LB-1083

A PHASE-REGULATED TRANSISTOR POWER SUPPLY

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Approved

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Transistors have many attractive features for power supply regulator circuits, and may be used effectively in Class A regulator circuits in series and shunt regulated supplies. Good regulation is obtainable and the low transistor operating voltage and useful features of complementary transistor circuits are particular advantages. However, the power handling capacity and high temperature performance of these power supplies are basically limited by the power-temperature restrictions of Class A transistor operation. Further, the use of physical heat sinks or forced air cooling to lessen these restrictions may substantially increase the size and weight of the associated equipment.

This bulletin describes the use of transistors in a constant-voltage phase-regulated power supply.[1] This power supply, which has been implemented previously with thyratrons, theoretically entails no dissipation in the regulation process.[2,3] The regulator transistors are operated as switches. Consequently, transistor dissipation is very low and large amounts of power can be controlled with small resultant transistor temperature rise, thus permitting substantial improvement in power-supply capabilities and high temperature performance over that of series or shunt regulated supplies.

A transistor switch circuit for use in this power supply is described, and operating and performance details are discussed. An experimental 0 to 1 ampere regulated power supply is described in which the average dissipation in the regulator transistors is 150 milliwatts. At 1-ampere load the overall efficiency is greater than 78 percent and with respect to the transistor switch circuits alone is 95 percent.

Phase Regulation

The technique of phase control of thyratrons for direct current regulation has been employed in a number of industrial applications as well as in regulated d-c supplies for laboratory and equipment use. This technique provides an efficient means for the control of high d-c power and makes efficient use of the power handling capabilities of the regulator elements through their employment as switches. The basic principles of power phase regulation are here summarized as introductory background for the discussion of transistors in this application.

A block diagram of a phase-regulated power supply is shown in Fig. 1. Elements A and B are rectifying switches which conduct alternately for one-half the period of the a-c input signal. The phase of the conduction intervals of A and B with respect to the input voltage can be adjusted over a 90 degree range. Using an inductance input filter with a sufficiently large inductor the current flow in each switch occurs approximately as square waves and, with full-wave rectification, results in a constant load current. By varying the conduction phase of the switches, the magnitude of current flow may

[1] This method of regulation has, at times, been termed "Thyratron Regulation." The term "Phase Regulation" is employed as a more general definition to avoid any ambiguity with respect to the type of device used.

[2] Prince and Vodges, MERCURY ARC RECTIFIERS AND CIRCUITS, McGraw-Hill Publishing Company, 1927.

[3] J. H. Burnett, "Thyratron Grid Circuit Design", *Electronics*, March 1951, p. 106.

be varied from 0 to the value determined by normal full-wave rectification. Voltage regulation is obtained by sampling the output voltage and deriving a suitable feedback signal to properly control the conduction phase of the switches to compensate for variation in input voltage and load current.

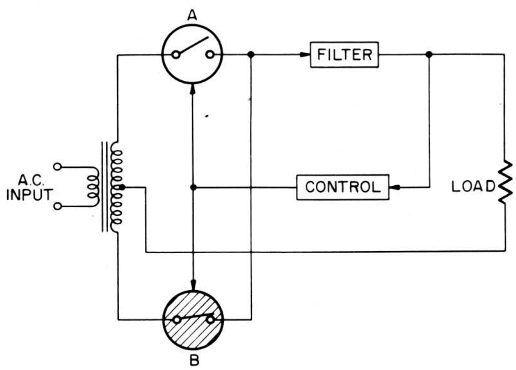

Fig. 1 - Phase regulated power supply.

The variation in direct current by phase control of the switches is illustrated in the circuit waveforms of Fig. 2 using an abbreviated diagram of the power supply. The voltage at the input to the filter, v_F, and the current in switch A, i_A, (idealized as square waves) are shown for different switch conduction phases, θ_C. The solid lines in the voltage waveforms define the conduction time of switch A. Its conduction is terminated when switch B turns ON and the filter voltage shown by the dashed lines is determined by conduction of switch B. v_A is the instantaneous transformer secondary voltage applied to switch A.

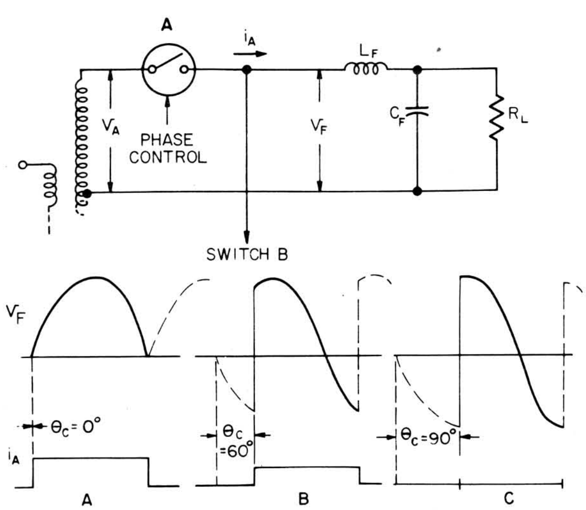

Fig. 2 - Phase control action.

For waveforms A, $\theta_C = 0$, the switches conduct in phase with the positive half cycle of the input voltage. The operation of the switches is that of a full-wave rectifier, and the d-c load current is maximum; equal approximately to $(2\sqrt{2}\,V_{A\,rms})/(\pi R_L)$ where $V_{A\,rms}$ is the rms input voltage.

The waveforms in B show switch condition delayed 60 degrees in the positive half cycle of input voltage. The filter input voltage becomes negative during the last 30 degrees of the switch conduction period. During this time, the filter inductor, L_F, effectively supplies energy back to the a-c line and the total average energy storage in L_F is decreased. Consequently, the current conducted by each switch is decreased, and for this particular case is one-half of its maximum value.

The waveforms in C are for the case, $\theta = 90$ degrees. Here, the filter voltage has equal positive and negative segments, resulting in zero average energy storage in L_F. The load voltage and current are zero.

The general relationship between d-c output voltage V_L and switch conduction phase for continuous current flow in the filter inductor is given by:

$$V_L = \frac{2\sqrt{2}\,V_{A\,rms}}{\pi}\;\frac{R_L}{R_L + R_S}\cos\theta_c \qquad (1)$$

where R_S includes the series resistance, of the switch, transformer, and filter inductor.

The switches in this power supply function as variable phase rectifiers. The filter performs as an a-c to d-c voltage converter which averages the positive and negative segments of input voltage transferred by the switches. The filter inductor also serves to maintain a forward voltage drop across the conducting switch during the negative excursion of input voltage.

The relationship between load voltage and switch conduction phase is obtained by simply transposing these variables in Eq. (1).

$$\theta_C = \cos^{-1}\left[\frac{R_L + R_S}{R_L}\;\frac{\pi V_L}{2\sqrt{2}\,V_{A\,rms}}\right] \qquad (2)$$

In design of a phase regulated supply, the value of θ_C is chosen to provide sufficient range for anticipated load and input voltage variation.

Note that the regulation does not provide filtering as in the case of a series regulated power supply. The output ripple voltage is determined solely by the amount of filtering used. The output impedance of the power supply is controlled below the fundamental frequency of the input voltage by the regulation process and at higher frequencies is determined by the filter output capacitor. The cutoff characteristic of the filter is the primary consideration in control loop feedback stability. Hence, the effective frequency range for regulator action is determined by the degree of filtering.

The power-handling capabilities of the power supply are determined primarily by the efficiency and power limits of the switch elements.[4] Transistors are well suited for this use because of their efficient switch properties and high current capabilities. For a specified output power, transistor dissipation and resultant temperature rise are substantially less than that incurred in Class A regulator circuits. An important result of this is that considerable improvement may be obtained in high temperature transistor supply performance.

Transistor Switch Circuit

There are several methods for implementing the switch circuits in the phase regulated power supply. One approach is to use amplifiers which are driven between OFF and ON conditions by suitable control circuitry. A more suitable technique, however, is the use of bistable circuits in thyratron fashion. This simplifies the control circuit mechanization, decreases switch drive requirements, and allows better decoupling between switch circuits.

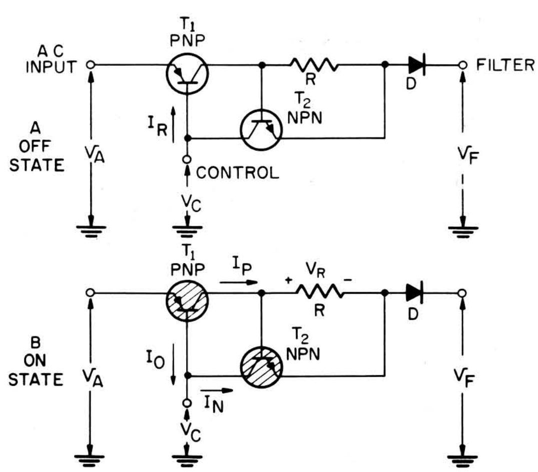

Fig. 3 - Transistor switch circuit.

A transistor circuit for the switch element in phase-regulated power supplies is shown in Fig. 3. This circuit is the basic complementary bistable configuration with regenerative current feedback provided by resistor R shunting the base-emitter junction of T_2. Its operation is similar to that of a thyratron.

[4] The filter performance at high currents may also be a major consideration.

The circuit is OFF, as shown in Fig. 3A, when the control voltage is more positive than the input voltage, v_A. This establishes an emitter-base cutoff bias on T_1 which prevents forward current flow. Diode D is required to prevent reverse conduction when the filter voltage, v_F, is more positive than v_C.

The transistor cutoff currents, including the reverse emitter and collector currents of T_1 and the collector current of T_2, are supplied from the control circuit. The collector current of T_2, when v_C is more positive than v_F, is the transistor I_{CO} slightly amplified by self-bias action from the base emitter resistor, R. When $V_A < V_C < V_F$, the total collector current of both T_1 and T_2 is limited to the reverse current of diode D.

The circuit turns ON as shown in Fig. 3B when v_A becomes more positive than v_C. T_1 is then forward biased and its collector current, I_P, produces voltage, V_R, across resistor R which forward biases transistor, T_2. T_2 then conducts and increases the base current of T_1 which in turn increases the bias for T_2. This action is regenerative and culminates in the condition where both T_1 and T_2 are in a saturation ON condition. T_2 then effectively shorts the base of T_1 to the output. Any change in collector current of T_1 adjusts the ON bias for T_2 to allow for the corresponding required change of T_1 base current. This lock-in action provides inherent circuit ON stability with switch current variation. The circuit remains stable in the ON condition until v_F is made more positive than v_A. The subsequent current reversal produces a cutoff bias on T_2 which causes the circuit to regenerate to its OFF condition.

Circuit Requirements

The OFF condition of the circuit is controlled solely by the emitter-base bias of T_1 as determined by the control and input voltages. As long as v_C is more positive than v_A, the circuit is unconditionally OFF. The value of the control voltage however is influenced by the voltage drop produced in the control source by the switch cutoff current. This becomes an important consideration at high temperatures and is discussed in the section on temperature effects.

The stability of the circuit in the ON state is determined by the complementary bias conditions necessary to maintain T_1 and T_2 in a saturated ON condition. To maintain T_1 ON, the collector current of T_2, I_N must exceed the minimum required base current of T_1 or:

$$I_N \geq \frac{I_P}{B_1} \qquad (3)$$

where B_1 is the d-c base-to-collector current amplification factor of T_1.

The total load current, I_L, is equal to $I_P + I_N$ and since $I_N << I_P$, the ON requirement for T_1 may be expressed approximately as:

$$I_N \geq \frac{I_L}{B_1} \tag{4}$$

A portion of the base current of T_1, I_O in Fig. 3B, will also flow in the control circuit. The magnitude of I_O depends upon the coupling between the control and switch circuits and varies as the magnitude and polarity of input voltage. When v_A is more positive than v_C, this current aids the ON switch bias. However, with v_A negative to v_C, current is supplied from the control source and adds to the collector current of T_2. If this reverse control current is significant, allowance must be made for it in the ON bias of T_2.

T_2 is biased ON by the base-emitter voltage, V_R, developed across resistor R. R is made small to limit power loss and the total switch resistance. Consequently, T_2 may be considered as voltage biased and V_R is given closely by:

$$V_R = I_P R \cong I_L R \tag{5}$$

The ON condition of T_2 is then determined on the basis of its d-c transconductance, factor, G_{M2}, and neglecting the reverse control current is expressed as:

$$V_R \geq \frac{I_N}{G_{M2}} \tag{6}$$

Allowance for the reverse control current is made by its addition to I_N in this expression.

Combining expressions (4), (5), and (6), the value of R to insure circuit ON stability is:

$$R \geq \frac{1}{B_1 G_{M2}} \tag{7}$$

For relatively constant current operation, R is determined by the appropriate values of B_1 and G_{M2}, with allowance for their variation over the specified current range. For wide current operation the value of R depends primarily on G_{M2} at the minimum operation current. At very low currents, a relatively large value of R may be required. This increases the total switch series resistance and, at high switch currents, may result in considerable power dissipation. An alternative bias method for T_2 for wide current operation is the diode bias shown in the switch circuit in Fig. 4. The relatively constant voltage forward characteristic of

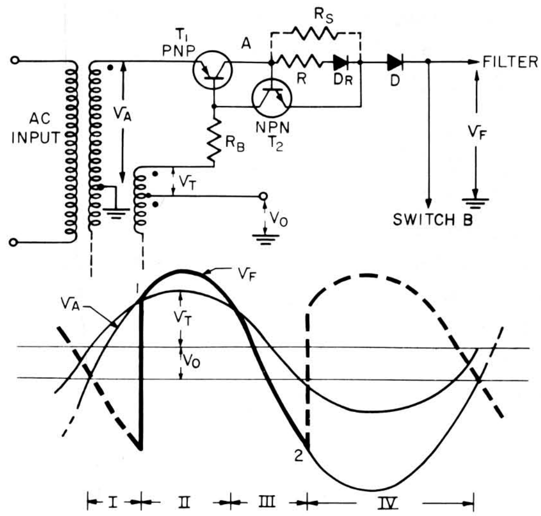

Fig. 4 - Power supply wave forms

diode, D_R, maintains sufficient voltage bias for T_2 at low currents, thereby allowing a small value of R for satisfactory high current operation. If the diode characteristic properly matches the saturation input characteristic of T_2, R may be eliminated entirely. A consequence of diode bias is that the cutoff current of T_2 may increase substantially from the greater self bias action. Resistor, R_S, shunting the diode, may be utilized to decrease this effect.

Power Supply Operation

The operation of the transistor switch circuit in a phase regulated power supply is illustrated in Fig. 4. Switch conduction is controlled by a variable d-c voltage, V_o, added to a constant a-c voltage, v_T, which is in phase with the input voltage. The switch conduction phase in terms of the circuit voltages is given by:

$$\theta_c = \sin^{-1}\left[\frac{V_o}{V_A - V_T}\right] \tag{8}$$

where V_A and V_T are peak amplitudes of the a-c input control signals respectively. The control sensitivity of the switch conduction phase depends both upon the change of V_o and the amplitude of V_T. By making V_T large, a wide variation in θ_C is obtained with a relatively small change in V_o. The definition of the time of switch conduction, however, becomes low for large values of θ_C where the slopes of v_T and v_A become equal. A minimum switch current may be specified on this basis for a required control precision. Improved timing definition may also be obtained by phase shifting v_T with respect to v_A although this decreases the control sensitivity of V_o and imposes the problem of suitable phase shift action with widely varying loads. The resistor R_B provides decoupling between switch and control circuitry during the switch conduction period.

The different operating conditions of the switch circuit are outlined in the circuit waveforms in Fig. 4.

In region I, the circuit is cut off with $V_o + v_T > v_A$. The filter voltage is negative as determined by the conduction of switch B. The maximum reverse voltage is impressed across the transistors just prior to the time of switch conduction and is greatest for $\theta_C = 90^0$. The transistor voltage limits specify for this condition the maximum permissible input voltage. This consequently sets the maximum load current (at $\theta_C = 0^0$) for a given power supply voltage. At point 1, the circuit turns ON as v_A exceeds $V_o + v_T$.

In region II, the circuit is conducting with $V_o + v_T < v_A$. This is the normal switch ON condition as described previously.

In region III, the circuit is conducting with $V_o + v_T > v_A$. During this time, a reverse control current flows to the switch and adds to the current conducted by T_2. The magnitude of reverse current depends on the value of R_B, and is greatest just prior to the time the circuit is turned OFF.

At point 2, the circuit switches to cutoff as the alternate switch is turned ON. In the commutating action, both switches turn OFF and turn ON nearly in synchronism. This results in a smooth transition of load current between switches. Typical switching times at 1 ampere are 5 to 10 microseconds using transistors having 500-kc cutoff frequencies.

In region IV, the circuit is cutoff with the filter input voltage positive with respect to the switch input and control voltages. The reverse switch current is then determined by diode D, and the voltage across T_2 and the collector junction of T_1 is small. The emitter junction of T_1, however, must withstand the difference between the maximum values of input and control voltages.

Temperature Effects

Two temperature effects are important in the switch operation. These are ON state stability at low temperature and variation in θ_C due to increase in transistor cutoff currents at high temperatures.

The circuit stability in the ON condition is determined by the B of transistor T_1 and the G_M of transistor T_2. Both gain factors decrease at low temperatures; consequently, their minimum values at the lowest operating temperature determine the circuit design for ON stability. The decrease in base-emitter d-c conductance of T_2 at low temperatures necessitates a proportional increase in its ON bias voltage. Using resistance bias, this will further increase the switch resistance and the power loss at high switch currents. With diode bias, however, the temperature variation of the diode d-c conductance compensates that of the transistor.

At any temperature, the cutoff currents of the switch transistors and of diode D will produce a voltage across the control circuit coupling resistor, R_B. In operating region I, this voltage adds with the control voltage in determing the conduction angle, θ_C. As temperature is raised, the increase in switch cutoff current will decrease the control voltage at the switch and thereby decrease the conduction phase.

In this case, θ_C is given by:

$$\theta_c = \sin^{-1}\left[\frac{V_o - I_R R'_B}{V_A - V_T}\right] \qquad (9)$$

where R'_B includes R_B and the d-c resistance of the control circuit. The maximum value of θ_C, and therefore the minimum regulated load current in power supply operation, is limited to:

$$\theta_{c_{max}} = \sin^{-1}\left[1 - \frac{\hat{I}_R R'_B}{V_A - V_T}\right] \qquad (10)$$

where $\hat{I}_R$ is the switch cutoff current at the highest operating temperature.

The amount of d-c coupling between switch and control circuit is a compromise between minimizing the cutoff current variation of θ_C and limiting the control circuit current during switch conduction. The output resistance of the control circuit together with the value of R_B also determines the amount of decoupling between switch circuits. Because of the mutual coupling by the control circuit impedance, any unbalance in switch cutoff currents particularly at high temperatures will result in unsymmetrical switch operation.

Developmental Power Supply

A developmental phase regulated power supply which was constructed to evaluate the performance of the transistor switch circuits is shown in Fig. 5. The feedback circuit is a conventional transistor d-c amplifier with a voltage gain of approximately 50. The 1N91 diodes are used to compensate for temperature variation in the operating point of the input stage. Control isolation between the switch circuits is provided by the 470-ohm resistors.

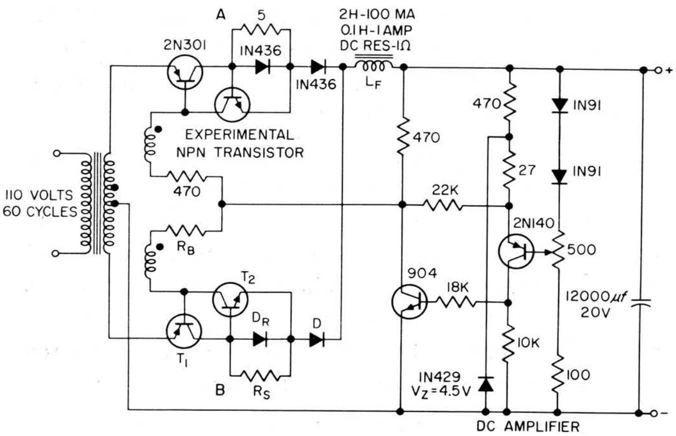

Fig. 5 - Developmental power supply.

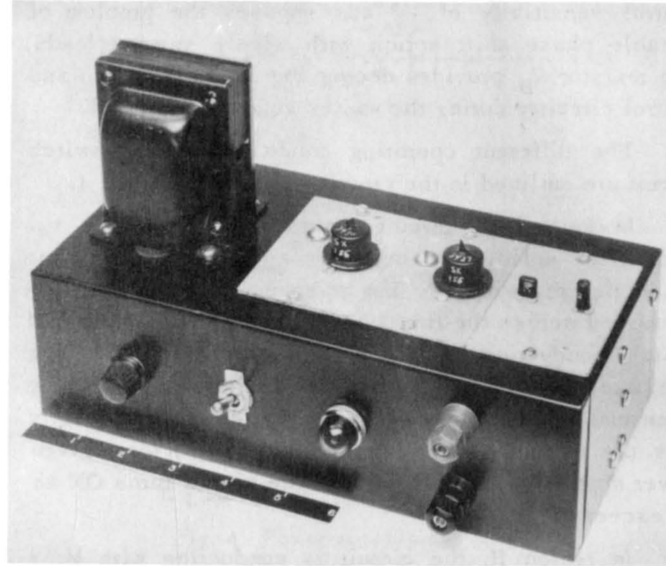

Fig. 6 - Developmental 12-volt 0-1 amp. regulated power supply.

For an output voltage of 12 volts, the following performance was obtained.

Regulation	0.4%	0 to 1.2 amp., load current
	1%	85 to 145 volts, input voltage
Ripple Voltage	16 millivolts	1 amp.
Overall Efficiency	78%	1 amp.

Phase control of the switch circuits is obtained from 5 degrees to approximately 80 degrees. The lower limit is set principally by the minimum output voltage of the d-c amplifier. The upper limit is established by the minimum switch current for stable ON circuit operation. The d-c amplifier and the 500-ohm sampling network provide the necessary bleeder for this minimum switch current.

The maximum load current of 1.2 amperes was imposed solely by the limit on the a-c input voltage due to the breakdown voltage of the particular transistors used in the switch circuits. It may be increased using higher voltage transistors and may be considerably larger if the minimum load current (condition of maximum reverse switch voltage) is increased. The load ripple voltage is determined by the filter elements. Good filtering is required to prevent overdriving of the feedback amplifier by the ripple signal. This is a major consideration in specifying the gain of the amplifier and hence the voltage regulation.

At 1-ampere load current the average dissipation in the p-n-p switch transistors is approximately 150 milliwatts, only 2 percent of the output power. The overall power supply efficiency at 1 ampere is greater than 78 percent.[5] A series regulated supply with comparable performance would entail a dissipation of perhaps 6 watts in the regulator transistor(s) with an overall efficiency of less than 50 percent. The high efficiency of the phase-regulated supply also reflects in a large reduction in transformer size.

Elevated temperature tests were made on the power supply at an output voltage of 8 volts. The operation was limited at 55 degrees C for 1 ampere load current due to phase modulation by the switch circuit cutoff currents.[6] Voltage regulation at each temperature in this range was relatively constant at 0.5 percent; however, a 2 percent overall shift resulted from d-c amplifier drift.

A photograph of the power supply is shown in Fig. 6. The switch circuits and d-c amplifier are mounted on detachable modules. The filter elements are mounted inside the chassis.

[5] Because the dissipation in the transistor circuits is only a function of load current the power supply efficiency depends on the output voltage, and may be much greater than the figure quoted with higher voltage operation.

[6] This restriction resulted only from the relatively high cutoff currents in the power transistors used. Low-power transistors (50 mw) having small cutoff currents provided satisfactory power supply operation above 75 degrees C at 300 ma load current.

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