A Transistorized Power Supply for an Ultra-violet Discharge Lamp

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A power supply is described which will operate a Flood-Hilger deuterium discharge lamp, which emits a continuous spectrum from the visible to below 2 000Å. The lamp filament is energized for 90sec and after the first 60sec, a 300V striking pulse is applied. The arc is then stabilized at a predetermined current in the range 0·35 to 0·55A, the load stability being better than 18 000:1. Semiconductor devices are used throughout.

(Voir page 762 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 769)

MANUAL and automatic spectrophotometers designed to operate in the ultra-violet region have become common in analytical laboratories and in industry[1], and it is normal practice to use an electrical discharge in hydrogen—or, more recently, deuterium—to provide a continuous ultra-violet source spectrum. Such discharge lamps work at quite high voltages which are subject to very wide variations from unit to unit. This article

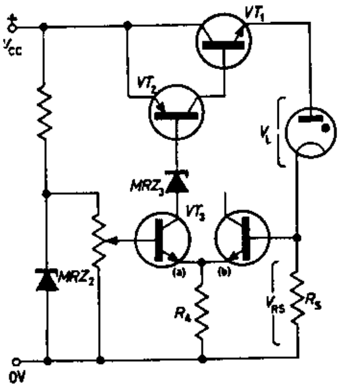

Fig. 1. Basic current stabilizer

describes a fully transistorized power supply designed to operate the Flood-Hilger deuterium discharge lamp—a source which will provide a continuous spectrum ranging from the visible down to wavelengths below 2000Å.

The Flood-Hilger lamp has the following striking sequence and operating characteristics:

(1) Filament 'on' for 90sec at 4V 8A r.m.s.

(2) Striking pulse of not less than 300V to be applied 60sec after filament is switched on.

(3) Arc current to be stabilized within the range 0·35 to 0·55A d.c. The drop across the lamp lies between 40 and 90V.

The basic stabilizing loop will be seen from Fig. 1 to be entirely conventional. However, only with the recent introduction of high-voltage transistors at reasonable prices has it become possible to use this type of circuit without including a thermionic valve as the series element.

If a Zener diode is to be used as a reference element, it is convenient to choose, where possible, a 5·6V unit so that the temperature coefficient is minimal. This is possible in the present case by making the value of $R_S = 10\Omega$ so that the voltage drop across it will not exceed 5·5V. This implies that R_S must be capable of dissipating at least 3W continuously without a significant change in resistance. A contact-cooled, metal cased 5W unit is excellent for the purpose, for it is of small size and can be screwed to the chassis for cooling purposes.

To determine the minimum supply voltage required, maximal voltage drops around the load circuit may be summed:

$$V_{CC(min)} = V_{CE1(min)} + V_{L(max)} + V_{RS(max)}$$
$$= 3{\cdot}5 + 90 + 5{\cdot}5$$
$$= 99\text{V}$$

Here, 3·5V has been assumed to be the minimum permissible value of V_{CE1}.

In a real circuit, the change in V_{CC} due to mains variations is usually taken as +10, −15 per cent, which means that, neglecting transformer regulation, the nominal value of V_{CC} will be:

$$V_{CC} = \frac{99}{(1-(15/100))} = 116{\cdot}5\text{V}$$

and $V_{CC(max)} = 116{\cdot}5\,(1 + (10/100)) = 128\text{V}$.

It is now possible to determine the maximum value of V_{CE} with which VT_1 must cope:

$$V_{CE1(max)} = V_{CC(max)} - V_{L(min)} - V_{RS(min)}$$
$$= 128 - 40 - 3{\cdot}5$$
$$= 84{\cdot}5\text{V}$$

VT_1 must also dissipate a maximum power of:

$$P_{diss(max)} = 82{\cdot}5 \times 0{\cdot}55 \simeq 45\text{W}$$

These factors lead to a choice between several transistors, and taking cost into account, the optimal unit at the time of writing is the Silicon Transistor Corp. 2N3234 which has the following characteristics:

$$V_{CEM} = 160\text{V}\ ;\ P_{tot} = 60\text{W}\ ;\ h_{FE} = 18 \text{ to } 40$$

Note that in the event of a short-circuit fault developing in the load, this transistor would not be immediately damaged by the application of almost the whole of V_{CC}, but would survive until thermally destroyed by the resultant maximum dissipation of 66W, which is the power corresponding to the 0·55A current level.

The pnp half of the stabilizing pair, VT_2, must also cope with almost the whole of V_{CC}, and must be capable of dissipating a power given by:

$$P_{diss(max)} = V_{CE2(max)}\ .\ I_{C2(max)}$$
$$= (82{\cdot}5 - V_{BE1})\ .\ (0{\cdot}55/18)$$
$$= 81{\cdot}5 \times 0{\cdot}03 \simeq 2{\cdot}5\text{W}$$

A unit fulfilling these requirements is the Motorola 2N3635, which has the following characteristics:

$$V_{CEM} = 140\text{V}\ ;\ P_{tot} = 5\text{W (with heat sink)}\ ;\ h_{FE(min)} = 80$$

The comparator/amplifier is a dual npn transistor, a choice which minimizes the number of resistors by virtue of accepting the base current of VT_2 as its collector current;

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and which makes for excellent stability. The device in question will have a maximum I_C in its left half of:

$$I_{C3(a)(max)} = \frac{I_{C2(max)}}{h_{FE2(min)}}$$
$$= \frac{0{\cdot}55}{18 \times 80} = 0{\cdot}38\text{mA}$$

By reason of its low cost and high V_{CEM} (60V), the Fairchild 2C425 is eminently suitable here. However, since the base of VT_2 can approach 128V, a series Zener diode, MRZ_3, must be included in the collector lead as shown.

The worst case situation is where $VT_{3(a)}$ approaches cut-off, and little voltage is dropped in R_4, when, if $V_{CE3(a)}$ is not to exceed 60V:

$$V_{Z3(min)} = V_{CC(max)} - V_{CEM3}$$
$$= 128 - 60$$
$$= 68\text{V}$$

A suitable Zener diode is consequently the 1N1792, which has a nominal Zener voltage of 75 ± 10 per cent; that is, $V_{Z3(min)} \simeq 68$V and $V_{Z3(max)} \simeq 82$V.

If a 'maximum' 1N1792 is fitted, then the lowest possible value for $V_{CE3(a)}$ is:

$$V_{CE3(a)(min)} = V_{CC(min)} - V_{Z3(max)} - V_{R4} - V_{BE2}$$
$$= 99 - 82 - 5 - 1$$
$$= 11\text{V}$$

The circuit of Fig. 2 shows how this basic current stabilizing loop has been incorporated into the complete system. Here, the reference Zener diode, MRZ_2, has itself been current stabilized by the regulator MRZ_1 and the resistor R_2. The actual load current is adjusted by RV_1, the lower limit being determined by R_3.

The remainder of the circuit accomplishes the striking sequence as follows:

At switch-on, the discharge lamp heater is fed from the filament transformer T_2 via the normally-closed contacts 5 and 6 of relay *A*. This transformer also supplies the heater of the 90sec thermal delay relay *B*, which upon operation, energizes relay *A*, which is then held on by its own contacts 11 and 9.

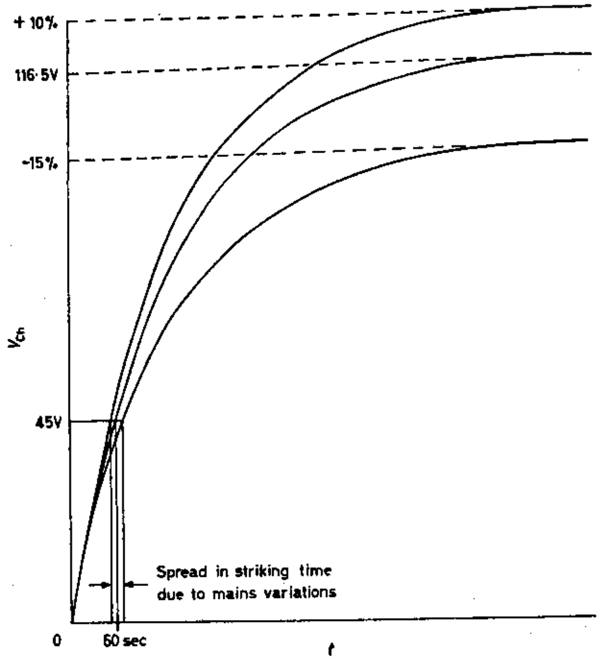

Fig. 3. Charging exponentials for C_h
(Shockley diode 1N3836 breaks over at 45V ±4V)

However, the lamp is scheduled to strike 30sec prior to the operation of the relays, and this operation is initiated by the firing of the Shockley diode D_h. The capacitor C_h is charged via R_h from the nominal 116V line. It will be noticed that this charging current is derived from the lower end of the relay *A* coil, which, after relay operation, reverts to zero potential so that C_h does not continue to charge.

The time-constant $R_h . C_h$ is so arranged that at

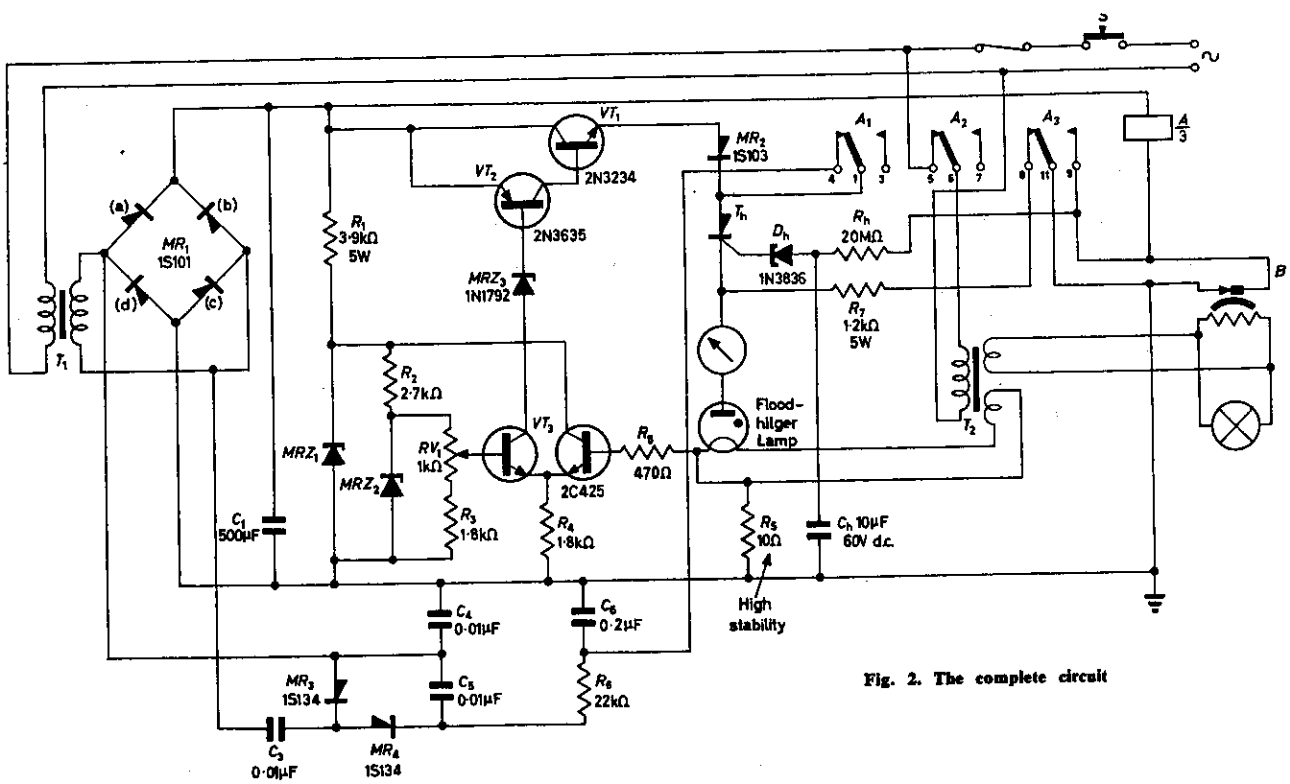

Fig. 2. The complete circuit

$t = 60$sec, the firing voltage of D_h is reached. It is convenient to choose a Shockley diode whose firing voltage is on the more linear part of the charging exponential (as shown in Fig. 3), for this ensures accurate timing.

The thyristor T_h is in turn fired by the discharge of C_h into its gate via D_h. This applies a 300V pulse to the lamp, which strikes and so completes the main stabilizing loop. The striking voltage is generated by a tripler circuit of the form shown in Fig. 4. When applied to the main circuit, one of the diodes, MR_{1c}, becomes one of the main rectifier diodes, with the result that not only is a component saved, but the negative end of the tripler output is automatically connected to the common line. The +300V end is now applied to the junction of the thyristor and the isolating diode MR_2 via contacts 4 and 1 of relay *A*.

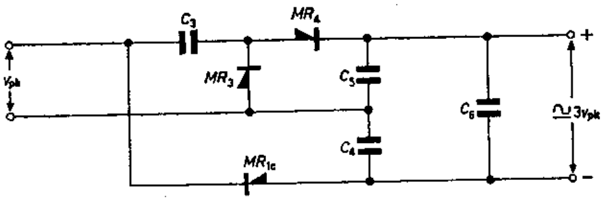

Fig. 4. Basic tripler circuit

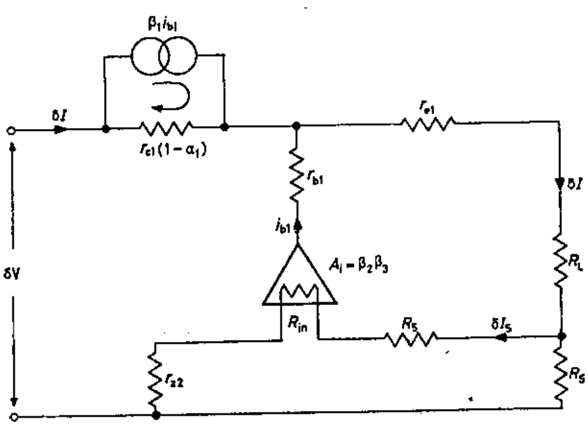

Fig. 5. Circuit for analysis

It will be noticed that R_7 has been included as a return path for the thyristor firing pulse, which would otherwise be connected only to the lamp, which is an open circuit prior to striking. R_7 must have quite a low value to ensure reliable striking, which means that it must be uncoupled from the circuit during normal operation, otherwise it would affect the stability by shunting the lamp and R_S.

This is accomplished by contacts 8 and 11 of relay *A* and is one of the five functions of that relay which, summarized, are:

(a) Self-latching

(b) De-energization of T_2 (filament switch-off at 90sec)

(c) Removal of C_h charging current

(d) Removal of lamp striking voltage

(e) Uncoupling of thyristor striking pulse return path.

The priming and striking functions are now completely isolated from the main stabilizing loop and will remain quiescent until the mains switch is manually switched off, then on again. This will be so even though the lamp may become extinguished, or be unplugged.

The analysis of the circuit can be most easily carried out by inserting the T-parameter equivalent of VT_1 as shown in Fig. 5. Summing voltages round the output loop gives:

$$\delta V = (\delta I - \delta I_S) R_S + \delta I\,[R_L + r_{e1} + r_{c1}(1-\alpha_1)] - \beta_1 i_{b1} r_{c1}(1-\alpha_1)$$

If r_{e1} is small and $\delta I_S \ll \delta I$, this reduces to:

$$\delta V = \delta I\,[(R_S + R_L + r_{c1}(1-\alpha_1)] - \alpha_1 i_{b1} r_{c1}$$

Now:

$$i_{b1} = \beta_2\beta_3\delta I_S = \beta_2\beta_3\frac{\delta I\,.\,R_S}{R_{in}}$$

where $R_{in}\ (\gg R_S)$ = input resistance of the long-tailed pair.

$$R_{total} = \delta V/\delta I = R_S + R_L + r_{c1}(1-\alpha_1) - \alpha_1\beta_2\beta_3\frac{R_S\,.\,r_{c1}}{R_{in}}$$

The effective resistance of VT_1 is therefore:

$$R_{eff} = R_{total} - R_L = r_{c1}(1-\alpha_1) - R_S\,(1-\alpha_1\beta_2\beta_3\,(r_{c1}/R_{in}))$$

If $\alpha_1 \to 1$, this approximates to:

$$R_{eff} = \beta_2\beta_3 r_{c1}\,(R_S/R_{in}) \quad\ldots\ldots\ldots\ldots\ (1)$$

This expression can be solved only if r_{c1} is known, and neither this nor the equivalent *h*-parameters appear in the published data for any of the transistors usable in the VT_1 position. Also, R_{in} is dependent upon the position of the slider of RV_1. If this is at the top of the track:

$$R_{in} = R_5 + 2r_{b2} + 2r_{e2} + r_{z2} \quad\ldots\ldots\ldots\ (2)$$

where r_{z2} is the incremental resistance of MRZ_2

(This expression, however, is very approximate, for R_4 cannot be considered large in relation to R_5 and r_{z2}.)

If R_{in} is assumed to be about 5kΩ and r_{c1} is estimated at 10kΩ, then R_{eff} is 15kΩ.

R_{eff} is clearly the resistance involved when both load and mains stability are being considered, and in fact, these parameters were measured and over much of the working range R_{eff} was found to be of this order.

The mains stability, $\delta V/\delta I$, was always better than 12 000:1 while the load stability was better than 18 000:1. Knowing that $V_{CE(VT1)}$ was required to range from 3·5 to 84V, it would be expected that the value of r_{c1} would change significantly, leading to a wide variation in stability.

The temperature performance of the power supply depends upon the temperature coefficient of MRZ_2, and since this can be a selected unit, the performance can be made very good. Using an unselected Texas 2056A, the regulated current was found to have a maximum slope of only 10μA/°C over the range 25° to 50°C ambient

When VT_1 and VT_2 are types 2N3234 and 2N3635 respectively, no high-frequency oscillation is observed because of their widely differing frequency responses. However, should alternative devices be used and such oscillation occurs, a small capacitor connected between the VT_2 collector and base will damp it successfully by reducing the high frequency loop gain around the VT_1 and VT_2 circuit.

The performance of this power supply makes it suitable for use in both double and single beam optical instruments, but should an even higher order of stability be required, the long-tail resistor, R_4, could be taken down to a negative line. Also, a second differential stage could be added to supplement the gain of VT_3.

Acknowledgments

Thanks are due to Messrs. Hilger and Watts Ltd (for whom the supply was developed), for their permission to publish this article. The author would also like to acknowledge the valuable assistance of Mr. A. Sanghvi who carried out much of the initial experimental work at Swansea.

REFERENCE

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