_2 can be minimized by procedures discussed in connection with practical circuits. The flux has now returned to the starting point t_0. General voltage waveforms associated with this process are shown in figure 1 for repetitive pulses v_i having a period of repetition τ.

LIMITATIONS OF REPETITIVE PULSE TRANSFORMER

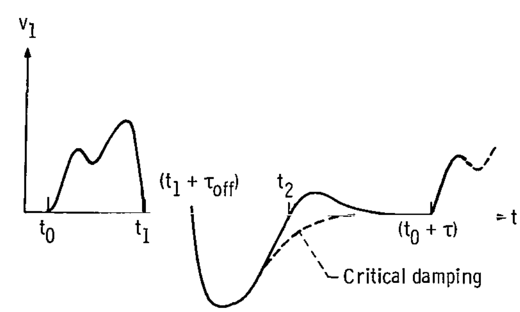
This circuit is subject to more limitations and undesirable losses than a simple transformer. Use of semiconductor diodes in series with the load current introduces an obvious power loss. The switch S, which shall normally be a transistor, introduces an even greater series voltage drop and requires energy to operate. In certain cases, these semiconductors may set the limit to the maximum speed of response. There are, however, two additional restrictions as a consequence to the mode of operation of this circuit.

Magnetic saturation sets an upper limit to the voltage-time area $\int_{t_0}^{t_1} v_i \, dt$ that can be handled. But the present circuit used only a fraction of the available flux swing of the core material. Therefore, the mass of the core is increased by a factor that is the reciprocal of this fraction, assuming the same maximum voltage-time area capability. Thus, one should pick a core material having a substantial difference between the saturation and the remanent flux. For square loop cores, the introduction of a small air gap can reduce the remanent flux and make their use practical in this application.

When operating at the limit of maximum voltage-time area, the pulse repetition period τ must be great enough to let the core complete its flux reset; that is, $\tau \geq (t_2 - t_0)$ in figure 1. If the voltage-time area is less than the maximum permissible, then one may decrease τ below this limit, forcing the cycled minor loop in figure 2 to lie higher up on the core characteristic of φ against H. The reset time $t_2 - t_1$ is the sum of the turnoff time τ_{off} of S (see fig. 3) and the duration of the oscillation. One can reduce the reset time by decreasing C, which compresses the reset oscillation. Such reduction of $t_2 - t_1$ is necessary to attain the highest permissible pulse repetition rate. It can also lead to the generation of very high voltages during reset, which can damage a switching transistor. These problems are discussed further in connection with optimum dissipation of the available energy of magnetization. This energy, which is proportional to the upper shaded area in figure 2, is returned to and partially or totally dissipated in the circuit external to the core.



(a) Circuit.

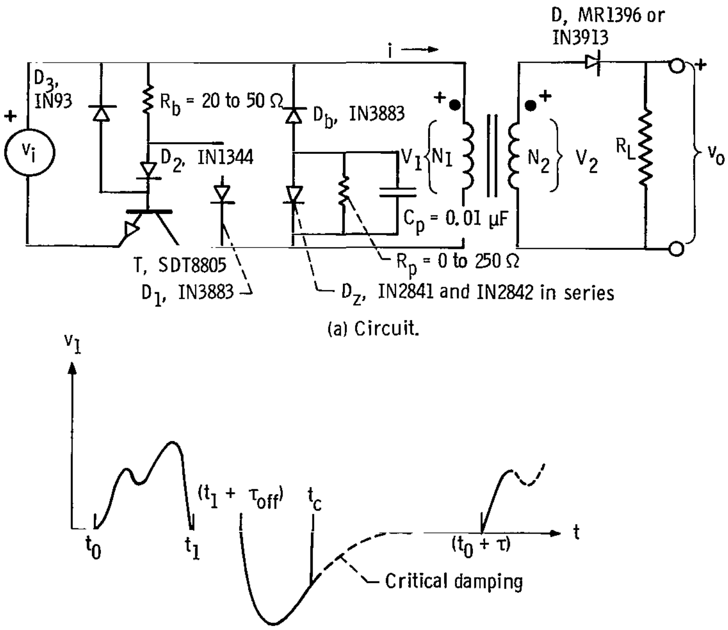


(b) Representative waveforms.

Figure 3. - Series version of practical pulse circuit.

DESIGN CONSIDERATIONS

In general, the circuit design is dependent primarily on the core material and transistors that are employed. The nonlinear characteristics involved make it difficult to supply explicit values for circuit parameters. There is a broad range of voltage, power, pulse repetition rate, rise time, and other requirements from which specifications for a particular transformer could be selected. Thus, the circuit drawn in figure 1 may not work well in many cases without some modification. A high-repetition-rate circuit that can transmit undistorted pulses requires a well-designed pulse transformer and nondestructive dissipation of the available energy of magnetization in a time interval that may be less than the pulse duration. The latter problem is discussed in connection with the circuits of figures 3 and 4, which are somewhat more specific than the circuit in figure 1.



(a) Circuit.

(b) Representative waveforms.

Figure 4. - Parallel version of practical pulse circuit.

Circuit Energy Dissipation

Two practical types of circuits are now discussed. They differ in that one circuit uses a series resistor-capacitor connection, whereas the other connects these elements in parallel.

Series circuit. - Four new components, D_b, D_z, R_s, and C_s were added to figure 3 to limit the peak voltage during flux reset and provide control over the dissipation of the available energy of magnetization. The zener diode D_z keeps the voltage within the safe operating area of the transistor for a given maximum exciting current that the transistor must interrupt. The zener diode is isolated from the input v_i by the diode D_b.

If the effects of the voltage drop in the winding resistance R_w are neglected, as soon as t exceeds t_1, $v_i = 0$, and $d\varphi/dt$ becomes negative. Because of stored charge, the transistor cannot turn off until the lapse of some turnoff time τ_{off}. Therefore, the magnitude of $d\varphi/dt$ during this interval will be quite low if R_w is small, there being a rapid rise after time $t_{off} = t_1 + \tau_{off}$. The available energy of magnetization thereafter starts to be partially dissipated in R_w and R_s and temporarily stored in C_s and in an equivalent winding capacitance C_w. The terminal voltage v_1 reaches a negative peak and decays as C_s discharges. Thus, at some time between t_1 and t_2, the exciting current i_e reverses polarity and becomes negative. Unless C_w is negligible, the current through R_s will differ somewhat from i_e. This process dissipates a significant part of the available energy of magnetization in $R_s + R_w$.

Reversal of the polarity of i_e implies that some of the energy stored in C_s and C_w is returned to the core. Appendix B shows that at time t_2

$$\int_{\varphi_{max}}^{\varphi_{min}} H \, d\varphi \leq 0 \tag{1}$$

the equality holding if $R_s = R_w = 0$. If the path $H = H(\varphi)$ is known and the equality holds, this equation determines φ_{min}. With reference to figure 2, equation (1) states that the shaded areas are equal if no losses exist in the winding and the external circuit. In this case, all the available energy of magnetization is put back into the core, which gives the minimum possible φ_{min}.

The flux rise from φ_{min} to φ_o may be quite small for cores having a significant remanent flux. Therefore, the additional flux reset generated by letting i_e become negative can provide an increased available flux swing with these types of cores.

Solving for the voltages and currents for a general core is not feasible because the relation between i_e and φ is not necessarily linear. Whenever a linear relation such as

$$N_1 \frac{d\varphi}{di_e} = L \tag{2}$$

is valid, the flux reset is oscillatory in nature if

$$\frac{C_s R_s^2}{L} \leq 4 \tag{3}$$

At equality, there is critical damping. The inequality (3) is based on a simple RLC-type circuit which neglects R_w and C_w. Sketches of representative terminal voltage waveforms are shown in figure 3(b) for the critically damped and oscillatory conditions. In the oscillatory case, one observes a positive voltage pulse across the load after time t_2. The magnitude of this pulse depends on the magnetic properties of the core and the circuit parameters, and its cause appears to be the flux rise from φ_{min} to φ_o. Generally, R_s and C_s can be adjusted to minimize the reset time $t_2 - t_{off}$ subject to the peak voltage limitations of the transistor for a given size input pulse. If D_z is a power zener diode, it too can be used to dissipate energy during flux reset.

Parallel circuit. - The circuit shown in figure 4(a) differs from the series circuit in that it uses a capacitor C_p and a resistor R_p in parallel to control the flux reset. The diode D_b isolates now not only D_z but also the C_p-R_p combination from the input v_i. A major difference between the parallel circuit and the series circuit is that, in the parallel circuit, neglecting the effects of C_w, the exciting current i_e cannot become negative.

The action of the diode D_b is now considered in greater detail. If the flux reset is oscillatory, there exists a time t_c, where $t_{off} < t_c < t_2$, at which i_e passes through zero. If C_w is neglected, i_e is then also the current at the terminals of N_1. Because D_b prevents the flow of a negative i_e, the oscillation of v_1 ends discontinuously at time t_c. At this time, the flux has returned to φ_o, and $i_e = 0$. Thus, for the parallel circuit, $\varphi_{min} = \varphi_o$ and point t_2 coincides with point t_0 in figure 2 if $C_w = 0$. For any real circuit, $C_w \neq 0$, so that $d\varphi/dt$ does not go discontinuously to zero at time t_c, although the drop can be quite abrupt. The winding capacitance C_w appears to be the cause of a small positive terminal voltage pulse that is observable under some conditions just after t_c.

The charge left on C_p at time t_c thereafter decays through R_p with a time constant R_pC_p. As R_p is increased, the t_c decreases until it approaches the time at which v_1 peaks. An increase in R_p decreases the reset time $t_c - t_{off}$ at the expense of generating a higher peak v_1. Since the R_p-C_p combination is isolated from v_i by D_b, the start of the next pulse may be immediately after t_c if the pulse width is at least several times R_pC_p.

If equation (2) holds, the condition analogous to the inequality (3) for the parallel circuit is

$$\frac{L}{C_pR_p^2} \leq 4 \tag{4}$$

Representative waveforms are shown in figure 4(b).

Little of the available energy of magnetization can be returned to the core in this circuit, and, consequently, the parallel circuit minimizes the undesirable positive voltage pulse, such as is shown after time t_2 in figure 3(b) for the series circuit. Thus, the parallel circuit may have the advantage of a better reproduction of the input waveform. The parallel circuit is also suited for use with cores having an inherently low remanent flux. In such cores, the final flux level φ_0 will be low, also, and no extra flux swing capability can be produced by the application of a negative H. However, with higher remanence cores, the series circuit would have a bigger available flux swing. In some applications, the isolation of C_p and R_p from the input v_i may favor the parallel circuit.

Energy transmission efficiency. - The total losses in these circuits consist of core loss, copper loss in the windings, losses in the semiconductors, and dissipation of the available energy of magnetization. Calculating the efficiency ϵ in general is difficult because of uncertainties associated with the semiconductors and the nonlinear properties of the core. It seems true, however, that the nonrecovery of the available energy of magnetization can account for a significant part of the total losses in certain cases.

Assuming a linear lossless core, one can calculate an upper bound for ϵ. The energy of magnetization E_{mag} for such a core is $\frac{1}{2} LI^2$, which is also the available energy of magnetization, as previously defined, because the core is lossless. Thus, in terms of the secondary voltage v_2,

$$E_{mag} = \frac{1}{2}\left(\frac{\bar{l}}{\mu A N_2^2}\right)\left(\int_{t_0}^{t_1} v_2\, dt\right)^2 \tag{5}$$

where μ is the total permeability, A is the cross section, and $\bar{l}$ is the mean length of the magnetic path of the core. The efficiency ϵ of the circuit must satisfy

$$\epsilon < \frac{E_L}{E_L + E_{mag}} \tag{6}$$

where

$$E_L = \frac{1}{R_L} \int_{t_0}^{t_1} v_o^2 \, dt \tag{7}$$

because other losses are omitted from the inequality (6). For a rectangular voltage pulse and $v_o \approx v_2$, the inequality (6) becomes

$$\epsilon < \frac{1}{1 + \frac{1}{2}\left(\frac{\bar{l}}{\mu A N_2^2}\right) R_L (t_1 - t_0)} \tag{8}$$

Power Switching Requirements

The only switching elements investigated for use in the present circuits were transistors; gate-controlled rectifiers were unsuitable because they could not easily interrupt a flowing current. This necessity to stop the peak magnetizing current under conditions in which the v_1 may reach over 100 volts in a few microseconds places severe demands on the transistor in a circuit using a high exciting current core. As used herein, the transistor may need a significantly higher V_{CEO} rating than the actual peak reset v_1 because the circuit encourages the phenomenon of reverse-bias second breakdown (ref. 1) The problem is more severe if a fast turnoff time τ_{off} is desired, because the higher speed power switching transistors are more susceptible to the second breakdown. Ideally, for a fast-rise-time power circuit, one would pick a switching transistor having a high V_{CEO} rating, sufficient power and current capabilities, and the minimum-gain bandwidth product f_t consistent with the desired speed. Limitation of the rate of rise and peak v_1 by use of the external capacitors has kept the voltages and currents within the safe operating area of the transistor in several experimental circuits.

Diodes D_1 and D_2 shown in figures 3 and 4 are used as feedback clamps to keep the collector out of saturation (ref. 2). Their presence increases the collector-to-emitter voltage drop but considerably reduces τ_{off}. A germanium diode D_3 reduces reverse leakage and speeds sweepout of stored charge. A τ_{off} as low as 3 microseconds could be achieved for some power switching transistors such as the 2N3599 or the SDT8805.

Significant Magnetic Properties of Core

The mode of operation of this pulse circuit requires a core having a substantial difference between its saturation and remanent flux. Low remanence results in a more complete utilization of the magnetic material and, hence, less core mass. Magnetically softer cores are preferred because the lower exciting current drawn by a more permeable core is easier to switch off.

Commercial supermalloy-tape-wound cores, which have the low direct current coercivity of about 0.8 ampere per meter (0.01 Oe), yielded a maximum available flux density swing of about 0.5 tesla compared with the saturation value of about 0.8 tesla. Magnetically soft square-loop materials such as supermendur can provide saturation flux densities of over 2 teslas, but because of their squareness, they are not useful in this circuit. The core can be modified by the introduction of an air gap to use a square-loop material. Appendix C shows that the introduction of an air gap equal to

$$l_{gap} = \frac{\mu_o H_c}{\alpha B_s} \bar{l} \tag{9}$$

in an ideally square-loop core will reduce it remanence from B_s to αB_s, where $0 < \alpha < 1$. The new saturation magnetomotive force M_s is then

$$M_s = H_c \bar{l}(1 + \alpha^{-1}) \tag{10}$$

but the coercivity remains unchanged at H_c. If one assumes that all the available energy of magnetization can be returned to the core on reset, the minimum B reached will be

$$B_{min} = -B_s(1 - 2\alpha) \tag{11}$$

This value will also be the B_o provided that $B_{min} \geq -\alpha B_s$, since the minimum possible B_o is $-\alpha B_s$ in any case. From the equation

$$-\alpha B_s = -B_s(1 - 2\alpha) \tag{12}$$

one finds that there is no point in having an α less than 1/3 for the case of a lossless return of the energy of magnetization with this type of core. The maximum flux density swing is

$$(\Delta B)_{max} = B_s - B_{min} = 2B_s(1 - \alpha) \tag{13}$$

and it can be no bigger than

$$(\Delta B)_{max}\Big|_{\alpha=1/3} = \frac{4}{3} B_s \tag{14}$$

Use of the parallel R_p-C_p combination precluded restoration of any energy to the core, so that circuits of that type can take advantage of an $\alpha < 1/3$. Even for the series R_s-C_s circuit, the presence of losses and various hysteresis loop geometries can lead to a best $\alpha < 1/3$.

EXPERIMENTAL CIRCUITS

Several circuits incorporating some of the ideas just discussed have been constructed. The very first bipolar transformer (ref. 3) had a rather large number of turns of fine wire that limited its frequency response and power handling capability. Now, two additional experimental circuits have been built, each of which can handle hundreds of watts of power. They were constructed from power components on hand without regard to optimization for any specific application.

Low-Frequency Bipolar Transformer

The circuit detailed in figure 5 combines two single-polarity circuits in a way that enables it to handle both positive and negative repeated pulses. It was constructed around two rather large 2-mil supermalloy-tape-wound cores, each having an effective cross section of 10.37 square centimeters and a mean circumference of 35.7 centimeters. Each core was wound with four 100-turn windings of 12-gage wire and one 50-turn winding of 10-gage wire, from which various winding combinations could be selected. The outputs of N_2 can be combined across a split load or across a single load by uses of transistor switches as shown. Darlington connected transistors were used to increase the gain at the expense of a higher voltage drop. The use of the germanium diodes D_2

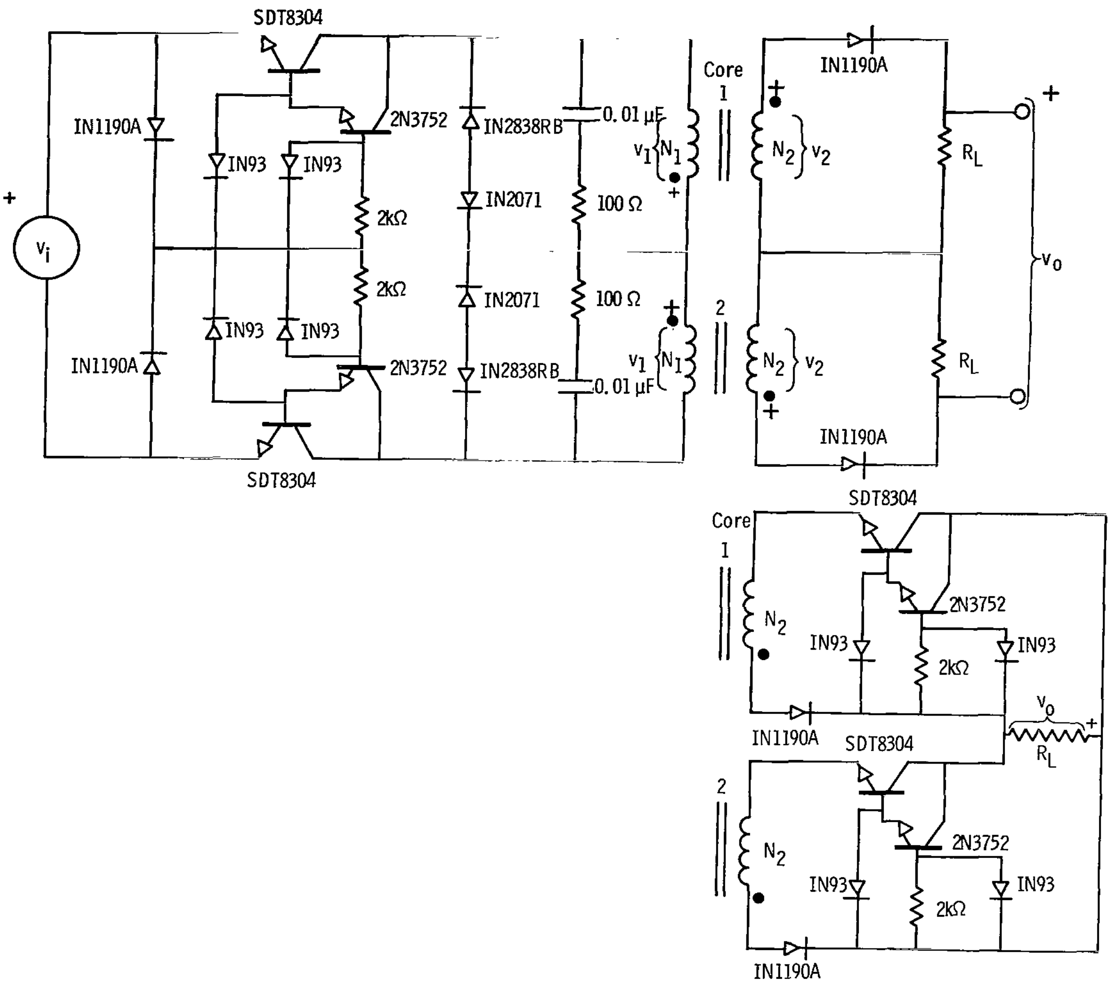
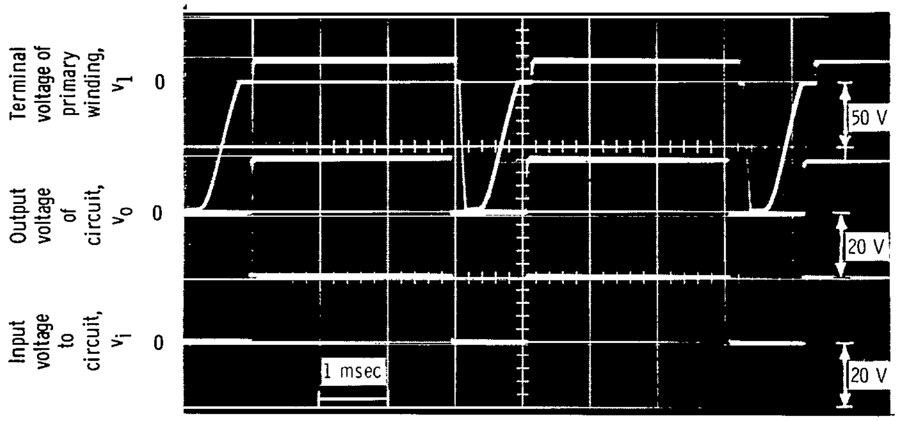


Figure 5. - Low-frequency bipolar pulse circuit. Cores 1 and 2, 50417-2F; each core includes four windings with 100 turns of 12-gage wire and one winding with 50 turns of 10-gage wire.

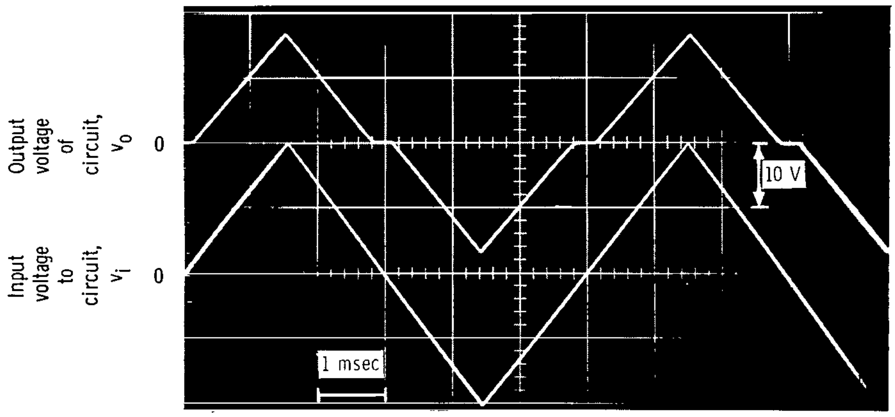
significantly reduced the reverse leakage and turnoff time of the transistors. The 40-ampere silicon rectifiers D_1 are relatively low reverse-leakage types suitable for low frequencies only.

Figure 6(a) shows the response of this circuit when excited by 20-volt rectangular positive pulses with a duration of 3 milliseconds and a period of repetition of 4 milliseconds. In this case, the load R_L was 5 ohms, and $N_1 = N_2 = 100$ turns. The middle trace shows that about 17 volts was available at the output, the 3 volts being lost in winding resistances and semiconductor drops. Thus, 58-watt pulses were being delivered at a 75-percent duty factor. Testing at higher powers could not be done because the necessary power pulse generator was not available.

Distortion caused by semiconductor attenuation of low voltages is apparent with the triangular wave input shown in figure 6(b).



(a) Response to repeated positive pulses.



(b) Response to triangular wave.

Figure 6. - Response of the low-frequency bipolar transformer driving 5-ohm load.

High-Frequency Unipolar Transformer

A simple circuit capable of responding to pulses narrower than 10 microseconds was constructed according to the diagram in figure 4. Four 24-turn windings of 14-gage wire were wound quadrifilar on a low-remanence powder core with a cross section of 7.26 square centimeters, a mean circumference of 24.8 centimeters, and a relative permeability of 150. This construction (see fig. 7) provided the low-capacitance tightly coupled windings needed for fast response.

It was necessary to use fast-recovery silicon power rectifiers such as the MR1396 or IN3913, which can handle 30 amperes and have a reverse recovery time of less than 0.2 microsecond. The use of the fast-recovery type IN3883 diode for D_b and D_1 was also desirable. The presence of the antisaturation and carrier sweep-out diodes D_1, D_2, and D_3 reduced the τ_{off} of the SDT8805, which is a 30-ampere 300-volt transistor,

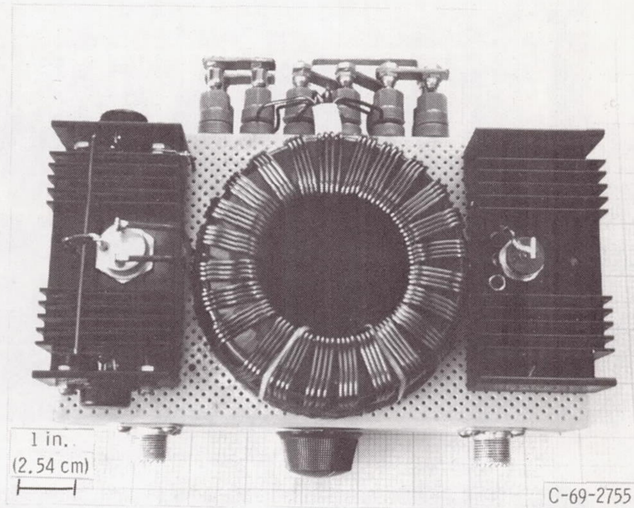
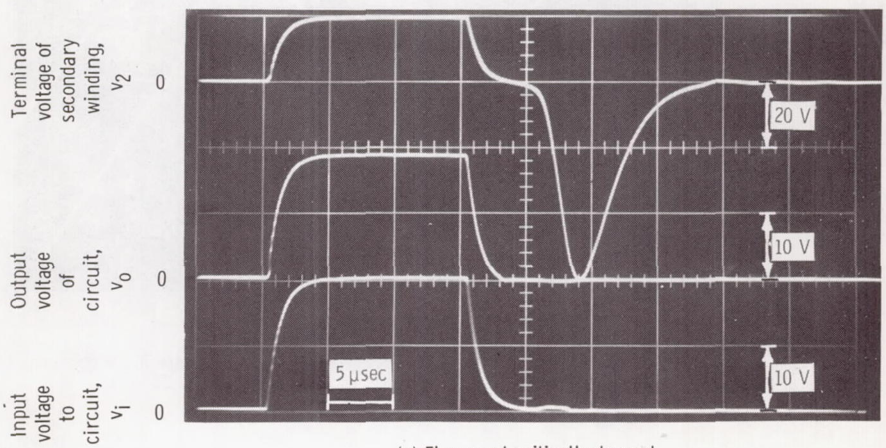
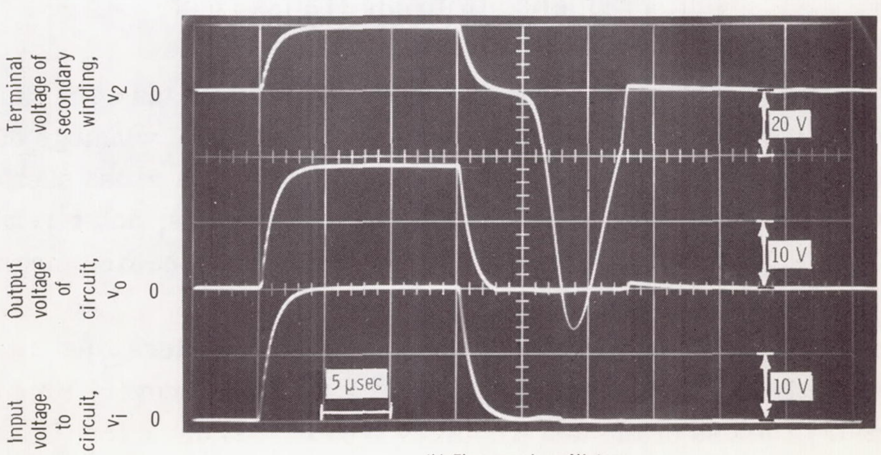


Figure 7. - Construction of high-frequency pulse transformer.

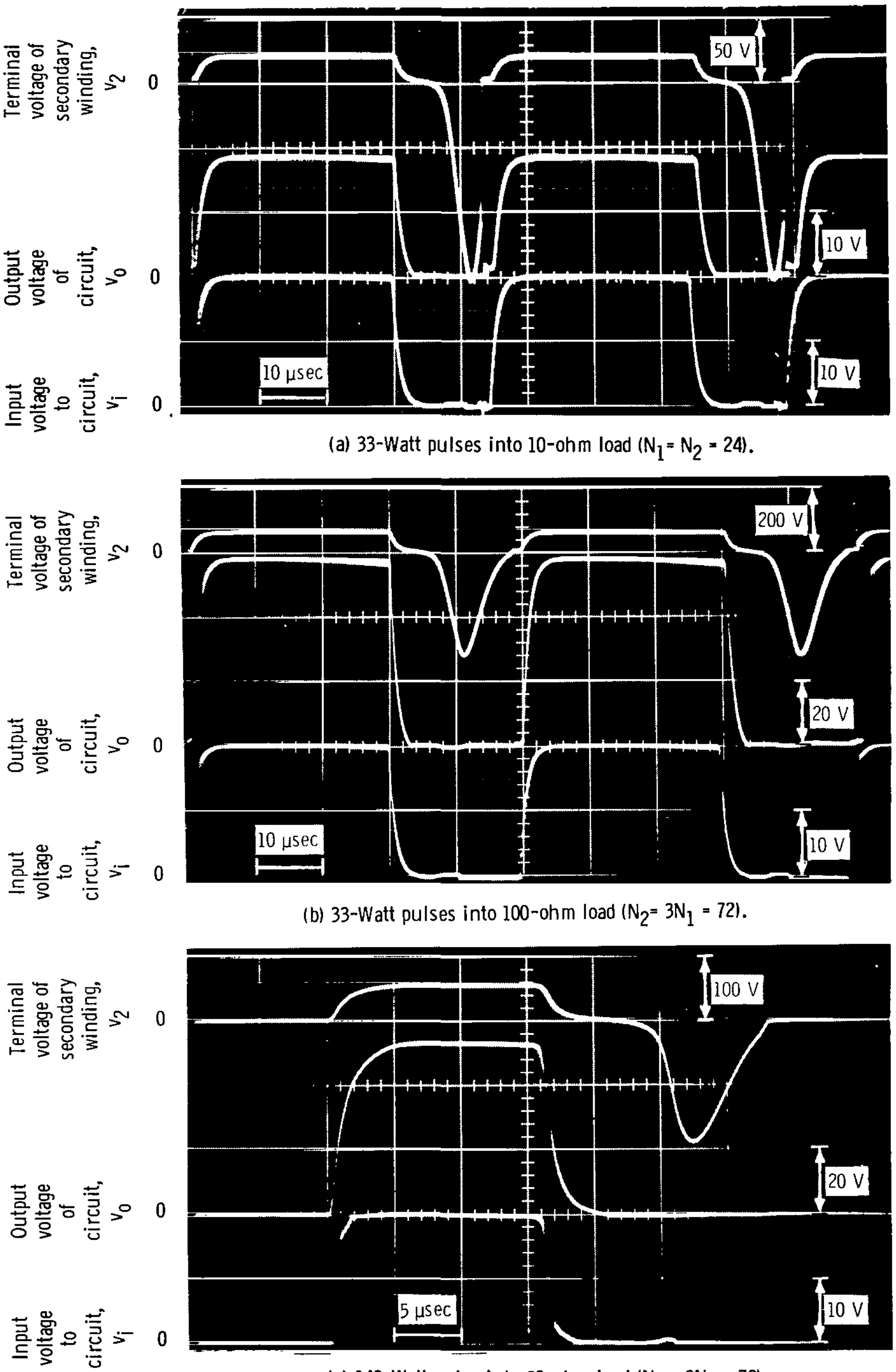


(a) Flux reset critically damped.



(b) Flux reset oscillatory.

Figure 8. - Typical response of high-frequency transformer with unity turns ratio and differing degrees of damping.



(a) 33-Watt pulses into 10-ohm load ($N_1 = N_2 = 24$).

(b) 33-Watt pulses into 100-ohm load ($N_2 = 3N_1 = 72$).

(c) 140-Watt pulse into 20-ohm load ($N_2 = 3N_1 = 72$).

Figure 9. - Response of high-frequency transformer at various power levels and loads.

to 3 or 4 microseconds. A 2N3599 was slightly faster at turnoff, but it has a maximum rating of only 100 volts.

Figure 8 illustrates some of the performance features observed when a 100-ohm load is driven with an applied v_i of 20 volts amplitude and 15 microseconds duration. The turns ratio is 1:1, with $N_1 = 24$. Figures 8(a) and (b) compare the observed traces when R_p was set for critical and less than critical damping, respectively. Thus, the turnoff and flux reset period could be compressed to slightly less than 10 microseconds.

The turns ratio, load, and pulse amplitude and duration were varied, and a few of the observations are presented in figure 9. The traces in figure 9(a) were observed while 33-watt pulses were delivered to a 10-ohm load at a duty factor greater than 50 percent and $N_1 = N_2$. The traces in figure 9(b) show pulses with the same duty factor delivering the same power to a 100-ohm load, with the transformer connected for a 3-to-1 stepup. The traces in figure 9(c) show a single 15-microsecond 140-watt pulse applied to a 20-ohm load. Testing at higher power levels and faster rise times could not be carried out for the lack of a suitable pulse generator.

Applying the inequality (8) gives $\epsilon < 0.68$ and $\epsilon < 0.95$ for the cases in figures 9(a) and (c), respectively.

SUMMARY OF RESULTS

A circuit was constructed to transform the voltage, either up or down, of repeated positive or negative pulses. Transistors, diodes, resistors, capacitors, and magnetic cores were combined in a passive circuit that overcame the problem of magnetic saturation by providing a rapid resonant flux reset.

Similar to an ordinary transformer, this circuit could drive multiple direct-current isolated loads from a single source. It could transform various pulsed waveforms that have a nonzero average value and are relatively closely spaced in time. Thus, this circuit also transformed the average value of the input voltage. Its primary disadvantages are increased core mass due to incomplete use of the total available flux swing and an efficiency below that of an ordinary transformer. The lower efficiency resulted from semiconductor losses and the loss of the available energy of magnetization. In addition, there is a minimum permissible spacing in time between successive pulses. Because of voltage drops in the semiconductors, the circuits tested would not operate with inputs below about 1.5 volts.

The experimental circuits built demonstrated their feasibility to transmit hundreds of watts to resistive loads. Operation at a duty factor greater than 50 percent was observed at pulse widths of several milliseconds and 30 microseconds with two high-power circuits. Finite turnoff time and the safe operating area of the power switching transistors were the chief limiting factors to the rate of pulse repetition.

One may be able to increase the efficiency and decrease the demands on the switching transistor by investigating variations on the circuits presented herein. For example, the possible advantages of a low-hysteresis, high-saturation, square-loop core with a small air gap may be significant in these circuits.

Lewis Research Center,
National Aeronautics and Space Administration,
Cleveland, Ohio, August 7, 1969,
120-27.

APPENDIX A

SYMBOLS

A	cross-sectional area of magnetic core
B	magnetic flux density
B_{max}	maximum B reached during cycle
B_{min}	minimum B reached during cycle
B_o	value of B at start of pulse
B_s	saturation B for core
$(\Delta B)_{max}$	maximum available change in B
C	capacitance
C_p, C_s	parallel and series capacitors, respectively, used to control flux reset
C_w	winding capacitance
D, D_b, D_1, D_2, D_3	diodes labeled in circuit diagrams
D_z	zener diode
E_L	energy delivered to load
E_{mag}	energy of magnetization
f_t	small-signal-gain bandwidth product of transistor
H	magnetic field intensity
H_c	coercivity of magnetic core
i_e	exciting current of core
L	inductance
$\bar{l}$	mean length of magnetic path
l_{gap}	length of gap cut in magnetic core
M_s	saturation magnetomotive force of magnetic core
N_1, N_2	number of turns on primary and secondary windings
n	N_2/N_1
R_b	base resistor

R_L	load resistor
R_p, R_s	parallel and series resistors used to control flux reset
R_w	winding resistance
S	switch
t	time variable
$\left.\begin{matrix} t_0, t_1, t_2, \\ t_{off}, t_c \end{matrix}\right\}$	critical times marked in cycle of operation
V_{CEO}	collector-to-emitter breakdown voltage of transistor with base open circuited
v_i	input voltage to circuit
v_o	output voltage of circuit
v_z	zener voltage of D_z
v_1, v_2	terminal voltages of primary and secondary windings
α	ratio of remanent-to-saturation magnetic flux density
ϵ	energy transmission efficiency of transformer
μ	total permeability of magnetic core
μ_o	permeability of free space
τ	period of pulse repetition
τ_{off}	turnoff time of transistor
φ	magnetic flux
φ_{max}	maximum φ reached during cycle
φ_{min}	minimum φ reached during cycle
φ_o	value of φ at start of pulse
φ_{sat}	saturation flux density
Superscript:	
	average value

APPENDIX B

DERIVATION OF ENERGY BALANCE INEQUALITY

The derivation of inequality (1) is based on the equivalent circuit drawn in figure 10. This circuit is an approximation, valid during flux reset, to the actual series version of the pulse transformer circuit. The following three equations determine the voltages and currents as functions of time:

$$v_1 - R_w i_e - N_1 \frac{d\varphi}{dt} = 0 \tag{B1}$$

$$v_1 + R_s i_s + \frac{1}{C_s} \int_{t_{off}}^{t} i_s \, dx = 0 \tag{B2}$$

$$v_1 - \frac{1}{C_w} \int_{t_{off}}^{t} (i_s - i_e) dx = 0 \tag{B3}$$

where x (and later y) is a dummy variable of integration taking the place of the time variable t. These equations assume no initial charge on C_s and C_w.

After substitution in equation (B1), the value for v_1, given by equation (B3) and multiplication by i_e, one obtains

$$N_1 i_e \frac{d\varphi}{dt} = -R_w i_e^2 + \frac{i_e}{C_w} \int_{t_{off}}^{t} (i_s - i_e) dx \tag{B4}$$

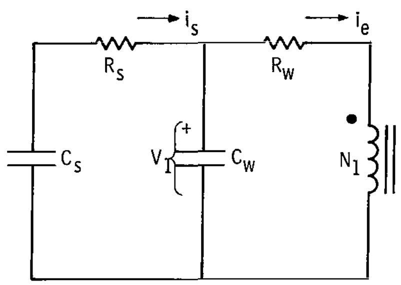


Figure 10. - Equivalent circuit for series version of pulse transformer during flux reset.

An integration from t_{off} to t transforms equation (B4) into

$$N_1 \int_{\varphi_{max}}^{\varphi} i_e \, d\varphi = -R_w \int_{t_{off}}^{t} i_e^2 \, dx + \frac{1}{C_w} \int_{t_{off}}^{t} i_e(y) \int_{t_{off}}^{y} \left[i_s(x) - i_e(x)\right] dx \, dy \tag{B5}$$

It helps in later manipulations to add and subtract the terms $i_s(y)$ to $i_e(y)$ in equation (B5). This operation changes the form of equation (B5) to

$$N_1 \int_{\varphi_{max}}^{\varphi} i_e \, d\varphi = -R_w \int_{t_{off}}^{t} i_e^2 \, dx - \frac{1}{C_w} \int_{t_{off}}^{t} \left[i_s(y) - i_e(y)\right] \int_{t_{off}}^{y} \left[i_s(x) - i_e(x)\right] dx \, dy$$

$$+ \frac{1}{C_w} \int_{t_{off}}^{t} i_s(y) \int_{t_{off}}^{y} \left[i_s(x) - i_e(x)\right] dx \, dy \tag{B6}$$

Before proceeding, a specialized formula for integration by parts is needed. Let $U(x)$ and $V(x)$ be arbitrary integrable functions of x. Then it is true that

$$\int_a^b V(y) \int_a^y U(x) dx \, dy = \left[\int_a^b U(x) dx\right]\left[\int_a^b V(x) dx\right] - \int_a^b U(y) \int_a^y V(x) dx \, dy \tag{B7}$$

In the case $U(x) \equiv V(x)$, it follows immediately that

$$\int_a^b U(y) \int_a^y U(x) dx \, dy = \frac{1}{2}\left[\int_a^b U(x) dx\right]^2 \tag{B8}$$

Returning now to (B6), one sees that, by letting $i_s - i_e = U$, applies to the first iterated integral on the right. Therefore,

$$N_1 \int_{\varphi_{max}}^{\varphi} i_e \, d\varphi = -R_w \int_{t_{off}}^{t} i_e^2 \, dx - \frac{1}{2C_w}\left[\int_{t_{off}}^{t} (i_s - i_e) dx\right]^2$$

$$+ \frac{1}{C_w} \int_{t_{off}}^{t} i_s(y) \int_{t_{off}}^{y} \left[i_s(x) - i_e(x)\right] dx \, dy \qquad \text{(B9)}$$

According to equation (B3), the second integral on the right side of equation (B9) is just $C_w v_1$. Further, from equations (B2) and (B3), it follows that

$$\frac{1}{C_w} \int_{t_{off}}^{y} (i_s - i_e) dx = -R_s i_s(y) - \frac{1}{C_s} \int_{t_{off}}^{y} i_s \, dx \qquad \text{(B10)}$$

which can be substituted in the iterated integral in equation (B9). After these substitutions, the result is

$$N_1 \int_{\varphi_{max}}^{\varphi} i_e \, d\varphi = -R_w \int_{t_{off}}^{t} i_e^2 \, dx - \frac{1}{2} C_w v_1^2 - R_s \int_{t_{off}}^{t} i_s^2 \, dy$$

$$- \frac{1}{C_s} \int_{t_{off}}^{t} i_s(y) \int_{t_{off}}^{y} i_s(x) dx \, dy \qquad \text{(B11)}$$

Formula (B8) again applies to the iterated integral. The general result proved thus is the energy balance equation

$$N_1 \int_{\varphi_{max}}^{\varphi} i_e \, d\varphi = -R_w \int_{t_{off}}^{t} i_e^2 \, dx - R_s \int_{t_{off}}^{t} i_s^2 \, dx - \frac{1}{2} C_w v_1^2 - \frac{1}{2C_s}\left[\int_{t_{off}}^{t} i_s \, dx\right]^2$$

(B12)

In particular, equation (B12) is valid at time t_2 when $\varphi = \varphi_{min}$ and $d\varphi/dt = 0$. If R_w and R_s are not both equal to zero, then (B12) requires that

$$\int_{\varphi_{max}}^{\varphi_{min}} i_e \, d\varphi < 0 \tag{B13}$$

If $R_w = R_s = 0$, then $v_1(t_2) = i_e(t_2)R_w = 0$ from (B1), $\int_{t_{off}}^{t_2} i_s \, dx = 0$ from (B2), and, consequently, (B12) reduces to

$$\int_{\varphi_{max}}^{\varphi_{min}} i_e \, d\varphi = 0 \tag{B14}$$

Inequality (B13) together with the equality (B14) constitute the desired result.

APPENDIX C

EFFECTS OF AIR GAP IN SQUARE-LOOP CORE

An ideally square-loop core has a saturation flux density B_s and a coercivity H_c. An air gap of length l_{gap} is cut into the core such that l_{gap} is small compared with the mean length $\bar{l}$ of magnetic path of the core. Also assumed is that l_{gap} is sufficiently small to provide a uniform field B_g in the gap. The core is equipped with an N-turn winding carrying a current i.

With fields in the core material denoted by the subscript m and fields in the air gap by the subscript g, Ampere's circuital law states that

$$H_m \bar{l} + H_g l_g = Ni \tag{C1}$$

Continuity of flux requires that, under the assumption of a uniform gap field,

$$B_m = B_g = \mu_o H_g \tag{C2}$$

Eliminating H_g between (C1) and (C2) results in

$$H_m \bar{l} + \frac{B_m}{\mu_o} l_g = Ni \tag{C3}$$

Note that, if the core is not saturated (i.e., $B_m = \alpha B_s$ where $0 < \alpha < 1$), then

$$H_m = \pm H_c$$

Consider now the case $i = 0$ and assume that the air gap is large enough to reduce the remanent flux density from B_s to αB_s. Then by equation (C3),

$$-H_c \bar{l} + \frac{\alpha B_s}{\mu_o} l_g = 0 \tag{C4}$$

and the length of the air gap is given by

$$l_g = \frac{\mu_o H_c}{\alpha B_s} \bar{l} \tag{C5}$$

Let the length of the air gap be given by (C5) and suppose that the current i is just large enough to saturate the core. Then, $H_m = H_c$ and $B_m = B_s$. Substitution of these values in equation (C3) yields the saturation magnetomotive force

$$M_s = H_c \bar{l}\left(1 + \frac{1}{\alpha}\right) \tag{C6}$$

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