A PULSE GENERATOR USING A FOUR LAYER DIODE

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PREFACE

With the advent of the transistor there has been continuing extensive efforts to control the electron while it remains within the interior of the solid material. As the electron remains in the solid at all times, it is not necessary to expend energy overcoming the work function of the material to produce electrons in vacuo where they may be controlled. This results in a more efficient device and usually a much smaller one. The pulse generator designed in this work has as its basic element a solid state device, the four layer diode.

The author wishes to express his appreciation to Texas Instruments, who furnished the necessary devices and to Dr. Harold T. Fristoe who directed the overall project. Specifically, the author wishes to thank Dr. Fristoe for his suggestions on temperature stability. The author also wishes to express his gratitude to his wife for her assistance and encouragement.

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CHAPTER I

INTRODUCTION

The PNPN four layer diode is a solid state device exhibiting a negative resistance region in its VI characteristics as depicted in figure 1-1. This implies two stable states, one of high impedance and one of low impedance, which prove to be of interest in considering the device as as switching element. The negative resistance region is of interest in considering the device as a possible oscillator. Some of the many and various uses of the PNPN are as a replacement in circuits using thyratrons, switching devices, relay applications and oscillator circuits.

A device similiar to the one described above was first discussed by Shockley[1] under the broader classification of devices with h_{FB} greater than unity. The operation of the device was explained on the basis of a hook region in the energy level diagram. This concept will be discussed more fully in the consideration of the operation of the present device, which differs from Shockley's common base element in that a common emitter configuration was used.

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[1]W. Shockley, <u>Electrons and Holes in Semiconductors</u>, D. Van Nostrand Company, Inc. New York, New York, 1950.

J.J. Ebers[2] published one of the first articles suggesting an equivalent circuit. The next two pertinent and descriptive articles were published in 1956 and 1957. Moll[3] ----etc. published the first paper which quantitatively and qualitatively described the device as a two terminal element.. Using Moll's two terminal device as a building block, I.M. Mackintosh[4] continued building to the theory of the PNPN parameters to include the device as a three terminal element. It was only after the study of the last two important papers and with all this background material at hand that any attempt was made to investigate the properties of the PNPN used in this project.

Device Operation

At this point it is helpful in understanding the PNPN switching mechanisim through an analogy between the control of a vacuum tube and a transistor. In a vacuum tube an excess or deficiency of electrons create a field about the grid which controls the flow of electrons in the space around the

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[2] J.J. Ebers, "Four-terminal PNPN Transistor," Proceedings of the IRE, Vol. 40, November, 1952, pp 1361-1364.

[3] J.L. Moll, M. Tanenbaum, J.M. Goldey and N. Holonyak, "PNPN Transistor Switches", Proceedings of the IRE, Vol. 44, September 1956, pp. 1174-1182.

[4] I.M. Mackintosh, "The Electricial Characteristics of Silicon PNPN Triodes", Proceedings of the IRE, Vol. 46, June 1958, pp 1229-1235.

grid, that is, there is a spatial separation between control electrons and the electrons constituting the main current. In a transistor the electrons in the base control the flow of carriers from emitter to collector. In a PNP the base electrons are in the conduction band and control the flow of electrons in the valence band, while in a NPN the reverse is true. So the separation of control and current electrons exist in the transistor by an energy band separation. It is important to realize that any excess of electrons in the base region in the energy band of the base's majority carrier will enhance the emission of carriers from the emitter by a factor of

$$\frac{h_{FB}}{1 - h_{FB}} .$$

The excess of electrons may come about by any cause such as light radiation, base contact emission or heat. In the particular case of the PNPN the control electrons in the P_{2-} base and N_{1-} base are the non-recombining electrons emitted by the end P and N regions and as such constitute the main current in the PNP and NPN transistors suggested by J. J. Eber's equivalent circuit.

A block diagram of the PNPN may be constructed as depicted in figure 1-2, while a possible equivalent circuit is shown in figure 1-3, as suggested by J. J. Ebers. If the circuit is biased as shown in figure 1-3, J_1 and J_3 will be

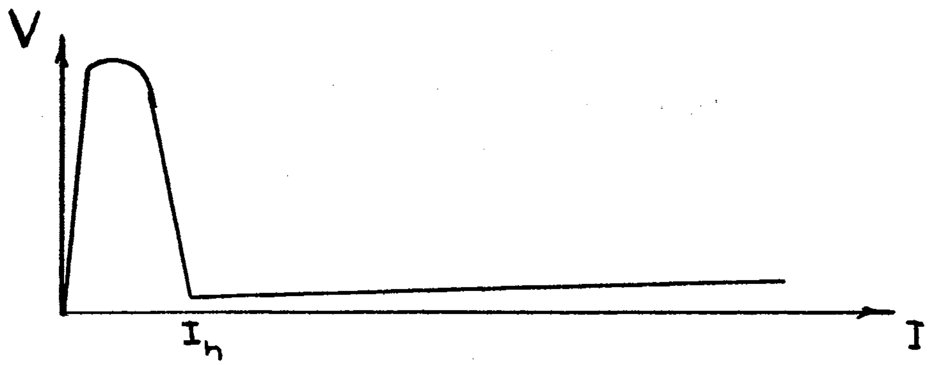

Figure 1-1. Characteristics of the PNPN

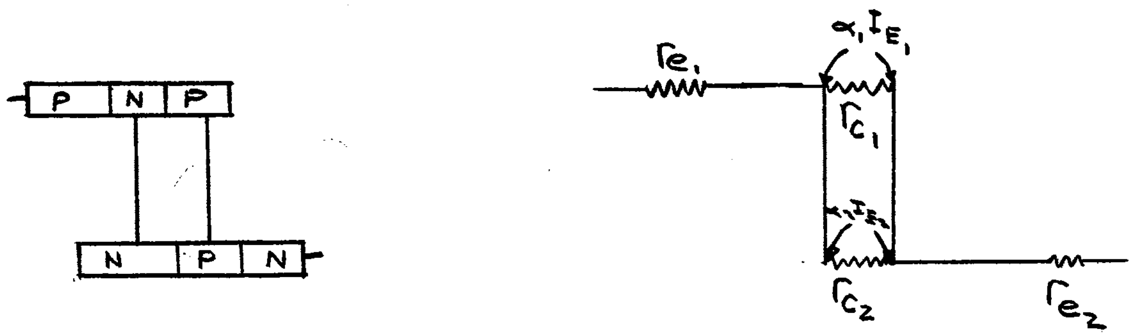

Figure 1-2. Equivalent Circuit of the PNPN

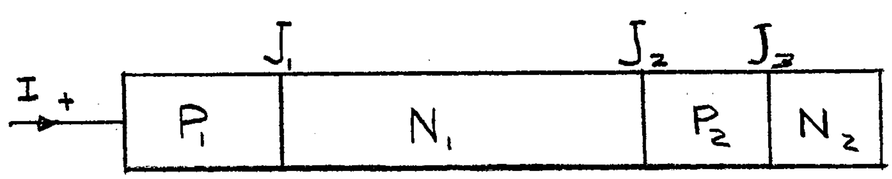

Figure 1-3. Junction Diagram

forward biased and J_2 will be reversed biased*. At this point the device will appear to be one reversed biased diode. Due to space charge generation in the reversed biased diode and junction leakage the current will continue to increase slightly as the voltage is increased. The holes injected by J_1 will diffuse across N_1, and will enhance an electron emission by the factor $$\frac{h_{FB}}{1 - h_{FB}} .$$

These holes will continue to cause an emission of electrons due to being caught in the potential hook. The electrons emitted from J_3 will in turn diffuse across P_2 and be collected by N_1. N_1 is the base of the PNP transistor, however, and these electrons in the conduction band will cause an enhanced hole flow in the valence band by a factor

$$\frac{h_{FB}}{1 - h_{FB}} .$$

As the voltage across J_2 is increased to avalanche magnitude the carriers crossing J_2 are multiplied due to the avalanche process by the factor M_n and M_p[4] for electrons and holes respectively. At some point in this regenerative process the magnitude of the negative electrons in the N_1 region will cause the positive charged depletion region to discharge and the opposite process will be accomplished on the P_2 side

*By forward bias it is meant the voltage existing across the junction is less than the equilibium voltage due to the shifting of Fermi levels to maintain a straight line. By reverse bias it is meant that existing voltage across the junction is greater than the equilibruim voltage.

of J_2. The state of affairs is such that now all three junctions are forward biased.

The above process may be stated analytically in the following manner. Again referring to figure 1-2.

$$I = M_p h_{FB_1} I_{j1} + M_n h_{FB_2} I_{j2} + I_{co} \tag{1}$$

$$I = \frac{I_{co}}{(1 - M_p h_{FB_1} - M_p h_{FB_2})}$$

If any factor is increased independantly in equation (1) the device may be switched to the "on" condition. This may be done by contacting a base lead to either base region; however, in the case considered, the base connection was made to P_2.

With the above theory in mind, expermentation on the device was began at this point. The first tests run on the device were with no base lead i.e., as a two terminal device. The anode voltage was slowly increased until the device switched to the on condition. This enabled, V_{bo}, the break over voltage of the device to be determined. This V_{bo} shall henceforth be referred to as the steady state break over voltage. A constant gate current of various magnitudes was then applied and V_{bo} was determined as a function of gate voltage and gate current.

In the process of evaluating the device parameters and determining its characteristics several noteworthy observations were made.

The first concerned the holding current. This is the minimum current the device will conduct in the "on" condition such that this current will maintain h_{FB} large enough to prevent the device from switching to the off or high impedance region. While in most units the current was a few microamps, it has been observed that some units require I_h as high as twenty mills.

Secondly, it was observed that the off impedance of at least 90% of the devices was in the order of megohms, while an occasional unit had a much lower impedance. The answer to this question appeared to be a dirty or leaky center junction since this junction has most of the reverse voltage across it in the off condition. This particular relatively low impedance device required a greater current in its emitter circuit before it switched on. This implies the entire unit is improperly cleaned and which in turn causes an appreciable surface current to exist. This surface current does not enhance more electron or hole current and it is not available in the interior to help saturate the traps and hence to increase the effective h_{FB}. This explains why a greater exterior current must exist in order to increase h_{FB} to the break over region.

Both of the above observations were of interest in this project for in considering this device as the heart of a relaxation type oscillator, turn-off and turn-on device parameters are of importance. If the device is to switch on and off, one possible means of turning the device off is to decrease the current below the holding current. Since this current varies from unit to unit then some type of compensation must be used to maintain a prescribed frequency. If some other means is used to turn the device off, such as a negative gate signal, then, of course, I_h variation is not critical. Variation in the emitter current necessary to switch the device on would also cause a variation in the free running frequency of this oscillator. In any circuit relying on an increasing emitter current to turn the device on, some type of stabilization would be necessary.

The third pertinent experimental result obtained concerned premature firing. By premature firing it is meant the device switches to the "on" condition at some gate or anode voltage of considerably smaller magnitude than the voltage indicated on the V-I characteristics. With a constant voltage on the gate, premature firing was observed when a step voltage, rather than a slowly increasing voltage, was applied to the anode. This indicated that the break over voltage of the device was a function of the time derivative of the applied voltage.

A series of experiments was performed to determine the cause and nature of this effect. The premature firing was a result of capacitive current flowing to charge the barrier layer of the reversed biased center junction. As h_{FB}

increases hyperbolically[5] with current, at some point the charging current increased h_{FB} sufficiently for the device to switch "on". The increase of h_{FB} with current is due to the saturation of traps[6]. The experiments were conducted and premature firings were observed at voltage sufficiently small in magnitude that the miltiplication of carriers was negligible. Hence, as indicated in expression (1), an increase in any factor would tend to switch the device to the low impedance state.

The voltage necessary to cause this type of premature firing is a function of the load resistance. For when the initial step is applied to the device and a resistance in series, the initial current is given by V/R. As was expected, for large values of R, no premature firing was observed. The magnitude of the static capacitance of the center junction was found to be in the neighborhood of twenty micro-micro farads. Also when the gate was tied to the emitter with a one-thousand ohm resistor, the step V_{bo} and the steady state V_{bo} approached the same value. This is understandable, as the gate resistor diverts current from the unit, preventing an increase in h_{FB}.

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[5] J.L. Moll, M. Tanenbaum, J.M. Goldey, and N. Holonyak, "PNPN Transistor Switches", Proceedings of the IRE, Vol.44, pp. 1174-1182, September 1956.

[6] W.Shockley and W.T. Read, Jr. "Statistics of the Recombinations of Holes and Electrons", Physical Review, Vol. 87, September 1952, pp. 835-842.

While the above consideration explained the process of premature firing it also suggested a problem. In determining the necessary conditions for premature firing, it was found that in some cases capacitive current could be an order of magnitude greater than I_h, and the device would not turn on. According to the V-I characteristics, the device should have been in the "on" condition, and if h_{FB} increased as predicted by the expression

$$h_{FB} = KI^{.5} \tag{1-a}$$

then equation (1) would indicate the device should have been in the "on" condition. The answer to this problem could be that h_{FB} does not increase according to (1-a) as this expression was derived for steady state condition i.e., the traps are in the condition indicated by the quasi-Fermi levels. Therefore before equation (1-a) can be considered to give a value for h_{FB} after the application of a step voltage, a finite time must elapse in order for the traps to reach a steady state value.

The results of the above series of experiments indicated premature firing of the device would have to be considered when employing the device in a circuit such that the anode voltage was not constant. In particular, this includes relaxation type pulse circuits. While an exact determination of the requirements on the step voltage, both in magnitude and duration, to cause premature firing would be necessary to entirely alleviate this problem, it was not done. The reason being that it would be necessary to consider the finite time required by the traps to reach an equilibruim state as indicated by the quasi-Fermi level. A problem of this type is beyond the scope of this work.

There is no need to consider the center junction charging current for slowly increasing anode voltage as the voltage across this junction follows the applied voltage and therefore there is no one surge of current. In fact, in using the PNPN in a circuit where it will be subjected to a dV/dt, the only requirement is that the time constant of the capacitance and device resistance be small compared to dV/dt. Due to the small value of the capacitance, this is not a stringent requirement.

CHAPTER II

CIRCUIT CONSIDERATIONS

The objective of the preliminary investigation was to determine the device parameters sufficiently to design a self excited temperature independant pulsing circuit using one PNPN device.

The first attempt to use the PNPN as a pulsing device was made by replacing the gas discharge tube in the classical saw tooth generating circuit with the PNPN. While this circuit could be made to operate with a particular unit, it would not operate in general nor was it temperature stable. Failure to cut off was the main problem encounted. This was due to the variation in I_h in the units over a range of micro amps to milliamps. Also the variation in circuit parameters would not allow the element to switch to the off state at elevated temperatures.

The above failure indicated a built-in shut off mechanism would have to be employed. This suggested the circuit shown in figure 2-2. The various wave forms generated by the circuit are illustrated in figure 2-3. Plate I shows the actual photographs of the generated waves.

Operation of Circuit

Basically the circuit operates in the following manner.

Capacitor C_1 charges to some voltage V_{bo} through R_1. At this time, t_1, the voltage across the unit and the current injected at the gate due to $R_3 - R_4$ is of the correct magnitude to cause the device to fire i.e., switch to the low impedance state. C_2 then begins to charge up to the voltage V_{bo} on C_1. C_1 is, however, discharging and hence V_{C_1} and V_{C_2} approach the same equilibruim voltage which indicates the current through the unit is becoming small. The voltage from gate to ground has been following the voltage from emitter to ground as they are in parallel due to the forward biased diode P_2N_2. As the current through the device decreases the P_2 N_2 diode becomes reversed biased and the gate voltage no longer follows the emitter voltage. When this happens the resultant gate to emitter voltage is negative and the device shuts off at time t_2. The two factors contributing to the switching of the device from the on to the off state are the negative gate voltage and the device voltage going to zero. The most important being the negative gate voltage.

In the following discussion of waveform generation two equivalent circuits have been used. The circuit of figure 2-4 represents the time during which the device is in the off state and this time has been designated as that interval from t_0 to t_1 and t_2 to t_3. Figure 2-5 depicts the circuit during the time t_1 to t_2 and is the on time of the device. As the pulse was only considered as a recurrent wave, time t_1 has arbitrarily been set as $t = 0$ in the calculations.

The device proper was considered as a simple switch, either opened or closed. No device transients were considered. Since rise and fall time of the pulse was of the same order of magnitude as the switching time of the PNPN device, the only improvement possible would be a faster switching PNPN unit. The key to the operation of the circuit is in determining the state of the P_1 N_2 diode and hence the correct equivalent circuit.

In order to clarify the description of the wave forms, the following constants were defined.

V_{dc} is the applied D.C. voltage.

V_{bo} is the voltage across the unit when it fires.

V_o is the voltage on C_2 when the unit swithces off.

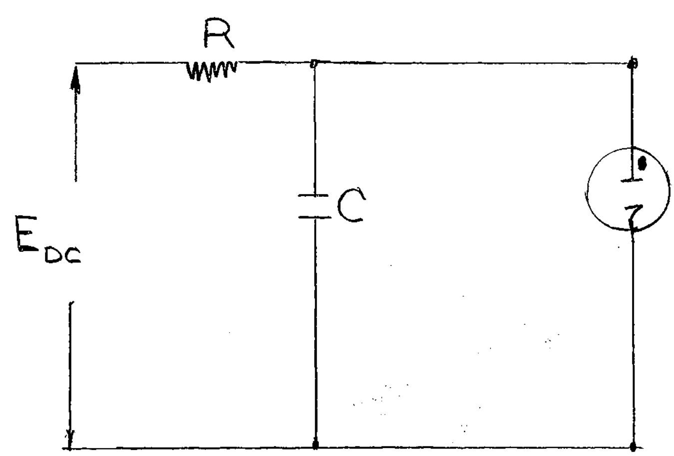


Figure 2-1. Classical Gas Discharge Sawtooth Generator

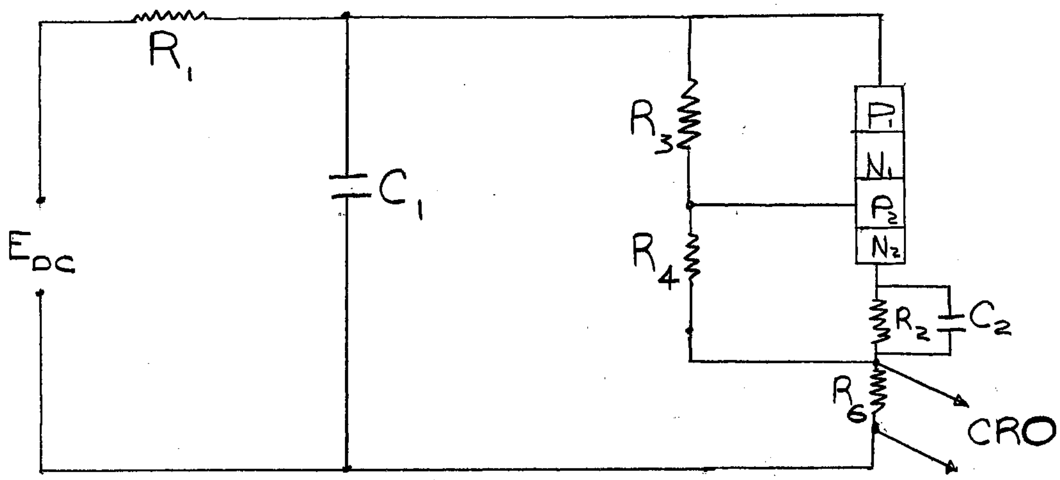


Figure 2-2. PNPN Pulse Generator with a Built-in Shut Off Mechanism.

Interpretation of Waveforms

The waveforms generated at various points in the circuit will be discussed next. Some of the waveforms will be discussed because of their possible applications, while others will be discussed in order to gain an understanding of the circuit. In particular, since the generation of a narrow pulse with a rise time in the order of several micro-seconds was the objective of this project the voltage across R_6 was of considerable importance. Furthermore, all mathematical derivations to follow concerns the pulse observed across R_6.

Inspection of figure 2-3C shows the voltage across R_6 rises to V_1, start to decay and then drop to zero. The rapid rise, .2 micro-seconds, is the result of the device switching to the on condition which effectively puts R_6 and C_1 in series as R_2 is shorted out by C_2 and hence a portion of the capacitor voltage must instantaneously appear across R_6. The voltage decays due to a decreasing current caused by the two capacitor voltages approaching the same value. The rapid fall, .2 micro-seconds, at t_2 is due to the device switching off, thereby reducing the current through R_6 and the voltage across R_6 to zero.

The wave-form at R_4 shown in figure 2-3E, may be explained by noting that at $t = t_1$ R_4 is also shorted by C_2 as it is shunted across R_2 and C_2 by the PN junction, which is a forward biased at $t = t_1$. As the current through the device approaches zero the PN diode's resistance increases so that R_4 is no longer in parallel with C_2 and hence V_{R4} falls below the increasing voltage on C_2. This effectively creastes a negative gate signal which shuts the device off. The voltage across R_4 then reduces to that DC value caused by the bleeder current through R_3.

The waveform shown in figure 2-3B, which is generated across C_2 and R_2 may be explained in the following manner. As the voltage across V_{R6} decreases, the voltage across C_2 is rising, hence the decay time of V_{R6} is the rise time of V_{C2}. Whenever the device switches to the "off" condition at t_2, the expotential decaying voltage is brought about by C_2 discharging through R_2.

The saw-tooth waveform shown in figure 2-3A is generated

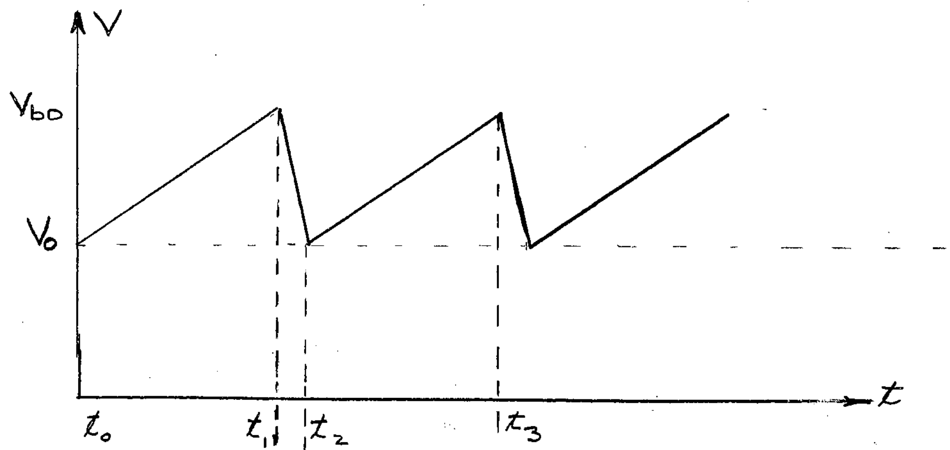


Figure 2-3A. Voltage from C_1 to Ground

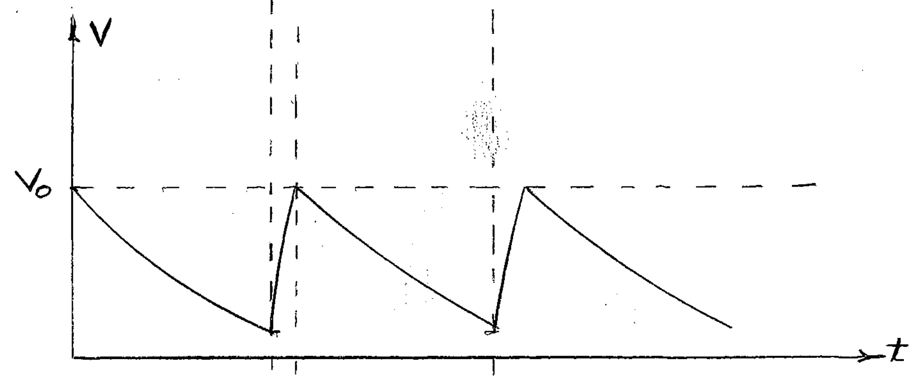


Figure 2-3B. Voltage Across C_2

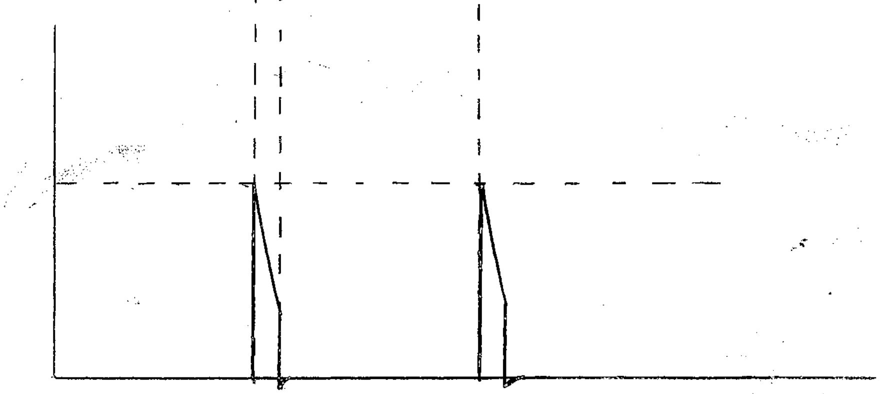


Figure 2-3C. Voltage from R_6 to Ground

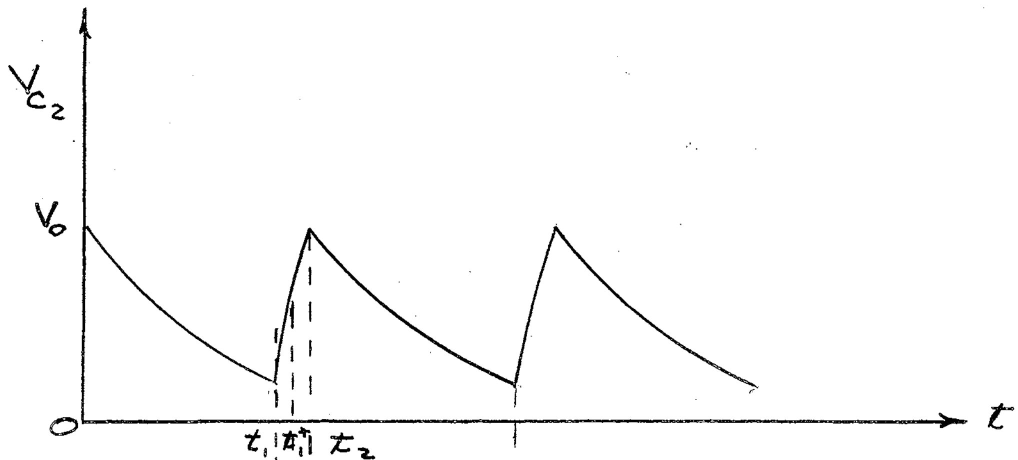


Figure 2-3D. Voltage Across C_2

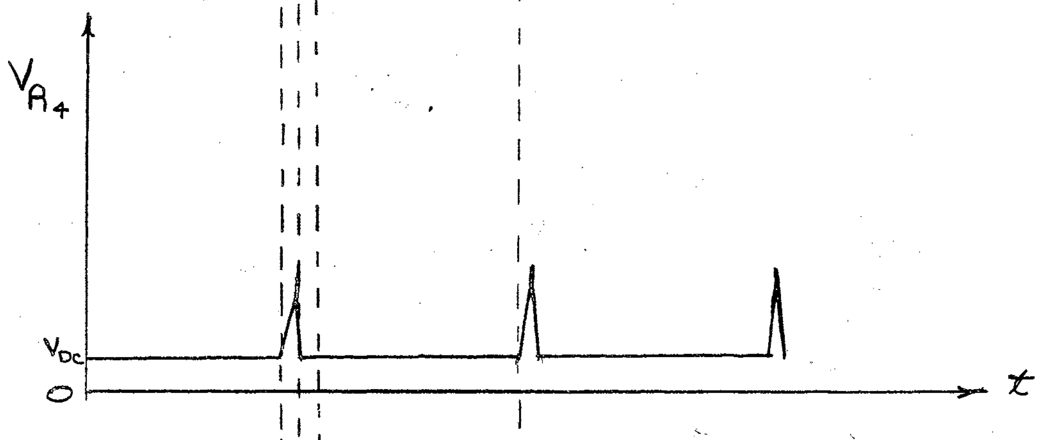


Figure 2-3E. Voltage Between Gate and Ground

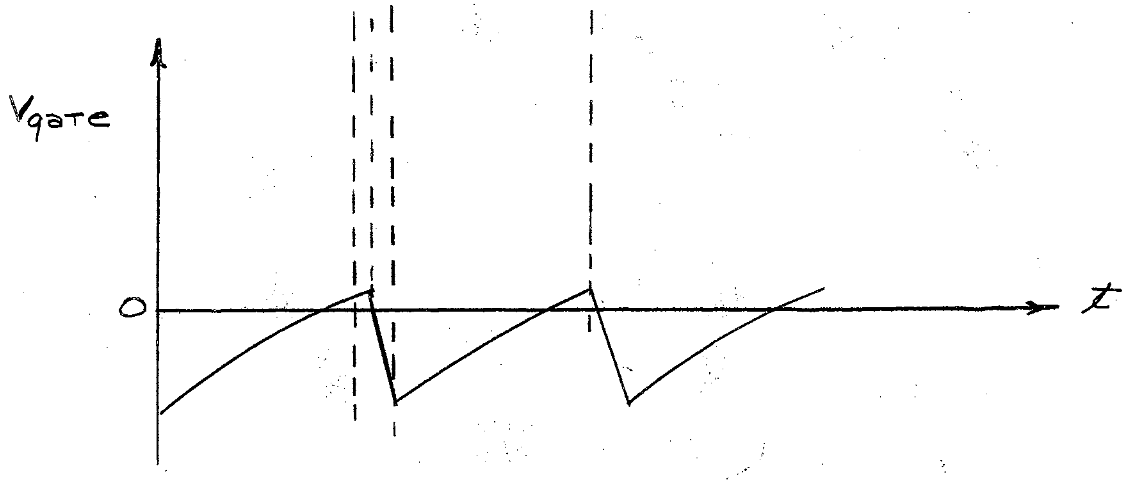
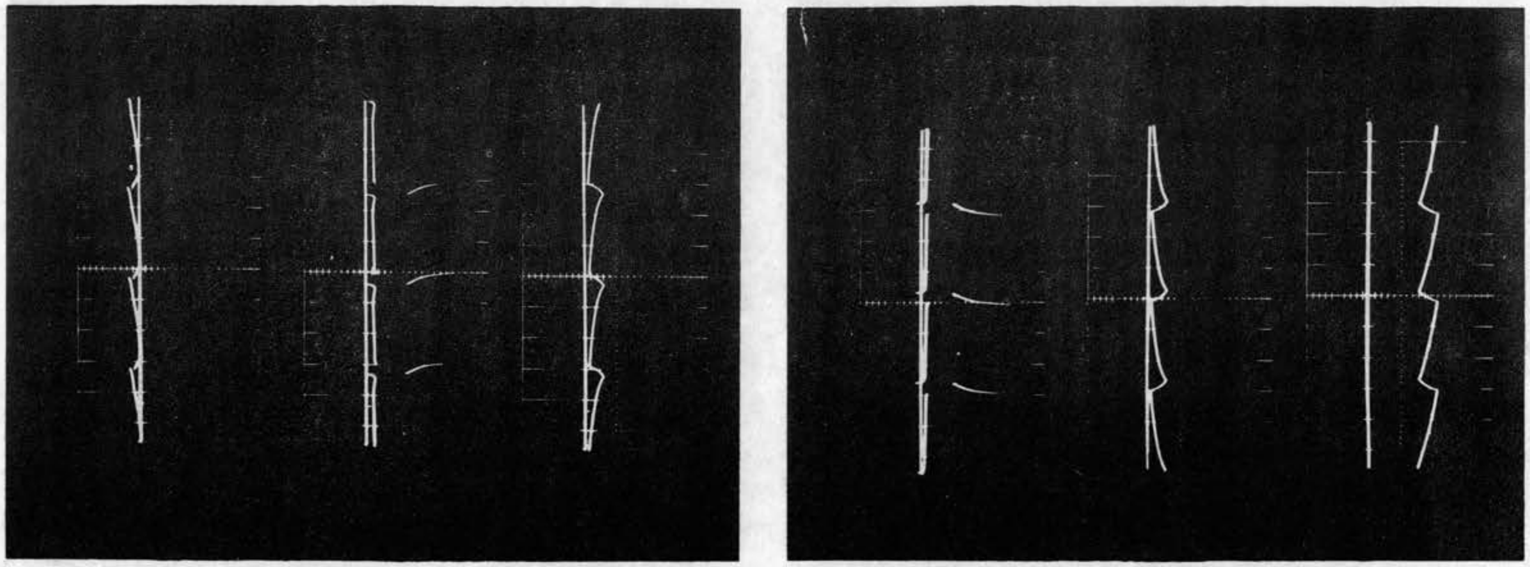


Figure 2-3F. Effective Voltage Existing Between Gate and Emitter, Which is the Difference Between V_{C_2} and V_{R_4}.

Plate I



Circuit waveform as described in
figure 2-3 a,b,c,d,e,f.

across capacitor C_1. The capacitor charges up to some voltage, V_{bo}, at which time, $t = t_1 = 0$, the unit fires. C_1 then discharges to some voltage, say V_0. The discharge curve is not an ordinary expotential for C_1 is discharging into another capacitor. This causes a much faster discharge. If R_6 is reduced sufficiently i.e., less than ten ohms, the discharge approaches a straight line with a fall time of ten micro-seconds. The charging curve can be made linear by making the applied voltage much greater than the V_{bo} of the device. While this wave was not worked with, nor were the circuit parameters varied to optimise this wave, it is believed that this circuit is capable of producing an ideal saw-tooth wave.

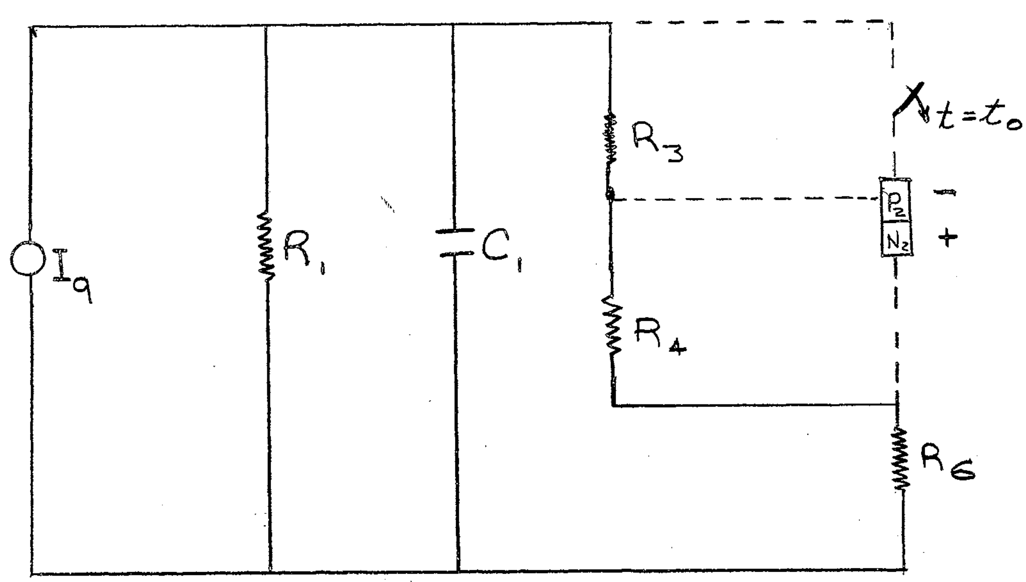


Figure 2-4. Constant Current Generator Equivalent Circuit for time $t_o < t < t_1$. (off time)

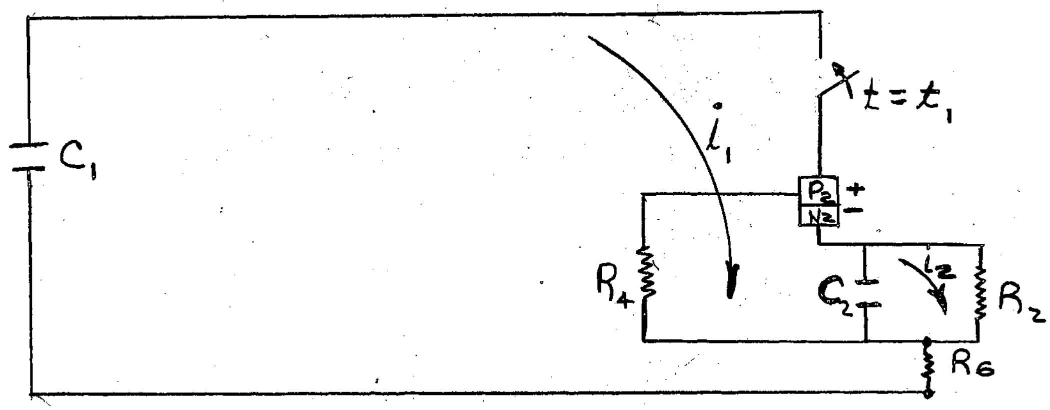


Figure 2-5. Equivalent Circuit
for time $t_1 < t < t_3$. (on time)

Time Constant Derivation

In the following derivations, expressions for three time constants were found. The first is an expression for pulse width at R_6. This is the time from t_1 to t_2. The other two constants concern the "off" time of the device.

Figure 2-5 depicts the equivalent circuit for time between t_1 and t_2. This is the "on" time of the device. For this time the diode P_2 N_2 is forward biased, effectively putting R_4 and R_2 in parallel. The actual device has been replaced with a switch which closes at $t = t_1$, this approximately shorts out R_3. If R_6 is much smaller than R_1 then the time it takes for the two capacitors to approach the same voltage is determined by R_6. Hence R_1 is omitted.

Writing loop equations for the circuit of Figure 2-5.

$$0 = (R_6 + R_c + 1/pC_1)\, i_1 - R_c i_2$$

$$0 = \qquad - R_c i_1 + (R_c + 1/pC_2)\, i_2$$

where $\quad R_c = \dfrac{R_2 R_4}{R_2 + R_4}$

Taking the Laplace of both equations and imposing the boundary condition $Q_0/C_1 = V_{bo}$ at $t = t_1$ there results

$$V_{bo}/s = (R_6 + R_c + 1/sC_1)\, I_1(s) - R_c I_2(s) \qquad (1)$$

$$0 = \qquad - R_c I_1(s) - (R_c + 1/sC_2)\, I_2(s). \qquad (2)$$

Solving (2) for $I_2(s)$ and substituting in (1) gives

$$V_{bo}/s = (R_6 + R_c + 1/C_1 s)\, I_1(s) - R_c \left[\frac{R_c}{R_c + 1/sC_2}\right] I_1(s).$$

$$I_1(s) = \frac{V_{bo}}{s}\left[\frac{1}{R_6 + R_c + 1/sC_1 - \dfrac{R_c^2}{R_c + 1/sC_2}}\right]$$

$$= \frac{V_{bo}}{s}\left[\frac{1}{\dfrac{sC_1R_6 + sC_1R_c + 1}{sC_1} - \dfrac{sC_2R_c^2}{R_c s C_2 + 1}}\right].$$

$$I_1(s) = \frac{V_{bo}}{s}\left[\frac{sC_1\,(sR_cC_2 + 1)}{s^2C_1C_2R_6R_c + s^2C_1C_2R_c{}^2 - s^2C_1C_2R_c{}^2 + sR_cC_2 + s(C_1R_6 + C_1R_c) + 1}\right]$$

$$I_1(s) = \frac{V_{bo}}{s}\left[\frac{C_2C_1R_c(s + 1/C_2R_c)}{\left(s^2 + s\frac{(R_cC_2 + C_1R_6 + C_1R_c)}{(C_1C_2R_cR_6)} + \frac{1}{C_1C_2R_6R_c}\right)\frac{C_1C_2R_6R_c}{1}}\right].$$

The denominator factors into

$$s = -\frac{R_cC_2 + C_1R_6 + C_1R_c}{2C_1C_2R_6R_c} \pm \left[\frac{(R_cC_2 + C_1R_6 + C_1R_c)^2}{(2C_1C_2R_6R_c)^2} - \frac{4}{2C_1C_2R_6R_c}\right]^{.5}.$$

Letting A and B equal the factors, there results

$$I_1(s) = \frac{V_{bo}}{R_6}\left[\frac{(s + \frac{1}{C_2R_c})}{\left[s_1 + (A - B)\right]\left[s_2 + (A + B)\right]}\right].$$

Taking the inverse Laplace of both sides

$$i_1(t) = \frac{V_{bo}}{R_6}\left[\frac{\frac{1}{C_2R_c} - (A - B)\ e^{-(A - B)t}}{(A + B) - (A + B)} + \frac{-\ 1/C_2R_c - (A + B)\ \ e^{-(A + B)t}}{(A + B) - (A + B)}\right].$$

$$i_1(t) = \frac{V_{bo}}{R_6}\left[\frac{(1/C_2R_c - A + B)e^{-(A - B)t}}{B}\right.$$

$$\left. + \frac{-(1/C_2R_c - A - B)e^{-(A + B)t}}{B}\right].$$

If the above equation is substituted into (2) and solving for $i_2(t)$, the expression is

$$i_2(t) = \frac{V_{bo}}{2R_6}\left[\frac{(A + B)e^{-(A + B)t} - (A - B)e^{-(A - B)t}}{B}\right] . \quad (3)$$

As the pulse width at R_6 is terminated when $i_2(t)$ goes to zero an expression for this time may be found by setting (4) equal to zero.

$$\frac{V_{bo}}{2R_6}\left[\frac{(A + B)e^{-(A + B)t} - (A - B)e^{-(A - B)t}}{}\right] = 0$$

$$(A + B)e^{-(A + B)t} - (A - B)e^{-(A - B)t} = 0.$$

Multiplying through by $e^{(A + B)t}$ gives

$$(A + B) - (A - B)e^{2Bt} = 0$$

$$e^{2Bt} = \frac{A + B}{A - B} .$$

Taking the ln of both sides

$$t = \frac{1}{B2} \ln \frac{A + B}{A - B} . \qquad (4)$$

By considering the ln term and substituting the values for A and B there results

$$\frac{A + B}{A - B} =$$

$$\frac{\dfrac{(R_cC_2 + R_6C_1 + R_cC_1)}{(C_1C_2R_6R_c)} - \left[\dfrac{(R_cC_2 + C_1R_6 + C_1R_c)^2}{(C_1C_2R_cR_6)^2} - \dfrac{4}{C_1C_2R_cR_6}\right]^{.5}}{\dfrac{(R_cC_2 + R_6C_1 + R_cC_1)}{(C_1C_2R_6R_c)} + \left[\dfrac{(R_cC_2 + C_1R_6 + C_1R_c)^2}{(C_1C_2R_cR_6)^2} - \dfrac{4}{C_1C_2R_cR_6}\right]^{.5}}.$$

Finding a common denominator

$$\frac{A + B}{A - B} =$$

$$\frac{\dfrac{R_cC_2 + R_6C_1 + R_cC_1 - \left[(R_cC_2 + C_1R_6 + C_1R_c)^2 - 4C_1C_2R_6R_c\right]^{.5}}{\cancel{C_1C_2R_cR_6}}}{\dfrac{R_cC_2 + R_6C_1 + R_c\,C_1 + \left[(R_cC_2 + C_1R_6 + C_1R_c)^2 - 4C_1C_2R_6R_c\right]^{.5}}{\cancel{C_1C_2R_cR_6}}} =$$

$$= \frac{R_cC_2 + R_6C_1 + R_cC_1 - \left[(R_cC_2 + R_6C_1 + R_cC_1)^2 - 4C_1C_2R_6R_c\right]^{.5}}{R_cC_2 + R_6C_1 + R_cC_1 + \left[(R_cC_2 + R_6C_1 + R_cC_1)^2 - 4C_1C_2R_6R_c\right]^{.5}}.$$

Dividing numerator and denominator by $R_cC_2 + R_6C_1 + R_2C_1$

$$\frac{A + B}{A - B} = \frac{1 - \left[1 - \dfrac{4C_1C_2R_cR_6}{(R_cC_2 + C_1(R_c + R_6))^2}\right]^{.5}}{1 + \left[1 - \dfrac{4C_1C_2R_cR_6}{(R_cC_2 + C_1(R_c + R_6))^2}\right]^{.5}} .$$

Since $R_c \gg R_6$ $(R_c + R_6) \approx R_c$, then

$$\frac{A+B}{A-B} = \frac{1 - \left[1 - \dfrac{4C_1C_2R_cR_6}{(R_cC_2 + C_1R_c)^2}\right]^{.5}}{1 + \left[1 - \dfrac{4C_1C_2R_cR_6}{(R_cC_2 + C_1R_c)^2}\right]^{.5}}$$

$$= \frac{1 - \left[1 - \dfrac{R_cR_6C_1C_24}{R^2_c\ (C_1 + C_2)^2}\right]^{.5}}{1 + \left[1 - \dfrac{R_cR_6C_1C_24}{R^2_c\ (C_1 + C_2)^2}\right]^{.5}}.$$

It is easily shown that $C_1C_2 < (C_1 + C_2)^2$ and $R_6 < R_c$, hence the parameter term is less than one and therefore the square root terms will be represented everywhere by a series expansion obtained by the binomial theorm. Expanding, there results,

$$\frac{A+B}{A-B} = \frac{1 - \left[1 + \dfrac{(-1/2)4R_6C_1C_2}{R_c\ (C_1 + C_2)^2} + \ldots +\right]}{1 + \left[1 + \dfrac{(-1/2)4R_6C_1C_2}{R_c\ (C_1 + C_2)^2} + \ldots +\right]}.$$

Using only non-squared terms

$$\frac{A+B}{A-B} = \frac{1 - \left[1 - \dfrac{2\,R_6C_1C_2}{R_c\,(C_1 + C_2)}\right]}{1 + \left[1 - \dfrac{2\,R_6C_1C_2}{R_c\,(C_1 + C_2)^2}\right]} = \frac{\left[\dfrac{2\,R_6C_1C_2}{R_c\,(C_1 + C_2)^2}\right]}{\left[\dfrac{2 - 2R_6C_1C_2}{R_c\,(C_1 + C_2)^2}\right]} =$$

$$\frac{\dfrac{R_6C_1C_2}{R_c(C_1 + C_2)^2}}{\dfrac{R_c(C_1 + C_2)^2 - R_6C_1C_2}{R_c(C_1 + C_2)^2}} = \frac{R_6C_1C_2}{R_c(C_1 + C_2)^2 + R_6C_1C_2}.$$

Dividing numerator and denominator by $R_6C_1C_2$

$$\frac{A+B}{A-B} \quad \frac{1}{\dfrac{R_c(C_1 + C_2)^2 - 1}{R_6C_1C_2}} = \frac{R_6C_1C_2}{R_c(C_1^2 + C_2^2) + C_1C_2(2R_c + R_6)} =$$

$$\frac{R_6}{R_c\left[\dfrac{C_1^2 + C_2^2}{C_1C_2}\right] + 1}.$$

Since $R_c \gg R_6$ and $(C_1^2 + C_2^2) \gg C_1C_2$ and as it is the ln that is of interest, there results

$$\ln\frac{A+B}{A-B} \approx \ln\frac{R_6C_1C_2}{R_c(C_1^2 + C_2^2)}.$$

In order to completely evaluate (4), the B term will be considered next. B is given by

$$B = -\frac{1}{2}\left[\frac{\left(C_2R_2 + C_1R_c + C_2R_6\right)^2}{\left(R_1R_2R_6R_c\right)^2} - \frac{4}{C_1C_2R_cR_6}\right]^{.5}$$

Making the same approximation as before $R_c \ggg R_6$ i.e., $R_c \approx 2.4K$ and $R_6 \approxeq 22$ ohm.

$$B = -\frac{1}{2}\left[\left(\frac{C_1R_c + C_2(R_c + R_6)}{C_1C_2R_6R_c}\right)^2 - \frac{4}{C_1C_2R_cR_6}\right]^{.5} =$$

$$= -\frac{1}{2}\left[\frac{R_c{}^2(C_1 + C_2)^2}{R_c{}^2R_6{}^2(C_1C_2)^2} - \frac{4}{C_1C_2R_cR_6}\right]^{.5}$$

$$B = -\frac{1}{2}\left[\frac{(C_1 + C_2)^2}{R_6(C_1C_2)} - \frac{4}{R_c}\,\frac{1}{R_6C_1C_2}\right]^{.5}.$$

$R_c \gg 4$, then

$$B \approxeq -\frac{1}{2}\ \frac{(C_1 + C_2)^2}{R_6{}^2(C_1C_2)^2} \approxeq -\frac{1}{2}\ \frac{C_1 + C_2}{R_6(C_1C_2)}\ .$$

Equation number (4) becomes

$$t = \frac{1}{2B}\ \ln\frac{A + B}{A - B} \approxeq -\frac{R_6(C_1C_2)}{(C_1 + C_2)}\ \ln\frac{R_6C_1C_2}{R_c(C_1{}^2 + C_2{}^2)}\ . \quad (5)$$

This expression gives the approximate pulse width, which is primarily determined by R_6, C_1 and C_2. The term containing R_c is a ln term and it does not have nearly the effect on t as R_6. Also R_c is a parallel combination and the variation of either of its parts does not cause as large a change in it.

Detailed consideration must be given to the time from t_2 to t_3, for when the "on" time, i.e., pulse width, is small compared to the "off" time, the period of oscillation, say T, is essentially this time. There are two time constants effecting this period. The most important is the duration of the negative pulse on the gate and the other is the R_1 C_1 time constant.

The negative voltage observed between the gate and emitter is the difference in voltage across R_4 and C_2. V_{R4} rises as quickly as V_{C2}, but drops to a DC value whenever the P_2 N_2 junction becomes reverse biased. Whereas V_{C2} decays expotentially through R_2 maintaining a negative potential on the gate until the decaying voltage of R_2 is less than the voltage drop across R_4 due to the bleeder current. If the duration of the negative pulse is longer than the charging time of R_1 C_1, the period T is primarily determined by R_2 C_2. This is understandable for the device will not turn on for a given anode voltage if there is a negative signal on the gate. For the other case when R_1 C_1 is larger than R_2 C_2 the period of oscillation is determined by both

$R_1 C_1$, $R_2 C_2$ and of course the magnitude of the DC voltage on R_4. It is well to note that for the former case, the device sees a constant voltage during the period $t = t_2$ to t_3 and hence R_1 and C_2 may be removed and the circuit will continue to operate, although at a slightly different frequency.

In order to compute the charging time of C_1 the equivalent circuit of figure 2-4 was used. The charging of C_1 from V_0 to V_{bo} is of course the "off" time of the device.

In this circuit R_2 and C_2 were omitted for they shunt R_4 through the P_2 N_2 diode which is reversed biased during the off period and are effectively out of the circuit.

Let $R_a = R_3 + R_4 + R_6$ and $I_g = E/R_1 = EG_1$.

Where $G_a = 1/R_a$ and $G_1 = 1/R_1$.

Boundary conditions are

$$V_{C_1} = V_0 \text{ at } t = t_0.$$

Writing a loop equation

$$I_g = VG_1 + C_1 \frac{dv}{dt} + V_1 G_a$$

$$EG_1 = VG_1 + C_1 \frac{dv}{dt} + V_1 G_a .$$

Separating the variables and integrating

$$\int dt = t = \int \frac{C_1 dv}{V (G_1 + G_a) - EG_1} + K_1 .$$

$$t = -\left(\frac{C}{G_1 + G_a}\right) \ln V(G_1 + G_a) - EG_1 + K_1$$

Evaluating constants at $t = t_0 = 0$

$$K_1 = \left(\frac{C_1}{G_1 + G_a}\right) \ln V_0(G_1 + G_2) - EG_1$$

$$t = -\left(\frac{C_1}{G_1 + G_a}\right) \Big[\ln V(G_1 + G_a) - EG_1 - \\ - \ln V_0(G_1 + G_a) - EG_1\Big].$$

Raising both sides to e, there results

$$\exp - t\left(\frac{G_1 + G_a}{C_1}\right) = \frac{V(G_1 + G_a) - EG_1}{V_0(G_1 + G_a) - EG_1} .$$

Solving for V (which is V_{C_1}) gives

$$V_{C_1} = \frac{EG_1}{G_1 + G_a} + \frac{V_0(G_1 + G_a) - EG_1}{G_1 + G_a} \exp - \frac{t(G_1 + G_a)}{C_1} .$$

The time constant is then given by

$$\tau = \frac{C_1}{G_1 + G_a} . \qquad (6)$$

If the time constant given by (6) is small compared to $R_2 C_2$, then T will be determined by the length of time the negative pulse is maintained on the gate. In order to determine this time it is necessary to know the voltage capacitor

C_2 charged up to before the device switched off. This voltage is approximately V_0, and is given by

$$V_0 = q_2(t)/C_2 .$$

$q_2(t)$ may be found by integrating $i_2(t)$, expression (3), with respect to time. The expression for $V_0(t)$ is then

$$V_0 = \frac{V_{bo}}{2BC_2R_6}\left[e^{-(A - B)t} - e^{-(A + B)t}\right]. \qquad (7a)$$

As $B \approx \frac{C_1 + C_2}{2R_6C_1C_2}$, (7a) becomes

$$V_0 \approx \frac{V_{bo}C_1}{C_1 + C_2}\left[e^{-(A + B)t} - e^{-(A - B)t}\right]. \qquad (7b)$$

By substituting the value of time given by (5), V_0 can be determined. When the device switches off the voltage will simply decay expotentially as C_2 discharges through R_2. When this decaying voltage is equal to the steady DC voltage on R_4 given by

$$E_{DC} = \frac{E \text{ applied}}{R_3 + R_4}\,\frac{R_4}{1}$$

the device will switch on. Equating the discharging voltage of R_2 to $E_{DC}(R_4)$ and solving for t, there results

$$t = - R_2C_2 \ln \frac{E \text{ applied}}{(R_3 + R_4)}\,\frac{R_4}{V_0} . \qquad (8)$$

This t is the period of oscillation, say T. This gives the "off" time of the device.

Parameter Variation and Stability Consideration

The pulse generated across R_6 may be varied in a number of ways. The two variable time constants are R_1 C_1 and R_2 C_2. Variation of other circuit parameters will also cause a frequency shift, but these parameters are usually fixed due to imposing circuit operating conditions.

The frequency of oscillation was varied by changing R_2 and C_2. Changing R_2 from ten thousand ohms to one thousand ohms varied the frequency from one half megacycle to one thousand cycles. While this was being done C_1 and C_2 were both one microfarad. The frequency was lowered to power frequencies by increasing C_2 to four microfarads. Changing R_2 over the above range of resistance did not change the magnitude of the pulse, nor did this variation cause a change in pulse width. The addition of the capacitance at C_2 did however, increase the width of the pulse, although its magnitude did not change.

Varying R_1 also changed the frequency, but only around a central frequency determined by R_2. Varying R_1 caused the magnitude of the pulse at R_6 to change slightly and its width to vary slightly. This variation was not indicated in the mathmetical expression, but that was as expected. The derivation was carried out under the assumption R_1 was of sufficient magnitude that its effect on pulse width would be small compared to other parameters. This was found to be true.

Of the above two methods of frequency control, variation of R_2 was by far the most effective and also its alteration of the wave shape was negligible. Although the circuit performed without C_1 and R_1, greater stability was obtained with these two elements in the circuit at high voltage.

The resistor, R_6, determine pulse width, as may be seen by inspection of equation (5). While this expression is not exact, it was found to agree within an order of magnitude with experimental data. It is certainly a good first approximation as device parameters also varied this much. Changing R_6 did not have an appreciable effect of frequency, and hence is an excellent means of pulse width control. Variatiom of R_6 from eleven ohms to two-hundred fifty ohms varied the pulse width from four micro-seconds to five-hundred micro-seconds.

The purpose of R_3 is twofold. If the device is to operate in a circuit where the applied voltage is less than the breakover voltage of the device, gate current injection must be used. R_3 furnishes a path for this current. Secondly, the R_3 and R_4 combination maintains self-bias on the gate which varies with temperature. This creates a certain degree of temperature stability. Variation of R_4 causes a change in pulse width and in frequency. Hence, it does not furnish a desirable independant control over either and is usually fixed by the magnitude of voltage the device is to operate on.

While it appears that the magnitudes of R_3, R_4 and R_2 might be chosen independantly, this was not found to be true. Several cases of why this should not be true will now be discussed.

In the process of turning the device on there will be a current at the node of the parallel combination of R_4 and the gate P_2 N_2, determined by R_3 and the voltage between the gate and anode. The amount of current entering the gate will be inversely proportional to the size of R_4. If R_4 is too small the device will not turn on and if R_4 is too large the device will turn on and fail to turn off. Hence there is a definite relationship between these parameters. For low voltage, it was found that the difference in magnitudes of R_3 and R_4 should not differ over ten thousand ohms. For high voltages, a difference up to fifty thousand ohms did not prevent operation.

If R_2 is too small, it will allow too much current to flow in the circuit when the device is trying to cut off. This will prevent the P_2 N_2 diode from becoming reverse biased and there will then be no negative gate signal; hence the device will remain in the on state.

Synchronazation

The circuit was found to be readily converted from a free running pulse generator to a synchronized generator.

Figure 2-6 shows the DC voltage on R_4 and the decaying voltage of C_2 superimposed on one another. Of course the decaying voltage appears at the gate as a negative voltage.

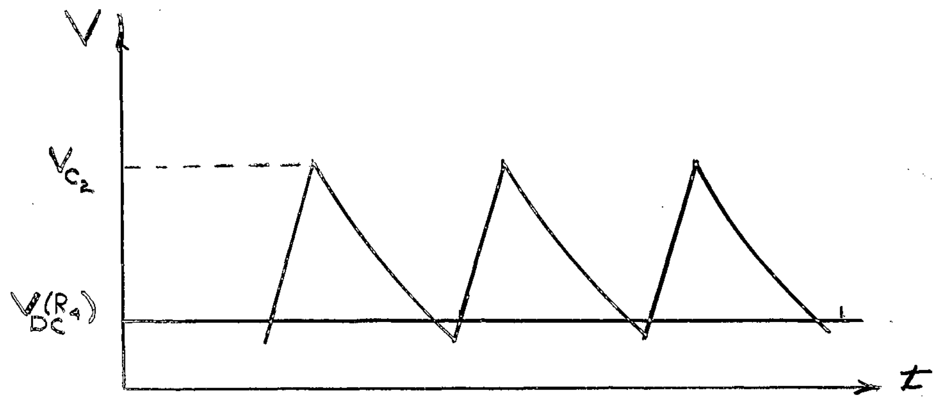


Figure 2-6. Voltage between Gate and Emitter

As it is the difference in these two voltages that is effectively on the gate, whenever the two voltages are of the same magnitude the device turns on, for there will then be zero or positive voltage on the gate. It is easily seen that by increasing the voltage on R_4, the device would fire sooner. This was done by superimposing a variable frequency A.C. signal on R_4. The circuit, designed for power frequency, synchronized from sixty cycles up to eight hundred cycles without changing any parameters. The necessary condition for this type synchronization is that the free running frequency must be lower than the applied signal. The variation in frequency was zero for a temperature range from 25° centigrade to 160°.

In general synchronization can be accomplished by periodically causing the gate and emitter voltages to become equal. This in turn, may be accomplished by either increasing the voltage on R_4, as in the above case, or by decreasing the voltage on R_2. The voltage on R_2 may be decreased quickly by imposing a negative step voltage or by creating a fast discharge path for C_2, such as a switching transistor in the on state shunted across C_2.

Temperature Stability

The importance of R_3 for temperature stability should not be minimized. Without R_3 in the circuit the repetition rate of the pulse generator was found to decrease by a factor of one half while changing the ambient from 25° C to 90°. In changing the ambient from 90° to 130° the repetition rate increased by a factor of two with respect to the rate at 25°. No concrete explanation was found for this phenomona. With R_3 in the circuit the percent variation in repetition rate was found to be a function of DC operating voltage. In changing the ambient from 25° C to 160° C the variation was from 1% at 280 volts to 10% at 24 volts. No particular ratios of R_3 and R_4 seem to decrease frequency variation.

The circuit, with parameters chosen for operation at 25° C, was also subjected to temperatures of -70° without malfunction. However, for large values of R_2 and small values of the operating DC voltage e.g. 24 volts, the circuit failed to operate at -90° C. This failure was due to the PNPN failing to turn on. As R_2 determines the current through the device, which determines h_{FB}, and since h_{FB} is temperature dependant, then evidently h_{FB} did not increase sufficiently to turn the device on. This operation failure was encountered at low frequencies whenever R_2 had been chosen to be a large value to obtain a low repetition rate. The same low repetition rate may also beachieved by selecting a larger C_2 and consequently a small R_2. This will give continuous operation down to -90° C.

Oscillation Stability

In any device exibiting a negative resistance region in its V-I characteristics, oscillation stability must be considered. Modulated one megacycle oscillations were observed in the wave form of the pulse generator of this project for a pulse repetition rate of one thousand. The sinusoidal oscillation was observed across R_6 and it was modulated by the wave shape of figure 2-3c. A thirty-three millimicro farad capacitor between gate and anode shorted this oscillation out without interfering with the operation of the circuit. For higher repetition rates, no oscillations were observed and it is believed that for the particular case discussed the circuit parameters were so selected that the load line intersected the negative resistance region at only one point. As the negative resistance region of the curve is almost a verticle line, an intersection with the load line at only one point is difficult to achieve. Also the parameter values to give this type of intersection would surely be different for two different four layer diodes.

The generated pulse depicted in figure 2-3c had a tendency to overshoot or oscillate on the trailing edge. This particular trouble was traced to the stray inductance inherant in the slidewire resistors used in building the pilot circuit. This trouble was stopped by using fixed resistors. Stray inductance in R_2 proved to be the most critical. Various values of inductance were purposely shunted across R_2 causing various damped sinusoidal oscillations. This particular type of operation was undesirable and was considered no further. It does indicate certain possible applications for a modification of the pulse generator considered here.

Design Considerations

The deciding factor in selecting R_3 and R_4 is the biasing of the PNPN in its operating range for a given voltage. The size of R_3 and R_4 should be such that the voltage on the gate is of sufficient magnitude to trigger the device with the given voltage on the anode. This may be determined from the V-I characteristics. As noted, the relative sizes of R_3 and R_4 must be considered in order for the device to switch "on" and "off" properly. This selection is more critical in the low voltage range.

While the circuit will operate without R_1 and C_1, their inclusion creates a better pulsing circuit. Stability is increased in the high voltage range and whenever the load current is appreciable. Also R_1 may be used to regulate the cut-off time of the device. With R_1 in the circuit a cut-off time of .2 micro-seconds was obtained as compared to the predicted time of 8 micro-seconds. The time constant of R_1 and C_1 should be small compared to the desired pulse repetition rate in order to simplify parameter selection.

The only requirement on R_6 is that it be of proper size to give the desired pulse width in accordance with equation (5). It should also be noted that R_6 should be small in comparison to the input resistance of following stage. This will insure frequency and pulse width independant of load.

For a given pulse repetition rate, equation (8) will determine R_2 and C_2, as the other variables in the expression have been fixed. If the pulse width is appreciable to the period of oscillation, then it must be added to this expression.

In order to test the validity of the above design equations, a circuit will be designed to operate on a DC voltage of 24 volts, with a pulse repetition rate of one thousand and a pulse width of 20 micro-seconds.

Design Procedure

R_1 and C_1 will arbitrarily be selected as one thousand ohms and one microfarad, with the condition in mind that the time constant of $R_1 C_1$ should be smaller than the period of operation. R_1 must also be much larger than R_6 so the equivalent circuit of figure 2-5 will be justified. It is well to note these two conditions are contradictory and a compromise must be made.

Using the V-I characteristics the values of R_3 and R_4 were determined to be 1.8K and 4.5 K in order that the unit will be biased at the specified operating DC voltage.

As the design equations contain R_2, C_2 and R_6 in expotential form, it is not possible to solve explicitly for these parameters. C_2 may arbitrarily be set at one microfarad since this leaves the values of R_2 and R_6 still free to be determined in accordance with the operating conditions i.e., pulse repetition rate and pulse width.

In order to determine R_2 equation (8) will be used. However, (8) contains V_0 which must be computed from (7). In order to evaluate (7), the two terms in the expotential will be investigated viz., A and B. As a reduced form for B has already been determined, a reduced expression, using the same approximation, will be derived for A. The complete expression for A is

$$\frac{1}{2} \quad \frac{(R_c + R_6)C_1 + R_cC_2}{C_1C_2R_6R_c} \quad .$$

But $R_c >> R_6$, then

$$A \approx \frac{1}{2} \quad \frac{R_c(C_1 + C_2)}{R_c(C_1C_2)R_6} \quad =$$

$$\approx \frac{C_1 + C_2}{2C_1C_2R_6} \quad .$$

The reduced values for A and B were not used in evaluating the pulse width for, physically, this ignores R_6, which is used to determine pulse width. However, in computing the period of the pulse which is long compared to pulse width, or in determining R_2, this is justifiable.

Using the reduced values of A and B and noting A = B, equation 7b becomes

$$V_0 \approx \frac{(V_{bo})\ (C_1)}{C_1 + C_2} \quad .$$

Substituting the values of the parameters

$$V_0 \approx \frac{(24)\ (1 \times 10^6)}{2 \times 10^6} = 12.$$

Solving (8) for R_2 and substituting the values of V_o, t, R_3 and R_4, there results

$$R_2 = \frac{t\ (\text{period of operation})}{C_2\ \ln\dfrac{(R_3 + R_4)V_0}{(E_{applied})\ R_4}} =$$

$$= \frac{1 \times 10^{-3}}{1 \times 10^{-6}\ \dfrac{(1800 + 4500)12}{(24)\ (4500)}}$$

$= 5500$ ohms.

(5) will be used to determine R_6. Solving for R_6

$$R_6 = \frac{t\ (\text{pulse width})}{\dfrac{C_1C_2}{C_1 + C_2}\ \ln\dfrac{R_6C_1C_2}{C_1^2 + C_2^2}}.$$

This is a trancendental equation which may be solved by successive approximation. However, a first approximation may be made by noting, from experimentation, that whenever R_6 is increased by 3/4 of an ohm the pulse width is increased by one micro-second. Sixteen ohms will be used.

In summary the parameters are

$R_1 = 1.000$ K	$R_4 = 4.500$ K
$C_1 = 1$ microfarad	$R_6 = 16$ ohms
$R_3 = 1.800$ K	$C_2 = 1$ micro-farad

A circuit was built using the above parameters and its operation observed. Listed below is a table of the specified and observed wave parameters of the generated wave.

TABLE I

GENERATED WAVE AT R_6

Wave Parameter	Observed	Specified or Calculated	% difference
pulse width	18 useconds	20	10 %
pulse rep rate	1045	1000	4.3 %
V_0	18 volts	12 volts*	33 %
V_{bo}	20	24 24 volts*	16.8 %

* calculated

The percent difference in the observed and specified values could have been due to the comprise in the selection of R_1, as C_1 did not have time to charge up to the full 24 volts before the PNPN fired. Thus, an error was introduced into the calculations.

Using a T I 110 PNPN this circuit operated successfully from -70° C to 160° C. Frequency variation over this ambient range was 15%.

For low frequencies and heavy current conduction the expression for pulse width predicted a much shorter pulse than the one actually observed because the finite time turn-off was not considered. Various other parameters were tried and in general the expressions proved to be dependable.

CHAPTER III

SUMMARY AND CONCLUSIONS

The goal of this work was to develope a temperature independant pulse generator utilizing a four layer diode, which could operate either in a free running or synchronized state. The circuit must be compatible with any unit selected from stock.

Preliminary investigation of the device established the device parameters and their variation, with respect to change in ambient and from unit to unit. Different values of I_h were found for various units and variation in the necessary emitter current to turn the device on was noted. It was also found that a step anode voltage of smaller magnitude than that indicated by the V-I characteristics would turn the device on.

In order to make the circuit independant of the variation in I_h the circuit was not designed to rely on a decreasing I_h to turn off. Instead a built in shut off mechanicism was employed to create a negative voltage at the gate. Fortunately, this same combination creates a positive voltage at the gate to turn the device "on" later in the cycle. Therefore the operation of the circuit is independant of variation in the above mentioned device parameters and the circuit functions properly for a device selected at random. Also the anode voltage time rate of change is determined by C_1, which is a microfarad larger than the capacitance of the center junction and hence the voltage across the center junction follows the increasing voltage of C_1. Consequently there is no surge of current causing premature firing.

Temperature stability was achieved by a form of self-bias. This involved a resistor, of the proper size, between the anode and gate.

In considering the design of the pulse generator the operating DC voltage will determine R_3 and R_4 while the prescribed pulse width will determine R_6. The R_1 and C_1 combination will be determined by the requirements that its time constant be small compared to the operating pulse repetition rate which in turn will determine R_2 and C_2.

In general the pulse generator might be employed in any system needing a narrow pulse with a rise time in the order of .2 micro-seconds. One of the possible applications of the circuit might be in an inverter circuit. It might also be used as an error detector in a closed loop servo system as its operation is a function of the DC voltage on the gate.

It is the belief of the author that the possibilities of this pulse generator are extensive and have scarcely been touched by this particular work. There is no reason to believe that the frequency of operation could not be extended far beyond the desired one megacycle achieved in this project. Also the number of useful wave shapes possible from this circuit is numerous. A suggestion for future work might be the variation of circuit parameters to obtain the optimum saw-tooth wave.

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