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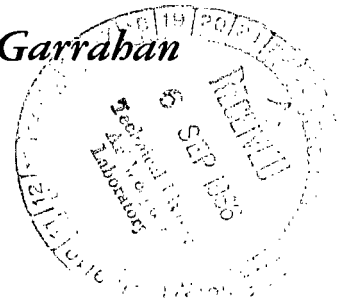
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LOW POWER NANOSECOND PULSE AND LOGIC CIRCUITS USING TUNNEL DIODES

by Stephen Paull, Ciro A. Cancro, and Norman M. Garraban
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ABSTRACT

A number of hybrid tunnel diode transistor pulse circuits have been developed to meet special requirements for cosmic ray experiments on scientific satellites and high-altitude balloons. These requirements include low power operation with pulse rise times as low as 50 nanoseconds, and compatibility for direct interconnection of threshold detectors, logic circuits, and commutating circuits without power-consuming amplifiers between stages. Circuit description and design information are given for five basic configurations, (1) threshold detector, (2) threshold detector and pulse stretcher, (3) single-pulse monostable, (4) double-pulse monostable, and (5) double-pulse monostable with feedback. Three functional combinations of these basic configurations are discussed to illustrate their compatibility for direct interconnection. They are (1) pulse width discriminator, (2) coincidence logic for particle detectors, and (3) pulse repetition frequency compressor. A basic microcircuit package is suggested, from which any one of the five basic configurations can be obtained by connecting appropriate shunts to external terminals.

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INTRODUCTION

A number of hybrid transistor-tunnel diode pulse circuits have been developed in connection with work on source-encoding electronics for cosmic ray experiments. This development was in response to several special requirements which had to be fulfilled in flight hardware for future scientific satellite payload and high-altitude balloon flight experiments. These requirements are the following:

- (1) Threshold detector, coincidence logic, and commutating circuits are required for random pulses with fast rise times (100 to 300 ns) and with relatively slow rise times (1 to 2 μ s).
- (2) A pulse repetition frequency compressor is required to prevent random pulses spaced closer than 2 μ s from feeding into the spacecraft telemetry.
- (3) A pulse rise time discriminator is required to distinguish between pulses with fast rise times and pulses with slow rise times.
- (4) Severe power limitation requires that all circuits use very low power.
- (5) Direct interconnection of circuits is required whenever possible to eliminate the need for amplifier and buffer stages.

These requirements have been met by a threshold triggering scheme using current switching between two transistors combined with tunnel diode circuits in five basic configurations. The design of each of these basic circuits is based on parameters defined in the tunnel diode static current-voltage characteristic (Figure 1). These characteristics are

I_p = peak-point forward current,

V_p = peak voltage, or voltage at which I_p occurs,

I_v = valley-point current,

I_f = absolute maximum forward current,

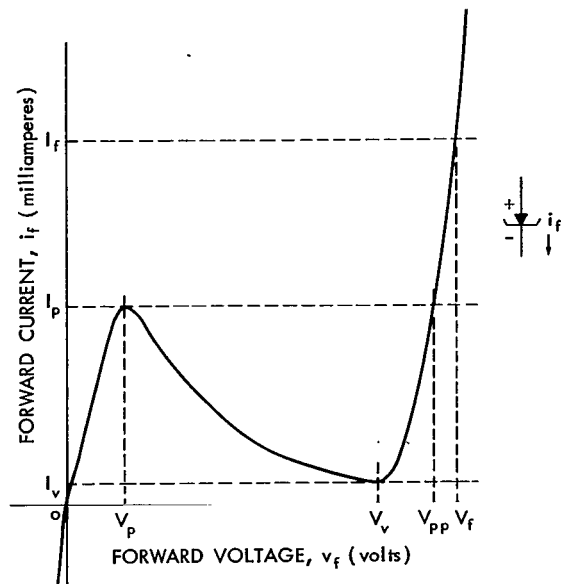


Figure 1—Tunnel diode static current-voltage characteristic.

V_v = valley voltage, or voltage at which I_v occurs,

V_f = any voltage in the forward region greater than V_v ,

V_{pp} = projected peak voltage, or the value of V_f at which the forward current equals I_p , and

R = tunnel diode junction dynamic resistance.

For design purposes, we assume the tunnel diode static characteristic to be approximately linear beyond the valley point, and we therefore may write

$$R = \frac{dV}{dI} \cong \frac{V_f - V_v}{I_f - I_v} \cong \frac{V_{pp} - V_v}{I_p - I_v} \quad (1)$$

Three design examples illustrate the manner in which the basic circuit configurations

may be interconnected to obtain threshold detection, pulse width discrimination, logic functions, and pulse repetition frequency compression.

BASIC CIRCUIT CONFIGURATIONS

Threshold Detector

A threshold detector with high resolution between closely spaced, random pulses must have a low threshold level in order to minimize the need for preamplification and consequent pulse broadening. Figure 2 shows a tunnel diode threshold detector with a transistor current-switching trigger circuit in which a threshold level as low as 200 millivolts can be obtained with high input impedance and high threshold stability. Total power drain is approximately 1.5 to 6 milliwatts, depending on the tunnel diode rated peak-point forward current, I_p . Threshold level is determined by the dc voltages set up in the resistive network comprised of R_1 , R_3 , R_4 , and R_5 . With transistor betas of 10 or more, threshold stability depends on the stability of I_p and not on transistor parameters. For high stability of I_p , the tunnel diode current should always be well below the absolute maximum rating.

In the quiescent condition, transistor Q_2 is cut off and transistor Q_1 conducts as an emitter follower. A negative input pulse applied to the base of Q_1 tends to make both emitters become more negative. Capacitor C holds V_{b_2} constant, however, and the current therefore switches from Q_1 to Q_2 . The Q_2 collector current now flowing through tunnel diode CR_1 rises to a value exceeding I_p , and CR_1 switches to its high voltage state. The resulting output voltage across CR_1 is $V_f > V_{pp}$.

(Figure 1). As shown on the waveform timing diagram in Figure 2, output pulse duration T_o is equal to the interval T_i during which the input pulse amplitude exceeds the threshold V_t . That is,

$$T_o = T_i = t_2 - t_1 .$$

Output pulse rise time is independent of input pulse rise time and depends only on the time constant $R_2 C_c$, where C_c is the sum of tunnel diode capacitance and stray capacitance between the Q_2 collector and output common.

Germanium tunnel diodes are particularly advantageous in this circuit for fast pulse operation, high threshold stability, and low power. For example, consider the type TD-9 with the following parameters.

Peak-point forward current	$I_p = 0.5 \text{ mA}$
Absolute maximum forward current	$I_f = 2.5 \text{ mA}$
Projected peak voltage	$V_{pp} = 0.5 \text{ V}$
Valley point capacitance	$= 2.5 \text{ pF}$

For high threshold stability, the circuit is designed so that tunnel diode current is

$$i = 2I_p = 1.0 \text{ mA} .$$

Let $E_{cc} = -3.0\text{V}$, and $R_2 = 1200 \text{ ohms}$. The voltage dividing network is designed so that the quiescent operating voltages shown in Figure 2 are

$$V_e = -1.8\text{V} ,$$

$$V_{b_1} = -1.3\text{V} ,$$

and

$$V_{b_2} = -1.5\text{V} .$$

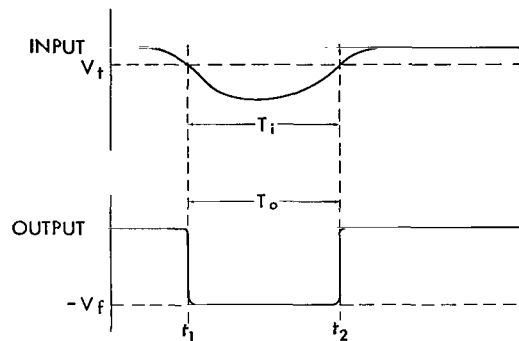
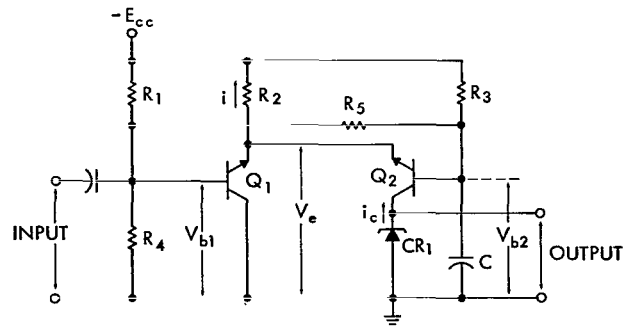


Figure 2—Threshold detector.

Assuming the output circuit stray capacitance to be approximately 2.5 pF, then $C_c \cong 5.0$ pf, and the output pulse rise time to 90 percent peak amplitude is

$$T_r \cong 3(1.2 \times 10^3)(5.0 \times 10^{-12})$$

$$\cong 18 \text{ nanoseconds.}$$

Output pulse peak amplitude for the TD-9 when $i = 2I_p$ is $V_f \cong 0.6V$. Power dissipation in the circuit is approximately 3 mW.

Threshold Detector and Pulse Stretcher

Figure 3 shows the current-switching transistor pair Q_1 and Q_2 arranged to trigger a tunnel diode monostable. This circuit operates as a threshold detector as previously described, and as a pulse stretcher for applications where output pulse duration T_o must exceed input pulse duration T_i . The waveform timing diagram (Figure 3) illustrates the pulse-stretching operation where $T_o = T_i + T$. The value of T is determined by the dc bias $-V_b$ and the values of L and R_6 .

Single-Pulse Monostable

In applications where output pulse duration T_o is required to be a fixed value less than input pulse duration T_i , the threshold detector of Figure 2 may be modified by connecting an inductance L across the tunnel diode as shown in Figure 4. For satisfactory operation, the input pulse must have a fast rise time, that is, its rise time must be small compared to the time constant L/R_2 .

The following design analysis shows how T_o may be adjusted to any value less than T_i by an appropriate selection of inductance L and tunnel diode parameters. During the interval T_i , we assume transistor Q_2 to be a constant current source, $i = I_f > I_p$ (Figure 1). We also assume that the tunnel diode and inductance may be represented by the equivalent circuit of Figure 5a, in which the resistive component of the inductive impedance is negligible compared with the tunnel diode dynamic resistance R (Equation 1). The tunnel diode is replaced by the voltage generator $v = V_v$ in series with R .

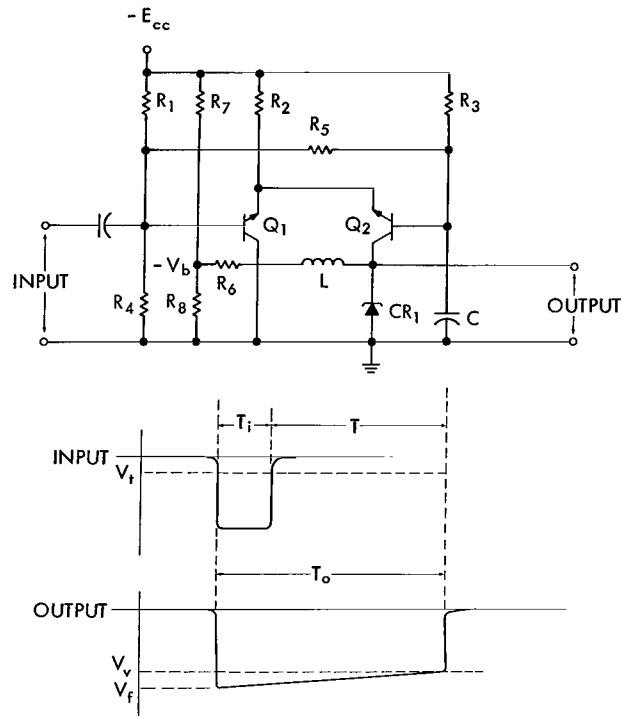


Figure 3—Threshold detector and pulse stretcher.

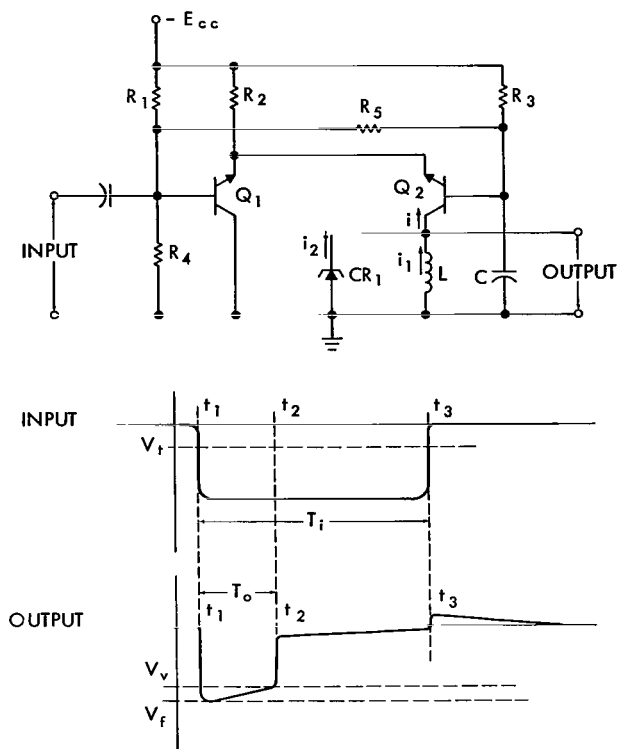


Figure 4—Single-pulse monostable.

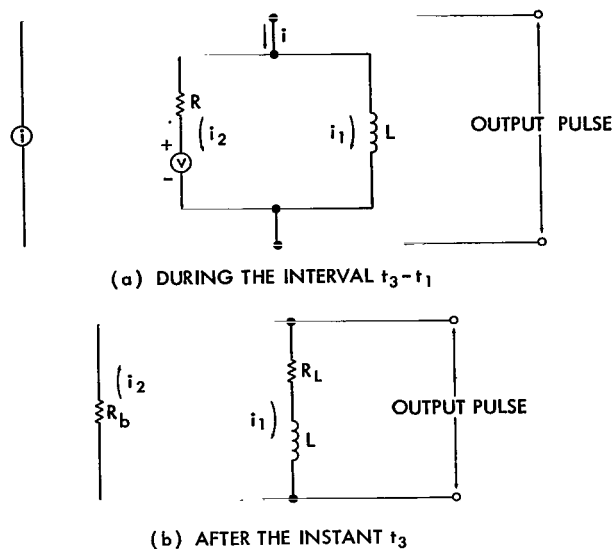


Figure 5—Equivalent circuits for design analysis: (a) During the interval $t_3 - t_1$; (b) After the instant t_3 .

The circuit equations for Figure 5a are

$$i = i_1 + i_2 \quad (2)$$

and

$$L \frac{di_1}{dt} - i_2 R = v \quad (3)$$

Solving for i_2 in Equation 2, and substituting the solution for i_2 in Equation 3 gives

$$L \frac{di_1}{dt} - (i - i_1) R = v \quad (4)$$

or

$$\frac{di_1}{dt} + \frac{R}{L} i_1 = \frac{v + iR}{L} \quad (5)$$

The solution to Equation 4 for i_1 is

$$i_1 = \frac{v + iR}{R} (1 - e^{-Rt/L}) \quad (5)$$

When $t = 0$ (at the instant t_1 in Figure 4), we get $i_1 = 0$ and $i_2 = i$, that is, all of the Q_2 collector current is flowing through CR_1 . According to Figure 1, the output pulse amplitude at the instant t_1 is therefore V_f . After t_1 , current begins to flow in L ; hence i_1 increases, and i_2 decreases. The instant t_2 in Figure 4 is the point at which i_2 reaches the value I_v and switches to its low voltage mode. This is indicated at t_2 in the pulse timing diagram where the output pulse amplitude falls to a negligible value. At the instant t_2 ,

$$t = T_o, \quad i_2 = I_v, \quad \text{and} \quad i_1 = i - I_v.$$

Substituting these values in Equation 5,

$$e^{-RT_o/L} = \frac{V_v + I_v R}{V_v + iR}. \quad (6)$$

From this, the following is obtained:

$$T_o = \frac{L}{R} \ln \frac{V_v + iR}{V_v + I_v R}.$$

If the circuit is designed so that $i \gg I_v$, then T_o may be approximated by the expression

$$T_o \approx \frac{L}{R} \ln \frac{V_v + iR}{V_v}. \quad (7)$$

Transistor Q_2 (Figure 4) remains conducting and transistor Q_1 remains off for the duration of the input pulse T_i . The circuit returns to the quiescent condition with Q_1 on and Q_2 off at the instant t_3 in Figure 4. From this it can be seen that the output pulse duration T_o can never exceed T_i . After t_3 the equivalent circuit in Figure 5b applies, where R_L is the resistive component of the inductive impedance Z_L , and R_b is the very small resistance of the tunnel diode with i_2 flowing in the back direction. If $R_b \gg R_L$, the following circuit equations may be written:

$$i_1 = i_2 \quad (8)$$

and

$$\frac{di_1}{dt} + \frac{R_L}{L} i_1 = 0. \quad (9)$$

The solution to Equation 9 is

$$i_1 = I_f e^{-R_L t/L} = i_2. \quad (10)$$

When $t = 0$ (at the instant t_3),

$$i_1 = i_2 = I_f .$$

A small positive pulse $i_2 R_b$ beginning at t_3 appears across the output terminals as shown in the waveform timing diagram in Figure 4. For normal circuit response to a subsequent input trigger, the current i_2 must be essentially zero. The time required for i_2 to decrease to a negligible value depends on the time constant L/R_L in Equation 10. This determines the effective dead time of the monostable and the minimum possible repetition period.

Double-Pulse Monostable

In Figure 6 the monostable circuit has a pair of tunnel diodes in a back-to-back configuration. The effect of adding the second tunnel diode is twofold:

- (1) The dead time is reduced to a fixed value $T_i + T_o$ independent of R_L .
- (2) A bipolar pulse pair is obtained on output 1 and a single negative pulse is obtained on output 2.

Circuit operation is described with reference to the waveform timing diagrams in Figure 6 as follows. During the time interval $t_3 - t_1$, current flows in the forward direction through tunnel diode CR_1 and in the backward direction through tunnel diode CR_2 ; the small backward resistance of CR_2 is negligible. Operation during this time is essentially the same as for the single-pulse monostable, and a negative pulse of duration T_o beginning at t_1 appears on both outputs 1 and 2. The value of T_o is determined by the tunnel diode parameters and the value of L as described for the single-pulse monostable. After the instant t_3 , the transistors return to the quiescent condition, with Q_1 conducting and Q_2 off. Current in the tunnel diodes is now reversed, and the CR_1 backward resistance is negligible. Tunnel diode CR_2 switches to its high voltage stage, and a positive pulse of width T_o appears on output 1 only.

It should be noted that if the polarity of both tunnel diodes is reversed, the bipolar

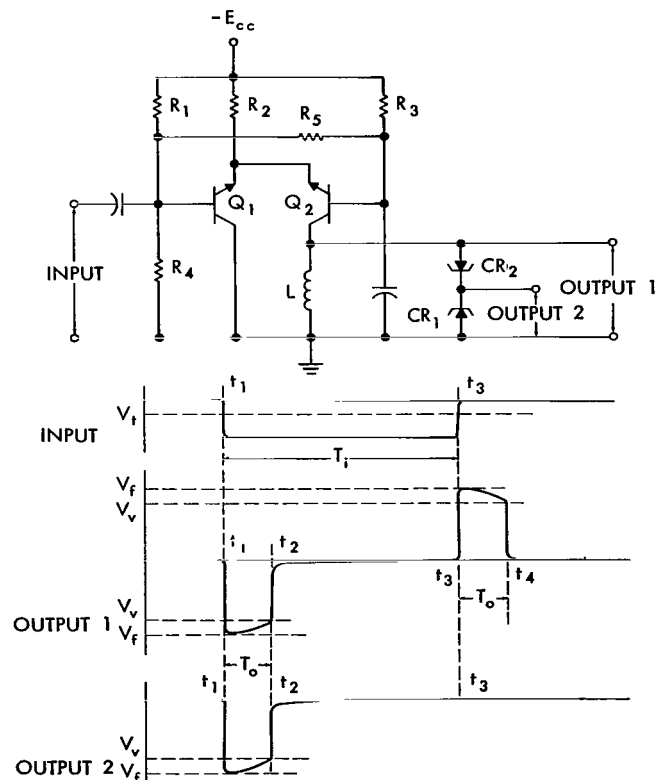


Figure 6—Double-pulse monostable.

pulse pair on output 1 remains unchanged; output 2 has no pulse at t_1 but has a single positive pulse of duration T_o beginning at t_3 . Furthermore, a slightly different configuration of the double-pulse monostable shown in Figure 7 may be used to produce the same outputs in response to a positive input signal instead of a negative input signal.

Double-Pulse Monostable with Feedback

In applications where output pulse duration of a fixed value, independent of input pulse

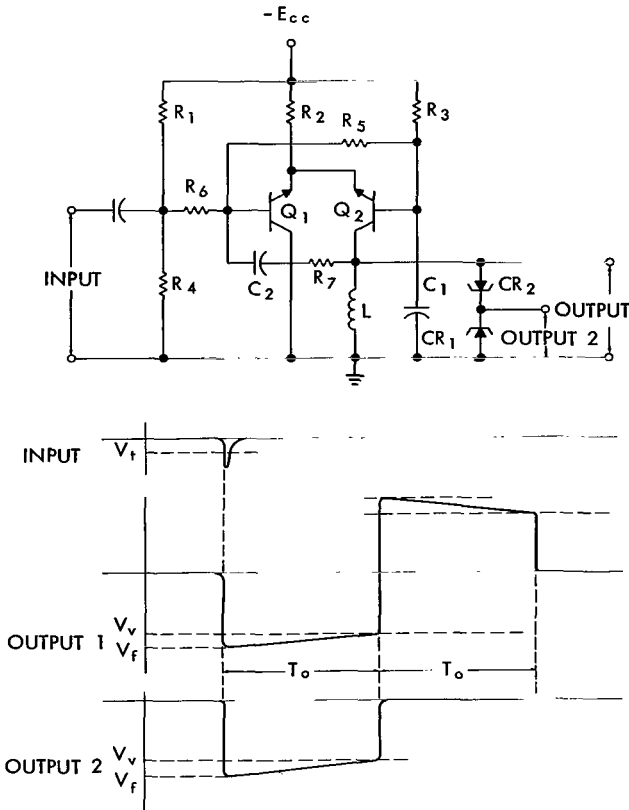


Figure 8—Double-pulse monostable with feedback.

of R_2 . The monostable may thus be designed to use germanium tunnel diodes for very fast pulse rise times with the two-pair series arrangement to obtain sufficient amplitude to drive an unbiased silicon transistor amplifier stage. The unbiased amplifier requires no standby power; therefore

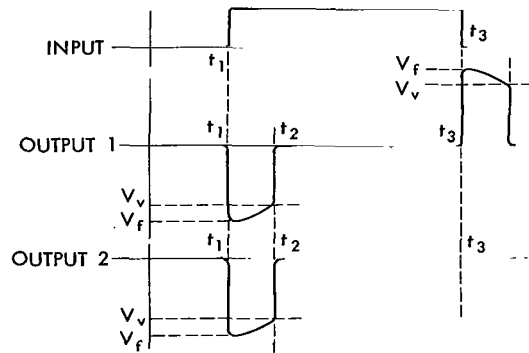
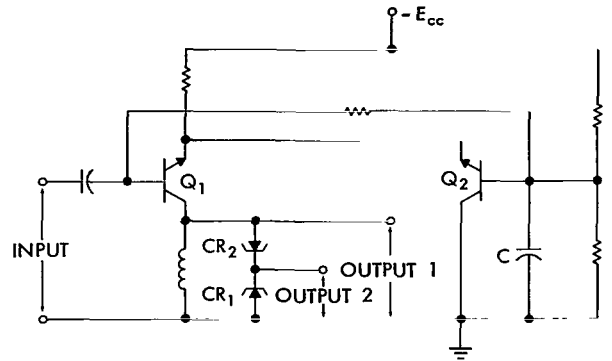


Figure 7—Double-pulse monostable for positive input.

duration, is required, the double-pulse monostable is modified by adding feedback path R_7, C_2 as shown in Figure 8. When input pulse duration T_i exceeds the required value for T_o , circuit operation is the same as shown by the waveforms in Figure 6. The feedback path is useful only when $T_i < T_o$ as shown by the waveforms in Figure 8. The feedback signal holds Q_1 off for the duration of the negative output pulse, and the resulting bipolar pulse pair on output 1 is a single square-wave cycle with period $2T_o$.

Output pulse amplitudes may be doubled with only small increase in power consumption by using four tunnel diodes in series, two facing in each direction, and by readjusting the value

at low duty cycle operation, the monostable and amplifier consume only slightly more power than that required by the monostable alone.

FUNCTIONAL COMBINATIONS OF BASIC CIRCUITS

Pulse Width Discriminator

Figure 9 shows three of the basic circuit configurations interconnected to operate as a pulse width discriminator. An output pulse of fixed width T_o is produced for every input pulse whose width T_i is less than a given value T , and no output occurs if $T_i > T$. Germanium tunnel diodes type TD-9 are used throughout, and the circuits are designed so that the transistor switching current in each stage is $i = 2I_p = 1$ mA. Total power consumption for the three stages is, including bias bleeders, slightly more than 12 mW.

As shown in the waveform timing diagrams in Figure 9, the first stage threshold detector produces a negative pulse A whose duration is equal to T_i , with fast rise and fall times independent

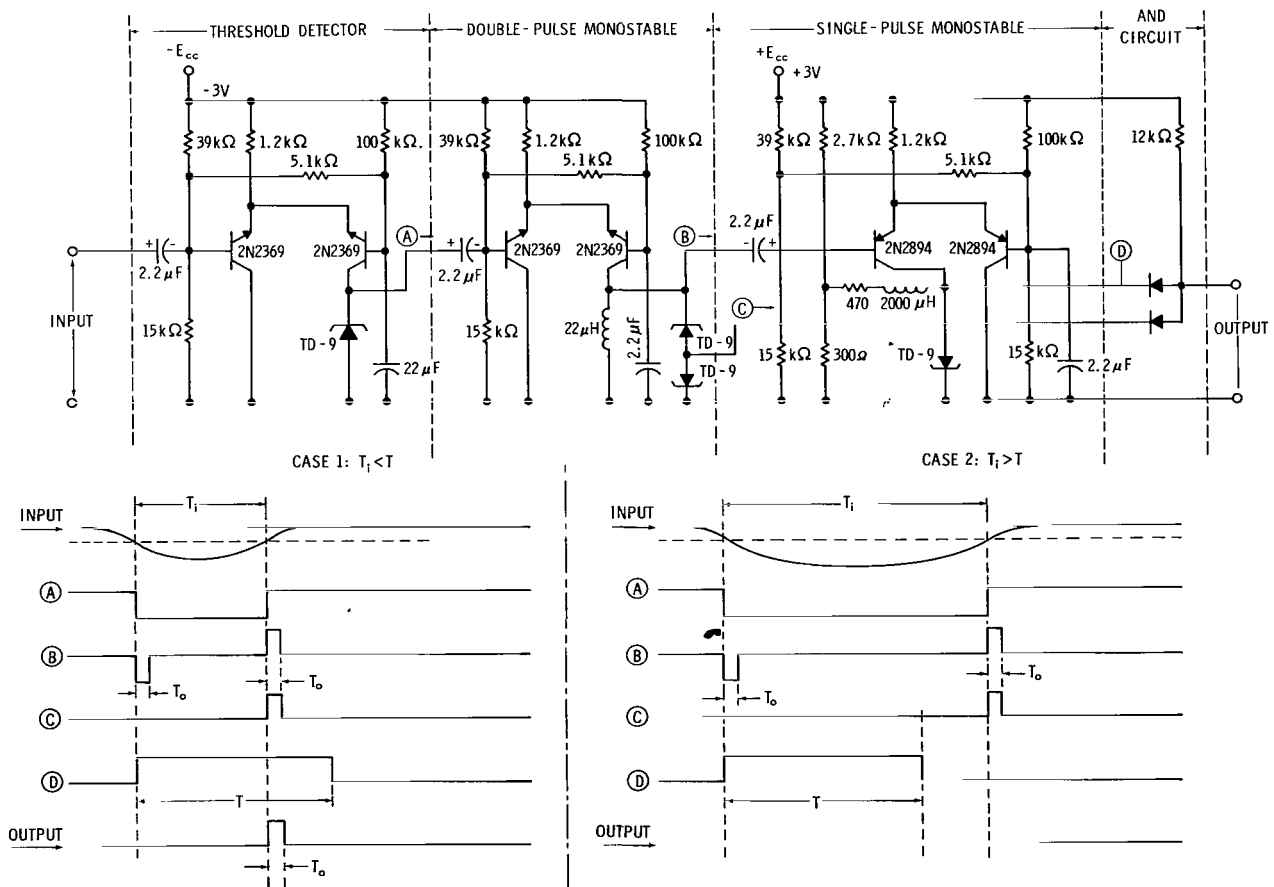


Figure 9—Pulse width discriminator.

of input pulse shape. This pulse drives the second stage double-pulse monostable to produce two outputs, B and C. Output B is a bipolar pulse pair. The first pulse is negative, with its leading edge at the leading edge of A. The second pulse is positive with its leading edge at the trailing edge of A. Output C is a single positive pulse with its leading edge at the trailing edge of A.

All B and C pulses have a fixed duration T_o determined by the circuit parameters and tunnel diode characteristics. For the TD-9, the manufacturer's published characteristics give

$$V_{pp} = 500 \text{ mV} ,$$

$$V_v = 350 \text{ mV} ,$$

$$I_p = 500 \text{ } \mu\text{A} , \text{ and}$$

$$I_v = 350 \text{ } \mu\text{A} .$$

Substituting these values in Equation 1 to compute the approximate tunnel diode dynamic resistance R gives

$$R \cong \frac{(500 - 350)(10^{-3})}{(500 - 60)(10^{-6})} = 341 \text{ ohms} .$$

In the circuit of Figure 9, $L = 22 \text{ } \mu\text{H}$, and $i = 1 \text{ mA}$. When these values for R, L, i, and V_v are substituted in Equation 7, the output pulse duration is

$$\begin{aligned} T_o &\cong \frac{22(10^{-6})}{341} \ln \frac{[350 + 341(1)] [10^{-3}]}{350(10^{-3})} \\ &= (0.645 \times 10^{-7})(\ln 1.97) = 43.8 \times 10^{-9} \text{ seconds} \\ &\cong 44 \text{ nanoseconds} . \end{aligned}$$

The negative pulse B triggers the third stage single-pulse monostable which produces a positive pulse D whose duration is T. This output is combined with output C in the diode AND circuit. In case 1, where $T_i < T$, the waveform timing diagram shows that output C occurs during the time D occurs, and this coincidence produces a pulse at the output. In case 2, where $T_i > T$, output C occurs after D has ended, and no pulse appears at the output.

Coincidence Logic for Particle Detectors

Figure 10 shows a three-parameter threshold detector and coincidence-anticoincidence logic system for low power operation in a cosmic ray experiment on a small scientific satellite. The

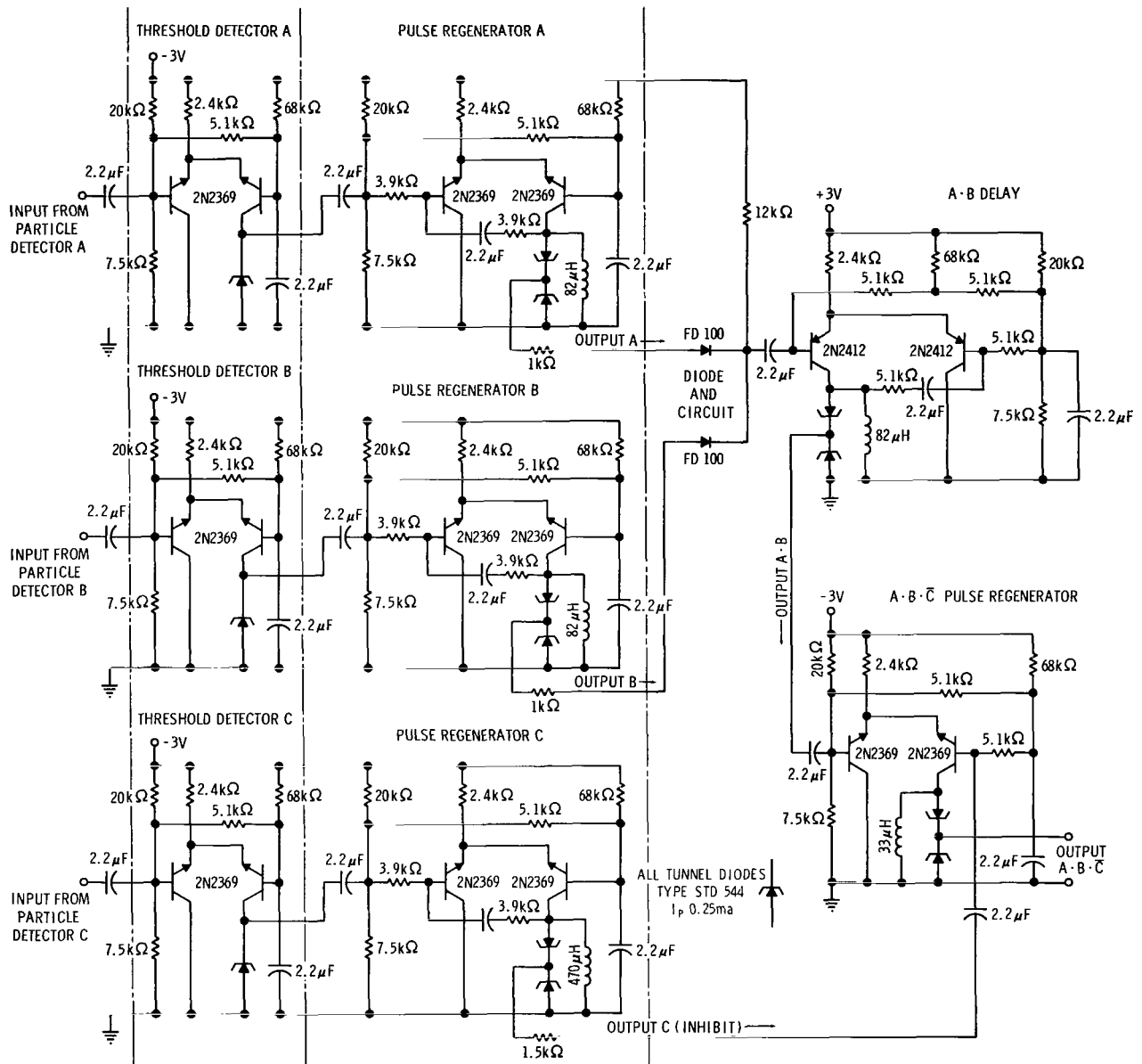


Figure 10—Low power, three-parameter, threshold detector and logic system.

logic circuits produce a single output gate pulse when, and only when, simultaneous input pulses are generated by two particle detectors A and B, and no pulse is generated by a third particle detector C. Pulses from all three detectors occur at random with peak amplitudes distributed over a very wide range, and rise times such that they all reach a threshold of approximately 200 millivolts in 100 nanoseconds or less. The complete system consists of eight hybrid transistor-tunnel diode stages using type STD 544 tunnel diodes with a rated peak-point current $I_p = 250$ microamperes. Each stage is powered from either +3 volts or -3 volts dc, and is designed so that tunnel-diode current is $2I_p = 500$ microamperes. Power dissipation, including that in the bias

voltage resistive network, is, therefore, slightly greater than 1.5 milliwatts per stage; the total power dissipation is slightly more than 12 milliwatts.

Negative input pulses from each particle detector are fed to a threshold detector similar to the basic circuit configuration shown in Figure 2. A threshold detector output pulse with fast rise time independent of input pulse rise time is generated whenever input pulse amplitude exceeds a threshold of 200 millivolts. This output pulse triggers a pulse regenerator stage consisting of the double-pulse monostable configuration with feedback shown in Figure 8. The pulse regenerator stage furnishes an output pulse of fixed duration T_D independent of trigger pulse duration. The back-to-back tunnel diode pair enables this stage to operate at high repetition rates with trigger pulses spaced as close as $2T_D$.

The output from the channel A pulse regenerator is combined with the output from the channel B pulse regenerator in a diode AND circuit. Whenever A and B pulses coincide, they trigger the $A \cdot B$ delay stage which operates in a manner similar to the double-pulse monostable configuration shown in Figure 7. The delay is included to allow time for the channel C threshold detector and pulse regenerator stages to respond in case a channel C input pulse occurs and requires 100 nanoseconds to reach threshold. The delayed $A \cdot B$ output triggers the $A \cdot B \cdot \bar{C}$ pulse regenerator if there is no inhibit pulse from the channel C pulse regenerator. The $A \cdot B \cdot \bar{C}$ pulse regenerator operates in a manner similar to the double-pulse monostable configuration shown in Figure 6.

The values of pulse duration and delay time in the logic circuits are chosen for maximum coincidence resolution and minimum rejection of truly coincident pulses reaching threshold at different instants. The channel A and channel B pulse regenerator stages have therefore been designed with $L = 82$ microhenries, so that pulse duration T_D is somewhere near 100 nanoseconds. The same value of L is used in the $A \cdot B$ delay stage in order to delay the $A \cdot B$ output by the same amount. The approximate value of T_D for these stages can be computed by substituting the appropriate parameters for the STD 544 in Equations 1 and 7. Rated parameters for the STD 544 are

$$V_v = 350 \text{ millivolts ,}$$

$$V_{pp} = 500 \text{ millivolts ,}$$

$$I_p = 250 \text{ microamperes ,}$$

and

$$I_v = 60 \text{ microamperes .}$$

The circuits are designed so that tunnel diode current is $i = 2I_p = 500$ microamperes. Substituting these parameters in Equation 1 gives an approximate value for the tunnel diode dynamic resistance R of

$$R \cong \frac{(500 - 350)(10^{-3})}{(250 - 60)(10^{-6})} = 790 \text{ ohms .}$$

From Equation 7,

$$T_D \cong \frac{82 (10^{-6})}{790} \left\{ \ln \left[\frac{350 + 0.5 (790)}{350 (10^{-3})} \right] \right\}$$

$$= (1.04 \times 10^{-7}) \ln 2.12$$

$$= 1.04 (0.75) (10^{-7}) = 78 \text{ nanoseconds .}$$

In the channel C pulse regenerator, $L = 470$ microhenries. From Equation 7, the computed duration for the channel C inhibit pulse is

$$T_c \cong \frac{470 \times 10^{-6}}{790} \ln 2.12 = 445 \text{ nanoseconds .}$$

In the $A \cdot B \cdot \bar{C}$ pulse regenerator, $L = 33$ microhenries. From Equation 7, the computed duration for the $A \cdot B \cdot \bar{C}$ output pulse is

$$T_o \cong \frac{33 \times 10^{-6}}{790} \ln 2.12 = 42 \text{ nanoseconds .}$$

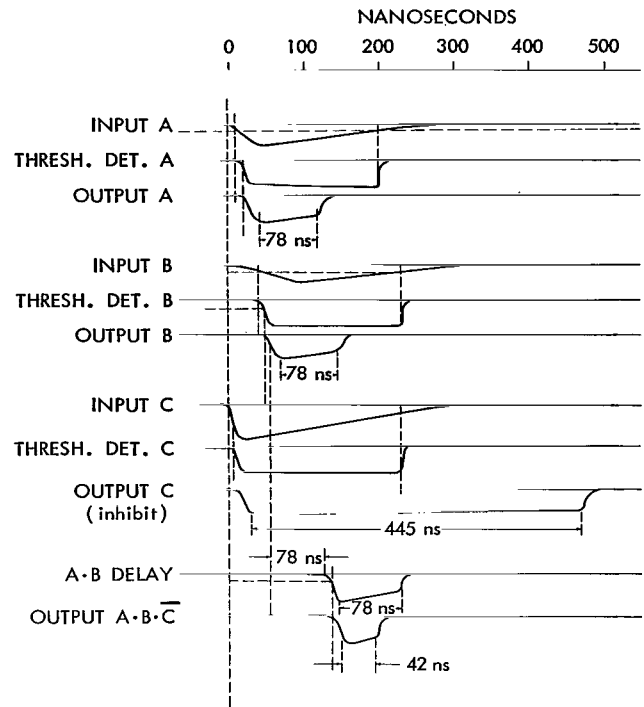


Figure 11—Pulse timing diagram for threshold detector and logic system.

The pulse timing diagram in Figure 11 illustrates the operation of the circuits in Figure 10 with the computed values $T_D \cong 78$ ns, $T_c \cong 445$ ns, and $T_o \cong 42$ ns. In this example, coincident input pulses A, B, and C reach threshold after different time intervals; 10 ns for A, 40 ns for B; and less than 10 ns for C. The output $A \cdot B \cdot \bar{C}$ is produced only when the input pulse on channel C is absent.

Pulse Repetition-Frequency Compressor

A portion of a scientific satellite experiment measures average repetition frequency of pulses generated by random particle impacts. This information is obtained by feeding the random pulses into registers in the telemetry encoder during fixed sampling periods. The registers are designed for average pulse rates up to 500 kcs, but do not operate at all at higher rates. In order to obtain pulse rate information above 500 kcs, a pulse repetition frequency compressor is required between the random pulse output and the register input. This compressor must reject a sufficient number of closely spaced pulses so that spacing of the remaining pulses is never less than $2 \mu s$.

The double-pulse monostable with feedback (Figure 8) behaves in this manner when input trigger amplitudes are sufficiently limited so that the feedback signal to Q_1 base holds Q_1 on for the entire positive portion of an output oscillation. This prevents further triggering of the monostable for the complete period $2T_o$ and results in the input-output characteristic shown in Figure 12. The maximum possible output frequency is $1/T_m$ where $T_m = 2T_o$.

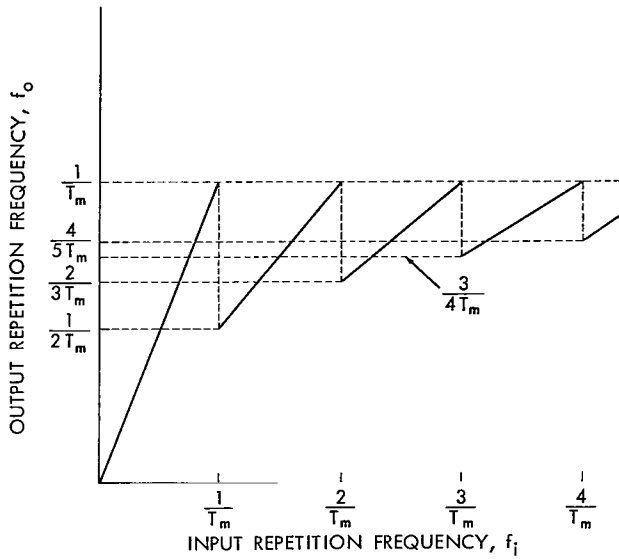


Figure 12—Repetition frequency compressor characteristic.

Figure 13 shows the double-pulse mono-stable with feedback preceded by an input-shaping circuit and followed by an unbiased amplifier. The input-shaping circuit consists of a back-to-back tunnel diode and shunt inductance configuration to limit input trigger amplitude and to produce a bipolar pulse pair for each input trigger pulse. The limiting action is sufficient to permit the monostable to operate as a repetition-frequency compressor, while the bipolar pulses prevent a charge from building up on the input coupling capacitor C_c . This in turn prevents a shift in triggering threshold which would otherwise occur when input signal duty cycle is high.

The germanium-type TD-9 tunnel diodes in the input-shaping circuit limit trigger peak

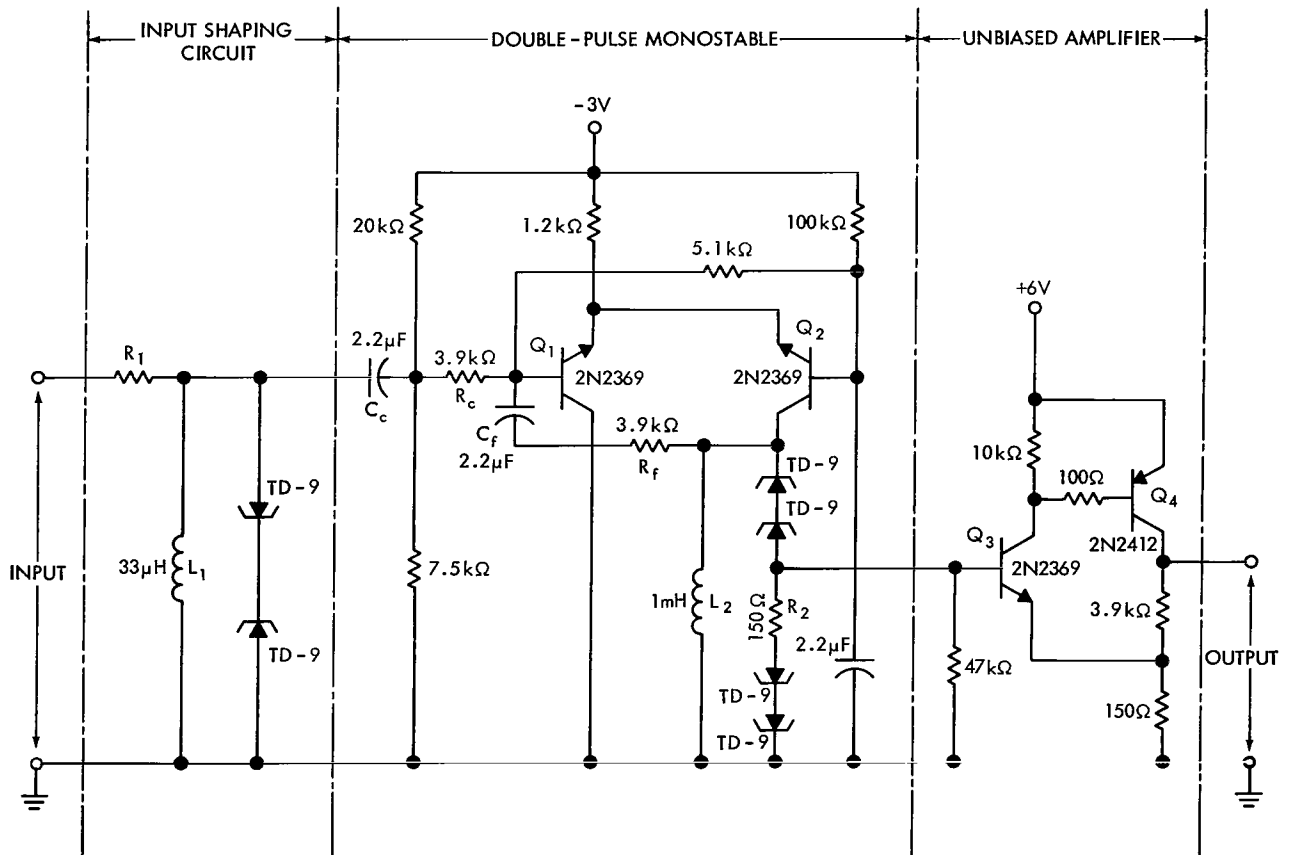


Figure 13—Repetition frequency compressor and unbiased amplifier.

amplitudes to a maximum value between 450 and 550 mV. The monostable itself has four type TD-9 tunnel diodes in series, two facing in each direction. With this arrangement, the output pulses always exceed $2V_v$ (700 mV plus the drop across R_2), and the feedback voltage to the base of Q_1 always exceeds the trigger peak amplitude. Furthermore, the output pulse amplitude is sufficient to drive the unbiased amplifier and thus obtain a voltage gain of approximately six with zero standby power drain.

For the type TD-9, $R \cong 341$ ohms, and $V_v = 350$ mV. The input resistance R_1 is adjusted for a given input signal amplitude so that tunnel diode current is $i = 2I_p = 1$ mA. Inductance $L_1 = 33 \mu\text{H}$ determines the width T_i of the input bi-polar trigger pulses fed to the base of Q_1 . According to Equation 7,

$$\begin{aligned} T_i &\cong \frac{33 \times 10^{-6}}{341} \ln \frac{(350 + 341)(10^{-3})}{350(10^{-3})} \\ &= 96.8 (\ln 1.97) (10^{-9}) \\ &= 96.8 (0.678) (10^{-9}) = 65.5 \text{ nanoseconds} . \end{aligned}$$

The monostable output pulse going to the base of transistor Q_3 in the amplifier has a duration T_o determined by $L_2 = 1$ mH and the dynamic resistance of two type TD-9 tunnel diodes in series, $R \cong 682$ ohms. From Equation 7,

$$\begin{aligned} T_o &\cong \frac{10^{-3}}{682} \ln \frac{2(350 + 341)(10^{-3})}{2(350)} \\ &= 1.47 (0.678) (10^{-6}) = 0.995 \times 10^{-6} \text{ sec} . \end{aligned}$$

Maximum repetition frequency is

$$f_m = \frac{1}{2T_o} \cong \frac{10^6}{1.990} \text{ c/s} , = 502 \text{ kc/s} .$$

BASIC MICROCIRCUIT PACKAGE

A basic circuit for packaging in a standard TO-5 transistor can or in flatpack form with nine external leads is shown in Figure 14. Any one of the five basic configurations can be formed from this unit by connecting external components and/or jumper wires to the appropriate external leads. Threshold level adjustment may also be made by connecting appropriate shunt resistors to terminals 1, 4, 6, and 7.

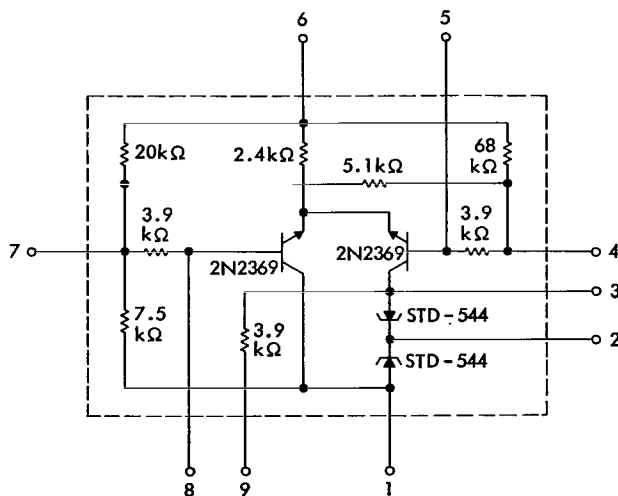


Figure 14—Basic microcircuit package.

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