During "read," Q3 is cut off and Q4 is saturated. Q4 takes the full write current through diodes D5 and D6, back biasing diodes D1 through D4. With Q4 saturated diodes D7 and D8 are back biased, allowing Q5 and Q6 to be saturated, thus connecting the two series-connected tracks to the read amplifier at a d-c level of approximately zero.

**Read Amplifier**

The read amplifier, Fig. 9, has two difference-amplifier stages and one output stage with more than enough gain to give a saturation output signal at a tape speed of 20 ips. In the first two stages, the common mode gain per stage is less than unity while the difference signal gain is approximately beta. The low common mode gain insures that power supply noise will not be amplified. Each transistor (Q1-Q4) is biased to a constant d-c operating point of approximately 3.8-smitter-collector volts and 1.9-ma collector current. The capacitors shown must only be large enough to have negligible signal attenuation at 20 ips, the lowest tape speed of interest. The lowest frequency signal will be at 20 ips and 100 cycles per inch (all ones or all zeros) for a frequency of 184 kcs. The micro-alloy 2N393 transistors have more than enough bandwidth for this application. The signal amplitude at the input to Q5 and Q6 is large enough so that, most of the time, one of these transistors is saturated. The output signals are of a computer-type amplitude (0 or -3) and are sampled by TP1 using conventional TX-2 logical circuits.

**References**


**The Dynamics of Toggle Action**

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It is common in the design of regenerative circuits to perfect the d-c (static) design on paper and the a-c (dynamic) design in the laboratory. In the process of working out the static design, a certain amount of engineering judgment can be used so that the resulting circuit will give approximately the required dynamic performance, but rarely will the d-c-designed circuit pass all the performance specifications without some changes dictated by laboratory tests.

Although it is awkward to calculate directly the effects of loading and component variation on stability, switching time, and triggering characteristics, it is even more awkward to measure these effects experimentally because interaction of all the components makes it difficult to find the worst combination.

This paper presents a design method using an intermediate step, i.e., the negative resistance curve, between the static and dynamic design. The effects of loading and component variation show up clearly as changes in the negative resistance curves and, in turn, the altered dynamic performance can be calculated easily from these curves.

The method is here applied only to toggle (flip-flop) design, but the extension to monostable circuits in simple. In principle, the method is applicable to other regenerative circuits such as blocking oscillators.

The paper includes the derivation of negative resistance curves from the circuit parameters, a method of evaluating toggle performance from the curves, and an example in the form of a transistor toggle.

**Description of the Curve**

Fig. 1(A) shows a typical toggle using n-p-n transistors. Suppose the resistor sizes have been chosen to produce some standard swing (V1 to V2) with some standard power supply voltages (E1 and E2). The object is to evaluate this proposed design.

A variable voltage source is shown connected to one collector, with suitable meters for current and voltage measurements. If the source voltage is either V0 or V1, no current will flow since V0 and V1 are the stable, open-circuit output voltages. Some current will flow at other voltages.

When T1 is conducting, and saturated, the current will be zero a voltage very close to zero. If T1 is forced more negative, current will rise rapidly along the saturated collector characteristic (Rc). This is shown as A, B in Fig. 1(B). Since collector current cannot exceed \( \beta_{ib} \), where \( \beta \) is the base current, T1 will pull out of saturation when the collector current reaches \( \beta_{ib} \).

The measured current will continue to rise, but more slowly, (C, D, E) and the slope is now that of \( R_{bi} \) and \( R_{c} \) and the grounded-emitter collector resistance of \( T_{1} \) all in parallel.

Up to this point, T1 has been cut off, but as the collector voltage of T1 rises, so does the base voltage of T2. When T2 begins to conduct, the circuit behavior alters radically. The current through T1 is reduced as the collector voltage of T2 drops, so less current, rather than more, is required from the external source. In fact, as the voltage is increased further, the current required drops to zero and reverses. (C, D, E on Fig. 1(B)). When T1 is completely cut off, the current again begins to rise, and the slope is now \( R_{bi} \) and a very large collector resistance all in parallel. At \( E_{R} = V_{i} \) the current crosses the axis and will continue to rise linearly until breakdown.

This negative resistance (NR) curve describes completely the output characteristics of the circuit, just as the plate curves describe a tube. Load lines can be drawn that will indicate loading capability, as will be shown later. In addition it will be shown that the NR curve is an aid in deducing triggering thresholds (duration and amplitude) and switching time.

**Derivation of the NR Curve from the Circuit Parameters**

It is not necessary to make the previous measurement to arrive at the NR curves for a circuit. Consider Fig. 2(A). Assuming the base swing is small compared to \( E_{R} \), the current through \( R_{bi} \) is nearly constant at \( E_{bi} / R_{bi} \). When \( T_{1} \) is saturated, its collector current \((\dot{I}_{C})\) is \( (E_{bi} / R_{bi}) - I_{b} \). The drop across the right
hand $R_1(E_{ao})$ is then $i_R R_2$ and that is the amount by which $T_2$ is cut off. The collector voltage of $T_2$ is $R_2E/E_1 + R_1$, neglecting the base-emitter drop of $T_1$. The base current ($i_b$) of $T_1$ is $(E_1/R_1 + R_2) - i_b$. The collector current of $T_1$ must be less than $i_R$ in order for $T_1$ to be saturated.

The foregoing paragraph must be obvious to anyone who has ever gone through the d-c design of a toggle. It serves only to define a few terms.

Again imagine the collector voltage of $T_1$ to be forced negative by some external device. The external current will rise steeply along $T_1$'s saturation curve until the collector current is $i_3$ minus the amount inserted through $R_3$, the current that the external device is required to insert. Since here $T_1$ is cut off, $r_s$ is large and the slope through $E_F$ is simply $R_1$, in parallel with $r_s$. $E_F$ is just $R_3E/R_1 + R_3$.

Starting at $E_F$ and reducing the voltage of the external source, the external current will be determined by the resistance $R_2R_3/R_1 + R_2R_3$. As the voltage is reduced, the base current of $T_2$ will be reduced, and $T_2$ will come out of saturation. When the collector voltage of $T_2$ has risen to $E_{ao}$, $T_1$ will again begin to conduct because its base voltage will have risen to ground. The current in $T_1$ when $T_2$ is on the verge of conduction is then $i_3 = (E_1 - E_{ao}/R_1) - i_3$ and this quantity divided by $\beta_2$ is the base current required in $T_2$. The current through the right hand $R_2$ will be $(i_3/\beta_2) + i_3$ and the required voltage on $T_1$'s collector is:

$$I_3 = \frac{E_F - E_{ao}}{R_2}$$

For very large $\beta_2$, this voltage approaches $i_R$ which is also $E_F$.

Notice that, beginning at zero volts, the point reached, $C$ or $C'$, depends only on $\beta_2$ and, beginning at $E_F$, the point reached, $E$ or $E'$, depends only on the term $\beta_2$.

The range of voltage $C - E$ is the active region, and, if the $\beta$'s are constant here (a good assumption), and $T_1$'s base-emitter drop is constant (a poor assumption), the lines $C - E$ will be straight. It is easy to take into account the effect of $T_2$'s base-emitter voltage, as will be seen later.

An additional useful bit of information is $I_3$, the maximum current the external source can remove from the circuit without preventing bistability.

**Evaluation of Circuit Using NR Curves**

**Drifting Capability**

The circuit margin with respect to $\beta$ changes can be seen on the NR curves since the points of circuit stability are the points where the NR curves cross the load line with positive slope. The unloaded case is a special one in which the load line is a zero-current line and therefore coincides with the horizontal axis.

Fig. 3 shows two load lines on a typical pair of NR curves. $R_3$ is returned to some supply voltage $E_X$, $R_2$ is returned to ground. These would be typical of "and" and "or" gates in computer circuitry. For the low $\beta_2$ curve, the toggle is not bistable with load $R_3$, since there is only one point of intersection of the load line and the NR curve. With load $R_3$, the toggle is bistable for both low $\beta_2$ and high $\beta_2$. Marginal load lines would pass through $C$ or $C'$ and $E$ or $E'$.

**Fig. 3. NR curves with typical load lines**

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Usually, other considerations come into play before lack of bistability becomes a factor. A certain tolerance on output voltage must be maintained. When this tolerance is specified, a load line can be constructed to produce limit voltages and the minimum load resistance thus determined.

Clamps are a special case of resistive loading. Their effects can be shown as in Fig. 4, where $D_H$ and $D_L$ standardize the output voltage at $E_H$ and $E_L$. $D_L$ may not be necessary for standardizing output swing, but is often used to prevent saturation in the transistors. Since $T_0$'s collector is not allowed to go to $E_p$, $T_0$'s base current is reduced. $I_e$ is then less than in the unclamped case. An alternative display of the effect of clamps is Fig. 4(C), where the diodes have been considered as part of the toggle, rather than as part of the load. The NR curve has been reshaped by adding together the diode current and the NR current at each output voltage. Similarly, built-in load resistors can be included in the toggle's NR curve by adding the resistor current to the NR current at each voltage. (When adding, it must be remembered that the load resistances were plotted as negative when they were loads, and must be reversed in sign when they are to be considered as part of the toggle.)

**D-C Trigger Threshold**

In many cases it is advisable to trigger a toggle by turning "off" an "on" transistor rather than vice versa because the trigger pulse is applied to an active (conducting) element. That type of triggering will be covered in this paper. Triggering "on" is a straightforward application of the same techniques.

Refer to Fig. 5. Suppose the trigger is to be applied to the base of $T_1$ (which is conducting). Current in $T_1$ must be reduced till its collector rises to $E_{co}$, the amount by which $T_2$ was cut off. At this point $T_2$ comes into conduction and helps cut $T_1$ off. The current in $T_1$ at the verge of regeneration is $i_2 = (E_1 - E_{co}/R_1) - i_3$. The base current $i_b = (E_1/R_1 + R_2) - i_3 - i_4$. Solving for $i_2$ (minimum), $i_2 = (E_1/R_1 + R_2) - i_4 + (i_3/\beta_1) - (E_2 - E_{co}/\beta_1 R_1)$ which is seen to be $I_c/\beta_1$.

Therefore, the minimum trigger current is just $1/\beta_1$ times the height of the NR curve above the axis. If $\beta_1 I_t > I_c$ there will be only one point of intersection (point $F$) and the toggle will flip to that state and stay there when the trigger is removed.

The effect of triggering can be shown by reshaping the NR curve to take into account the reduced base current. Fig. 5(B) shows an NR curve with several different triggering levels.

**Pulse Trigger Threshold**

Because of capacitive loading and finite current rise times in transistors, one can expect a higher trigger threshold when triggering a toggle with short pulses. This threshold is difficult to specify in general, but for any particular case, graphical methods will produce a good answer.

The same statements apply to switching time and waveform for different triggering levels. The numerical example to follow will illustrate all the aforementioned techniques and also demonstrate the method for solution of the waveform and threshold problems.

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*Fig. 4. Two representations of the effect of clamps*

*Fig. 5. NR curves showing the effect of trigger current*

*Fig. 6. Calculated and measured NR curves for a typical circuit*

*Fig. 7. Threshold trigger current versus trigger time, with and without capacitive load*

*From the collection of the Computer History Museum (www.computerhistory.org)*
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On the NR curves, Fig. 1(B), D is a point of unstable equilibrium, as the load line (in this case the horizontal axis) crosses the NR curve in a region of negative slope. If the circuit is brought near point D by an external force and then released, it will regenerate to point A or point F, depending on which side of D it was on when released. To inquire as to what will happen if the starting point is exactly D, one must consider ever-present noise, and similar practical difficulties. Toggles will not stay at point D for the same reason that pencils will not balance on their points. The important point is that for successful triggering, the circuit operating point must pass $E_B$ by the time the trigger has gone away. Otherwise, the circuit will return to its starting point.

**Example**

Fig. 6 shows a typical toggle circuit and its measured and calculated NR curves. The disagreement at low voltage, point $B_1$, is due to the erroneous assumption of constant $\beta$ at low voltage and the neglect of $r_e$. The offset in the negative resistance region is due to the fact that the base-emitter drop of $T_2$ must be added to the calculated straight line $CDE$. This results in excellent agreement with the measured curve. The calculated threshold trigger current is:

$$I_C/\beta_1 = 12.3/57 \approx 215 \text{ microamperes}$$

The measured threshold trigger current is 220 microamperes. Note that the "complete triggering" current is 300 microamperes. That is the trigger current that reduces $T_1$'s base current to zero.

The current available for the external load at any voltage is the height of the NR curve at that voltage. If the load is largely capacitive, the available current can be integrated to find the time required to turn on $T_2$. Knowing $T_2$'s base current and $\beta$, $T_1$'s collector current can be integrated into its load capacity until the collector voltage changes enough to supply the threshold current to $T_1$. At this time the trigger can be removed and the regeneration can continue. Before this time has elapsed, the toggle will resume its former state if the trigger is removed. To this threshold time must be added the "built-in" delays in the transistors. If tubes are used, these built-in delays can be neglected.

Fig. 7 shows the trigger amplitude versus duration threshold for the typical toggle. Curve $A$ indicates the magnitude of the built-in delays in these particular transistors. Curve $B$ shows the similar curve with 1,000 $\mu$F on each collector. The difference at $t_i=300$ microseconds is 1.1 microseconds. Note the agreement between measured and calculated low-speed threshold.

Three hundred microamperes represents complete triggering in this case, since that is $T_1$'s base current. Above 300 microamperes, the trigger is removing stored carriers from $T_1$'s base, and this speeds up the transistors considerably. (Refer now to Figure 8A.) We will now calculate the additional trigger duration required when each collector is loaded with 1,000 microfarads ($\mu$F). The point of unstable equilibrium ($E_B$) is 2.6 volts in this circuit, so that is the voltage that $T_1$'s collector must descend to before the trigger is removed. The sequence of events can best be shown graphically, as in Fig. 8(A). At $t=0$, the current in $T_1$ is reduced, and the collector voltage increases as the load capacitor charges through $R_0$ (920 ohms). This voltage asymptotically approaches $-5.5$ volts in the case of complete triggering. When the collector voltage of $T_1$ exceeds 1.6 volts ($E_B$), $T_2$ begins to conduct, and its current increases with the same waveform that appeared on $T_1$'s collector, but with the obvious time displacement. The collector current of $T_2$ flows into a load consisting of $R_0$ and 1,000 $\mu$F in parallel. Eventually the collector voltage of $T_2$ will fall to $E_B$ ($-2.6$ volts), and regeneration will take over. At that instant, the trigger current can be turned off and regeneration will complete the change-of-state. Fig. 8(B) shows the essential steps in the calculation of minimum trigger time for the case of complete triggering. Note that $V_{CS}$ in Fig. 8(B) represents the change in $T_1$'s collector voltage from the initial state of $-5.5$ volts. The change
required is $5.5 - 2.6 = 2.9$ volts. Upon substituting, it is seen that $V_2$ reaches 2.6 volts after a delay of 1.07 microseconds, and this figure compares favorably with the 1.1 microsecond found experimentally and illustrated previously in Fig. 7.

Another good check point is near the point of marginal triggering. Notice the NR curve for adequate triggering in Fig. 9(A). The current is of the form $(E_{c}/R_{c}) (1 - e^{-t/R_{c}C})$ between $E=0$ and $E=E_c$. The voltage on the capacitor is obviously a simple $RC$ rise. From $E_c$ to $E_R$ there is another $RC$ rise, but here $R$ is negative, so the exponent is positive. In a sense, the current "approaches" $E_{c}/R_{c}$ as $t \to \infty$. Above $E_R$, the rise is again a simple $RC$, approaching $E_c$. For triggering levels just over the margin, the waveform on $T_1$'s collector should resemble the three exponentials. Fig. 9(B) shows the actual and predicted waveforms for two values of triggering current just over the minimum and for complete triggering.

THE EFFECT OF A COMMON EMITTER OR CATHODE RESISTOR

A common emitter resistor affords another path of regeneration and can therefore be expected to alter the shape of the NR curve. Fig. 10 illustrates this effect. The voltage drop across $R_E$, the common emitter resistor, has been added to the active circuit voltage and the NR curve of the composite circuit is shown dashed. It is assumed that the emitter return voltage is changed along with $R_E$ in such a fashion as to keep the emitter current the same as before at points $A$ and $F$.

When $R_E = -R_N$, the composite NR curve becomes vertical in the regenerative region. For larger $R_E$, the slope reverses in this region. It may seem surprising to have three consecutive intersections of the NR curve and the horizontal axis in which the NR curve has positive slope, and the casual observer may be led to believe that there are three points of stable equilibrium. Intersections with positive slope, however, are points of stable equilibrium only in systems that are open-circuit bistable, and in cases where $|R_N| > |R_E|$, it is both open-circuit and short-circuit bistable.

One was able to explore the NR curve experimentally when $R_E$ was small because the constant-voltage source constituted a vertical load line. That is, there was only one intersection of the (vertical) load line and the NR curve. But for large $R_E$, the current is no longer a single-valved function of voltage and the circuit is both open-circuit and short-circuit bistable. When $R_E$ is bypassed with a capacitor, the behavior becomes more complicated. For short input pulses, $R_E$ can be ignored, but for d-c loading, it must be taken into account.

SPECIAL CONSIDERATIONS WHEN USING TUBES

For tube circuits, the NR curve is not so easy to find analytically. The experimental method in the beginning of this paper can be used, although it is slow. The power supply can be replaced by a transformer, though, and the NR curve traced at a 60-cycle per second rate on an oscilloscope. Connect the plate voltage to the horizontal amplifier and insert a metering resistor to derive a current signal for the vertical amplifier. Then, the effects of variations in tubes, resistors, or voltages can be seen immediately. Since tubes go into or out of conduction more gradually than transistors, the corners of the NR curve will be rounded. This could be an advantage in analysis because the NR curve may be well represented by the simple cubic $I = K_1(E-E_D)+K_2(E-E_D)^3$ where $E_D$ is again the point of unstable equilibrium.

Conclusions

The usefulness of NR curves in toggle analysis has been demonstrated, and a method has been given for the derivation of such curves from the circuit parameters. The next step, apparently, is to turn things around and use the NR curves as the intermediate step in toggle synthesis. Given the requirements, output voltage and tolerance, load driving requirements, etc., it should be possible to draw an NR curve that fits the requirements, and then to derive the component values and tolerances directly from the NR curve. Whether this can be done in practice remains to be seen.

Direct Access Photomemory

Part I. Prototype Machine System

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IN RECENT years, the International Business Machines Corporation (IBM) research laboratory at San Jose, California, has sustained a group effort directed toward the investigation of a very-large capacity direct-access digital storage system using a unique self-developing photographic medium.

A feasibility model, designed around this film medium, with a $10^9$ alpha-numeric character capacity, has been completed. It was under taken purely as a research project to explore the capabilities of this medium, and there are no present plans to produce it. This machine, the Direct Access Photomemory, was used as a carrier to determine the problems associated with the radically different photo file as opposed to the more familiar magnetic systems. Early in the machine's development, design parameters were established with the primary purpose of forcing advanced technological development in all phases of the work. These were as follows:

- Capacity: $10^9$ alpha-numeric characters, consisting of $10^7$ 100-character records
- Storage density: 1,000 bits per lineal inch, equivalent track spacing of 0.006 inch
- Access time: one second maximum
- Input: 100-kc bit rate, asynchronous
- Output: 100-kc bit rate, asynchronous

Direct Access Photomemory

A description of the Direct Access Photomemory will reveal little resemblance in basic components to magnetic storage systems. Light sources, lenses, and a transparent film replace the more familiar magnetic head and oxide coated tape, disk, or drum surfaces. One of the major differences encountered from the point of view of system is the char-

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