RADIOTRONICS

Vol. 28, No. 10 October, 1963

IN THIS ISSUE

10

AN AWV PUBLICATION

REGISTERED IN AUSTRALIA FOR TRANSMISSION BY POST AS A PERIODICAL.

CERAMIC PICKUP CARTRIDGES

by B. J. Simpson

The introduction of the ceramic cartridge, and the developments that have taken place in that field in recent years, have brought to users a high-grade unit. Modern ceramic units, however, have posed new problems for the designer of amplifying equipment, particularly when transistors are used. This article describes the problems and some of the methods of dealing with them; it is the result of an investigation of the subject in the "Radiotronics" laboratory.

Ceramics

Most people when faced with this term think of fine china and pottery, and of course they are basically right. As far as the modern ceramic cartridge is concerned, however, the similarity ends with the good quality that is common to both the cartridge and the pottery. The fact is that the active element of the cartridge is allied in composition with the materials used in pottery, but is vastly different in its precise formulation.

In order to get the ceramic cartridge into perspective, it is first necessary to see where it fits into the general picture of record replay transducers. Many such devices have been produced over the years, but if we ignore such unusual types as magnetostriction, capacitance and eddy-current pickups, which as far as I know have vanished from the scene today, we are left with two basic types, the magnetic and the piezoelectric.

The term "magnetic" embraces all the varieties of electromagnetic and dynamic cartridges. The piezoelectric types, which we are to discuss here, originally consisted of what were loosely referred to as crystal pickups. A typical construction would consist of two slabs of a suitable material cemented together with an electrode between the two slabs and a second electrode in contact with the two outer faces of the assembly.

The slabs are cut in such a manner that a torque (in the case of a "twister" type assembly) or a flexure (in the case of a "bender" assembly) will produce a potential difference across the electrodes. Most pickups of this type have used the torsional system. The output from this type of device is directly proportional to the amplitude of the stylus displacement.

Earlier crystal units used sodium potassium tartrate (Rochelle salt). This material produces a high output, but is very susceptible to temperature and humidity. The humidity problem can be solved by sealing the unit completely, but this still leaves the temperature problem. Few units use this material today, if any. Other materials without these disadvantages have been used, such as ammonium dihydrogen phosphate, and generally produce a lower output.

In the days when these units just described were the only ones of their type available, it must also be remembered that we had neither stereo nor transistors. Most types would produce outputs of the order of 500 millivolts or more on programme, working into loads of the order of half a megohm or more. In general they had a fairly flat characteristic above the crossover point (around 500 cps), and a rising characteristic below that point. That is, they had a built-in equalisation for the lower portion of the recording characteristic. The upper portion of the characteristic was catered for partly by a natural falloff in output with rising frequency, and partly by a capacitor connected in parallel with the load.

This system worked in less critical days, when in any event we had only the old shellac 78's, and at a time when there was no standardisation of the record and playback characteristics. But in recent years, ceramic piezoelectric devices have appeared, using such materials as barium titanate in ceramic form. Apart from the basic principle of operation, the more recent ceramic units are so far ahead of the older crystal types as to be in a separate category.

Not only have the disadvantages of susceptibility to temperature and humidity been overcome, but the response and general quality of reproduction have made the newer ceramic cartridges a strong challenge to the entrenched position of the magnetic unit in the high quality field. This is particularly true when one considers the extremely low wear imposed by the later units, and the fact that (in valve amplifiers) the very sensitive amplifier required by magnetics is not required. Further, the later units in the ceramic field, at least those available to us from Britain, are so designed as to incorporate equalization, and produce a "flat" response into the recommended load.

Problems

As the quality of the ceramic cartridge has improved, the output level in general appears to have fallen, so that figures of 30 to 100 millivolts per centimetre per second are now being quoted for them. This is very much lower than the older crystal types gave us, but is also very acceptable in relation to the vastly improved quality. It is however a fact that must be remembered, and will be referred to again later.

The fact that the cartridge gives a "flat" response into the recommended load is an advantage, but here again, it appears in general that improvements in quality and even response have been attended by an increase in the value of recommended load. Few, if any, of the better quality units available today require a load of less than 2 megohms if the stated response in the bass region is to be obtained. This is another important point. Even with valve amplifiers, few of them will present 2 megohms or more to the

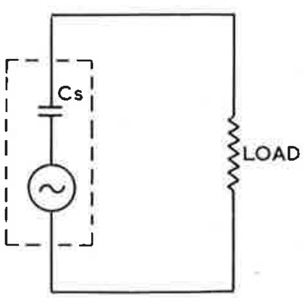


Fig. 1

cartridge, so that in fact some further equalization could be required.

When we turn to transistorized amplifiers, the last stated problem appears at first to be a large one, especially when it has to be solved in relation to cost. As will be seen later, not only is it difficult to present a load of the order required with economic transistor amplifiers, but the noise problem intrudes. As will be seen later, a completely different approach is required.

Equalisation

Before dealing further with the subject, it is necessary to digress a little into the subject of equalization, so that readers will be fully aware of what has to be done. The piezoelectric cartridge is to all intents and purposes a low-impedance generator with a series capacitor. The equivalent circuit is shown in Fig. 1. The usual mode of operation is to feed the cartridge into a resistor of sufficiently high a value that the current through the resistor, and the voltage drop across it, remain substantially constant at all audio frequencies of interest.

This is not quite true, but is a close approximation. In fact, the generator voltage will not be constant over the frequency range, but is usually arranged to rise towards the low frequencies, at which the reactance of the series capacitor is becoming large in relation to the value of the load resistance. It is immediately obvious that, apart from either accidental or designed variations in the generator output, the value of load resistor for reasonable response is determined by the value of the series capacitor Cs.

In some of the older types of crystal cartridge, Cs was quite large, values up to 0.002 mfd being on record; in cases of this kind, the recommended value of load resistor was much lower, perhaps of the order of 0.1 megohm or even lower. Later versions of crystal units had lower values of series capacitance, meaning higher load resistors in most cases, and today we are likely to meet values of the order of 500 picofarads or so, often meaning still higher values of load.

It is unfortunate that most makers do not quote the electrical equivalent circuit of the cartridge, particularly the capacitor Cs, as this could assist in those cases where it is not possible to show the unit the recommended value of load. If, for example, a unit with Cs = 500 pf and recommended load = 2 megohms is required to work into 0.5 megohm, a much more likely value in practice, circuit elements could be adjusted accordingly.

For example, if we assume that we have one of the newer units with a "flat" response into the recommended load, then the time constant of the circuit, from the figures just quoted, is 1,000 microseconds. This time constant should still obtain if the unit is fed into a lower value of load, and can be adjusted to the required value by connecting a capacitor across the load.

For example, if the unit is to feed into a 0.5 megohm load, then a capacitance of 2,000 pf is required in the circuit, and this can be made up by connecting a 1,500 pf capacitor in parallel with the load. This should reasonably maintain the correct response, although it will be seen in the example chosen that from a listening point of view there would be many people who would not notice anything amiss if the additional capacitor were left out. However, if we are going for quality, then we should do the right thing right down the line.

Because the piezoelectric unit is a capacitive device, placing pure capacitance in shunt across the unit will have no effect on performance, but will lower the effective output of the transducer. Note that the 1,500 pf capacitor mentioned in the previous paragraph was not a pure capacitance, as it was in shunt with other elements. This is the reason why quality of a piezoelectric unit is not affected by long screened leads.

It will be seen that connecting the ceramic units into valve amplifiers is no problem, even if the equalization does have to be adjusted a little for the optimum results. It must however be realised that the example chosen was based on one of the later units with a "flat" response. This does not apply to all the units available today, some of which require in any case external equalization as stipulated by the makers. In this type of case when a lower value of load resistance is required, the stated values of the equalising components may be adjusted following the theory outlined.

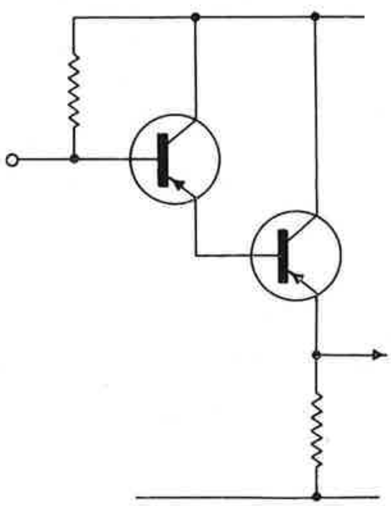


Fig. 2

Transistors

The main area in which trouble has been experienced in using ceramic units has been in the field of transistor amplifiers. It will readily be seen that the recommended loading is extremely high for transistor units, especially when economics have to be considered. There are a number of alternative approaches to the problem, and the more important of these will be dealt with briefly below. They are:—

1. Presentation of the recommended resistive load,
2. Presentation of a lower or much lower resistive load,
3. Conversion to a velocity characteristic,
4. Presentation of a capacitive load.

Recommended Load

The presentation of a load of 2 megohms or so when using transistors is not easy, and will certainly require at least two stages. Two common-collector stages in cascade is probably the most obvious first line of enquiry, similar to the circuit sketched in Fig. 2. There are other ways of doing this of course, but in most cases the disadvantages are low to zero gain and a noise problem. From the investigations made so far it would appear that this solution is not practicable. Much better results could be obtained with the use of a field effect transistor in the input stage, but these units are not yet available in the quantities and at the prices that would make them an attractive solution for consumer equipment.

Lower Loads

As mentioned previously, the presentation of a lower value of load resistor than that recommended will result in a loss of bass response unless measures are taken to equalize the response. The situation, as one may expect, becomes more acute the larger the departure from the recommended value. Further, a stage will be reached where the lower effective input level to the amplifier will call for more gain.

The lack of precise data on the characteristics of a pickup means that if this resort is to be adopted, then the characteristics must be measured, or the required degree of equalization must be determined experimentally. In any case, experimental verification would be required, because the load presented to the cartridge would vary with frequency, as would the generator output, both with frequency and due to the load variation.

The situation is therefore complex, although susceptible of fairly easy solution when a tone-band test record is available, together with the necessary measuring equipment. The point is that this is a laboratory exercise, and cannot be solved by simple rule of thumb methods in the home.

There are two other points worthy of mention in this regard. One is that the generator output, as opposed to the cartridge output, rises in the lower bass region in order to preserve the "flat" characteristic. The amount of this rise varies with different makes of unit, as one may suppose, and when a lower value of load is used, the rise is also affected by this factor. The second point is that one common method of "holding up" the treble response of the unit is to employ a controlled resonance, which is incorporated in the design of the unit. This appears in most cases to reside in the 6 to 10 Kc region, and does not of course materially affect performance, being very smooth and causing a rise in output that is within the stated characteristics of the cartridge. It is necessary to remember that it is there when checking frequency response.

One very simple approach to the presentation of a lower load value does not in fact do that. What is done is to prepare an amplifier with an input impedance of say 0.1 megohm, which can be done with transistors if required, and then to "build out" the impedance by installing a series resistor in the pickup lead so that the total becomes the required value. When this method is used with a valve amplifier, where the input impedance could be say 0.5 megohm, then the series resistor required would be 1.5 megohms, and the insertion loss 12 db. This situation is generally workable, particularly if the pickup chosen has an adequate output. See Fig. 3.

When we try to do this with transistors, however, where we are likely to be limited for practical and economic reasons to a maximum input impedance of 0.1 megohm, then the required series resistor would be 1.9 megohms and the loss 26 db approximately. It can readily be seen that this is hardly likely to recommend itself for quality work, as an acceptable signal to noise ratio is hardly likely to be achieved. So far, this line of enquiry does not appear to show any promise along either of the methods described.

Velocity Characteristic

One very intelligent approach to the problem of combining ceramic cartridges and transistor amplifiers has been used in the U.S.A. and by at least one U.K. manufacturer. This method is particularly appropriate for the higher-quality lower level cartridges, and consists of applying the cartridge output through a passive network to the standard magnetic input terminals of an amplifier. The passive network converts the amplitude or near amplitude characteristic of the cartridge to a velocity characteristic, and also of course involves a reduction in level. The signal is then equalised by the standard replay characteristic provided for the magnetic cartridge. In the U.S.A., small adaptor units are available containing the passive network, one of course for each channel.

This is a very practical suggestion when one considers matching up various units of commercial equipment into a system. It does however seem a pity that double processing of the signal is involved. Further, the network required will vary with the make of the cartridge, so that it is not possible for the maker of the amplifier to build in the necessary circuit, or to provide an alternative equalising network for ceramic cartridges.

When we turn to the home constructor, in whom we are especially interested in these pages, then he usually has a pickup of one type or the other, and duplication of facilities is therefore not required in most cases. If he wants to use a ceramic cartridge, surely it is better to have two alternative circuits and let him build one or the other, not to provide all facilities for a magnetic cartridge and then add on another circuit still.

Capacitive Load

So far in the discussion we have seen that one acceptable solution is available, other than the provision of the recommended load. It now remains to be seen what can be done with a completely different approach.

We know that the ceramic cartridge is a capacitive device. If the cartridge has a "flat" characteristic, we know that we can place pure capacitance across the output without affecting the frequency response; all we will do will be to reduce the effective output. Incidentally, we also know that the transistor, being essentially a low-impedance device, is better used that way.

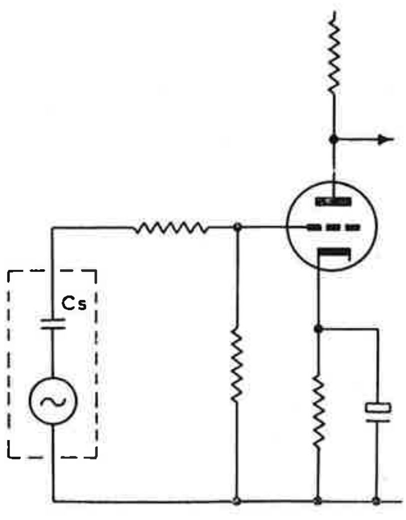


Fig. 3

One method tried, partly for interest and partly to prove the point, was to shunt the cartridge with a large capacitor, very large in relation to the probable internal capacitor Cs. A 0.47 mfd unit was used, and the output measured across the shunt capacitor was astonishingly flat, and certainly better than the measured response of the same cartridge working into the recommended load. For reasons which will appear later, this line of enquiry was not pursued further, so that the relationship between the two measured responses could have been anomalous.

This however was the possible basis of a system, but it would require a high-gain low-noise amplifier. Because the noise in an amplifier is derived almost entirely from the first stage, and this system would provide an extremely low impedance across the input, it is conceivable that a satisfactory signal to noise ratio could be obtained, provided the gain required could be made economic. So this is possibly something to be looked at again later. See Fig. 4.

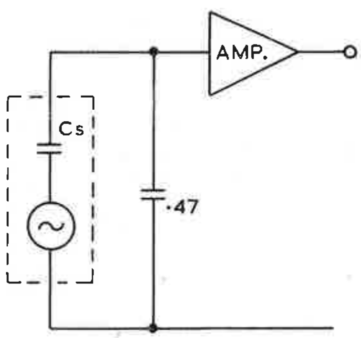


Fig. 4

The apparent promise here resulted in a decision to look further into the question of a circuit which, by using heavy feedback that is substantially capacitive, presents a capacitive circuit to the cartridge. One such circuit was used in these pages over a year ago in conjunction with a transistorized stereo amplifier. The basic configuration of the circuit is likely to look something like Fig. 5. Here the collector-to-base resistor supplies bias and contributes very little feedback, but the capacitor in parallel provides heavy capacitive feedback. The desired effect is to make the input of the amplifier look like a capacitor to the cartridge.

The extent to which this object is achieved depends among other things on the characteristics of the cartridge, remembering that the input capacitance of the amplifier, or if you like the virtual capacitor connected across the input of the amplifier, is not a pure capacitance. The input of the amplifier is capacitive at all frequencies at which the reactance of the virtual capacitor is low in relation to the shunt impedance resulting from the rest of the circuit.

The reactance of the virtual capacitor must also be low at all frequencies of interest compared with Cs. As one may expect, the problem lies in the lower bass region. In this circuit, frequency response can be improved at the cost of gain by increasing the value of the feedback capacitor, that is, by making the virtual capacitor larger. Readers will readily see the similarity between this situation and the interesting little experiment explained earlier.

It is also of importance that the value of the virtual capacitor is of course limited by the amount of feedback available, that is, by the gain of the transistor. A high gain transistor with a gain of 80 or more is required in this position. It will also be realised that the actual signal level at the base of the transistor will be very small, and although the low input impedance will assist in getting an acceptable signal to noise ratio, a low-noise transistor should obviously be used.

The value of the virtual capacitor (those brought up on valves will call it Miller capacitance) is the product of the feedback capacitor value and the voltage gain of the stage. If the precise characteristics of the pickup including the value of Cs are known, the signal level at the base of the transistor can readily be calculated, being Cs/Cv + Cs times the pickup generator voltage, where Cv is the value of the virtual capacitor. The lower cutoff frequency is determined by the expression $1/2\pi$ CsRi, where Ri is the input resistance of the stage.

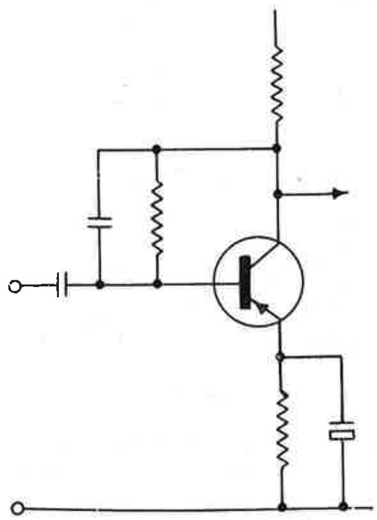


Fig. 5

Circuits of this kind that have been published in various places have caused some confusion. Firstly, it is difficult for the layman to find out, for example, what the value of Cs is. Secondly, there is good reason to suppose that the complex behaviour of the cartridge when applied to a low value of load (which in this case cannot be a pure capacitance) means that the simple relationships quoted above do not always hold in practice. Further, whilst some cartridges have a "flat" response under normal conditions, others do not. It would appear likely, for example, that one of those cartridges that have a rising output characteristic below the 500 cps point would be easier to cater for than a "flat" cartridge.

Because we were anxious to publish circuits suitable for ceramic units, and to ensure that they would in fact result in high quality reproduction, a complete investigation of this type of circuit was decided on as being the most likely solution to the basic problem of matching ceramics with transistors. In the course of this work, a standard basic circuit has been developed, together with data on the feedback circuits required for the more popular types of ceramic cartridge on sale in Australia. At this time, no effort was made to cover rarer types not appearing in quantity in this country. At the same time, it must be admitted that no one knows everything that is happening, and it may well be that there are other good pickups and cartridges on sale in this country that were not selected or named due to oversight. If this is so, I will be happy to check them, to recommend a suitable circuit, and to publish the data in these pages, on application by the maker or distributor. Whilst on this subject, it must be pointed out that pickups and cartridges named in this article were selected as representative types, and no endorsement of these units, beyond the facts stated, is necessarily implied by this article.

Practical Circuit

The practical circuit that has been evolved for this type of application is shown in Fig. 6. It consists of the basic circuit of Fig. 5, but with some re-arrangement, together with a direct-coupled amplifying stage following. The extra stage was added because the levels in the first stage are quite low, and it was felt that it would be better for the first amplifier after the input stage to be regarded as an integral part of the circuit. At the same time, the addition of the direct-coupled stage allowed the saving of three components in the first stage as originally designed.

In this version of the circuit, the feedback capacitor is a 0.047 mfd component, whilst in other versions the feedback circuit is modified according to the cartridge. Details of these changes are given later. The 0.01 capacitor at the input is there merely to isolate the cartridge from dc. Bias for the first stage is derived via the 10K resistor and a tapping on the emitter resistor of the second stage.

In order that this circuit could be used with other arrangements under preparation, an output level of about 100 millivolts minimum with heavy

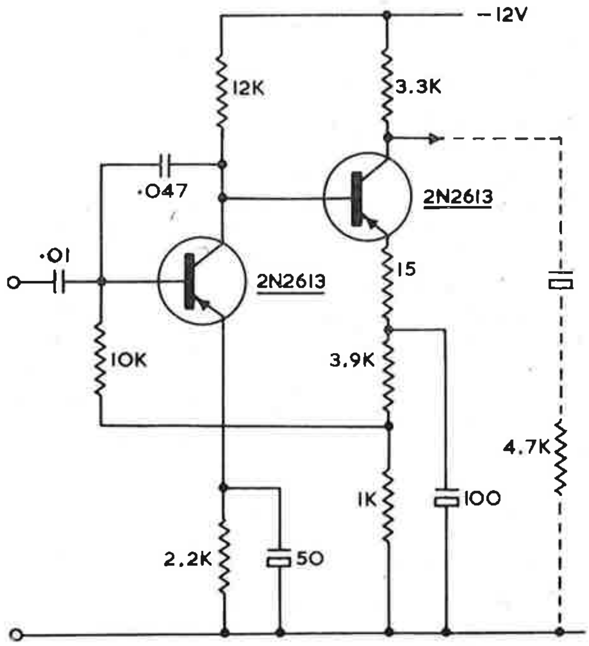


Fig. 6

programme material was aimed at, and easily exceeded with all the cartridges tested. At the same time an ample margin for overload at the input is provided for.

The signal to noise ratio of the circuit shown is more than adequate at better than —60 db, remembering that the noise output, as is usual with transistor circuits, is all white noise. It is my opinion, based on many listening tests, that where the noise output of an amplifier is all white noise, it sounds about 10 db better than a noise output with the usual discrete frequency components with which we have been familiar in the past, even when the measured signal to noise ratio is the same. This is a reasonable proposition, although the actual factor will depend on a comparison of any two specific units of equipment.

It is worthwhile remembering this when examining specifications of transistorized amplifiers. It explains why makers sometimes are contented with a signal to noise ratio of between —50 and —60 db in complete systems, whereas valve units, which are claimed to be noisier, can easily equal these figures on paper.

As one might expect from a circuit in which there is a great deal of feedback, the distortion is low. Due to the nature of the circuit, distortion measurements were made with a signal input which would produce an output of 150 millivolts at the output. As already indicated, this is about the order of signal level for which the circuit was designed, although an output considerably in excess of that figure can be achieved with higher inputs. The input signal was supplied from a low-impedance source with a 500 pf capacitor in series to simulate a pickup, and the distortion measured at not greater than 0.1% total.

Tests with Pickups

Because of the interdependent variations in levels and similar considerations already touched upon, the procedure in determining the more important circuit elements was to decide on an approximate value and then to determine the final value experimentally with the circuit driven with the chosen pickup or cartridge. In this way, the variations mentioned could be automatically taken care of, and the required results obtained quickly and accurately. Tests were carried out with two ceramic cartridges and one ceramic pick-up, as listed:

1. Decca "Deram" transcription cartridge,
2. Acos turnover cartridge type GP81-1N,
3. Acos "Hi-light" pickup.

The first mentioned of these components, in the version used, is a microgroove stereo cartridge with standard mounting centres. It is one of two similar units, the transcription version which was used in the tests and a "changer" version. As the differences between the two units are playing weights and general quality, and not the basic mechanism, recommendations made for the "Deram" transcription unit will also apply to the alternative version of this cartridge.

The second component is a turnover unit with provision for playing microgroove (stereo) and 78 records. In this case tests were carried out only on the microgroove side for convenience; the same circuit would of course apply for the 78 side.

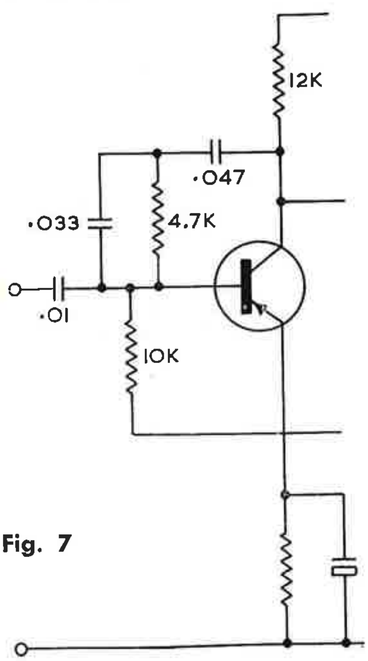


Fig. 7

The third unit is an integrated pickup with interchangeable heads for microgroove stereo and microgroove mono. It is understood that a further head is available if required for playing 78's.

In each case, the component was first checked against the makers' specification, not so much to check the figures as to make sure that the unit being used had not been damaged in any way. In all cases this test was carried out into the high-impedance load recommended in the instruction sheets, and in all cases the specified figures were found to be correct. The three units are discussed in further detail below.

The aim in determining the circuit values to use these cartridges with a capacitive input circuit was to achieve a frequency response equal to or better than that claimed for the same unit working into the normally recommended high-impedance load. This was achieved in all three cases. There was some small evidence that the use of the capacitive circuit had some effect in smoothing out the overall frequency response. Time has not so far permitted this to be investigated further at this time.

Decca "Deram"

The transcription cartridge used has a stated and verified frequency response of 40 cps to 12 Kc ± 3 db working into 2 megohms, whilst the "changer" version has a similar frequency response. The outputs are respectively 80 mv/cm/sec and 50 mv/cm/sec. The transcription unit has higher vertical and lateral compliance, lower tip mass and a lower recommended playing weight. The recommended circuit for both is as shown in Fig. 6, which, with the unit tested, produces a frequency response of 30 cps to 15 Kc ± 3 db, that is, better than claimed.

On the heaviest programme material, the transcription unit gave an output from the circuit in excess of 200 millivolts, whilst the output with the "changer" version is about 30% lower, as one would expect. The target output of 100 millivolts is easily reached.

Acos GP81-1N

This cartridge has a frequency response of 100 cps to 12 Kc ± 3 db into a 2 megohm load, and an output of 90 mv/cm/sec. The recommended capacitive input circuit is as shown in Fig. 6, except that a modification is required to the feedback circuit. The 0.047 mfd capacitor remains, but two other components are added, as shown in Fig. 7. Heavy programme material produced an output in excess of 200 millivolts, and a frequency response of 70 cps to 12 Kc $\pm$ 3 db was achieved.

Acos "Hi-light"

This is an integrated pickup with a frequency response of 20 cps to 20 Kc $\pm$ 3 db and an output of 30 mv/cm/sec into a 2 megohm load. The recommended circuit is as shown in Fig. 6, and produced a frequency response with the unit under test of 20 cps to 20 Kc $\pm$ 2 db, which is better than claimed. The output of this unit is lower, and the available output from the circuit on heavy programme material exceeds 100 millivolts, which was the target figure.

Summary

This then is a description of one answer to the problems put forward in the early part of the article, and it appears from several points of view to have notable advantages over alternative systems. As previously pointed out, it may in some cases be necessary to modify the feedback arrangements to suit a particular pickup. For this reason it is again pointed out that tests have been carried out only with the units mentioned herein, and that whilst the circuit would obviously be suitable for use with any unit provided any necessary adjustments were made, data regarding the use of the circuit with other units is presently not available.

A warning must also be given with regard to testing of circuits of this kind. Because the circuit is intended to be fed from a particular type of source and is adjusted to compensate for variations of equivalent generator output with frequency, it is difficult, unless all the characteristics of the cartridge are known, to simulate a frequency check on the circuit. Testing along customary lines will produce a frequency response far from that achieved in practice with a cartridge input. Testing is normally carried out using the chosen cartridge and a tone-band test disc.

It has been suggested that where the output of a pickup or cartridge is ample, then the frequency response could be improved at the expense of gain, as indicated earlier in the article. This is not entirely true, because if this is done, the response in the bass region rises progressively with respect to the middle and highs, besides of course extending further downwards. Not only does this then produce a frequency response which is departing more and more from a linear response as the degree of feedback is increased over what appears to be the optimum, but there are indications that if this is carried too far, then trouble will result from instability in the sub-sonic region. As it is, there is some indication that with this type of circuit, when incorporated into a complete amplifier, somewhat better decoupling will be required than may otherwise be considered adequate. As the circuits shown stand, they are adjusted for optimum frequency response, and no improvement can be expected from an alteration in the degree of feedback employed.

During normal operation, the maximum signal level at the first transistor base is likely to be of the order of 1 to 3 millivolts, and care should be taken to keep leads as short as possible. Adequate shielding of the input leads will be required, and normal care must be exercised in general layout and grounds. As previously mentioned in these pages, optimum noise performance from transistorized low-level stages is obtained only by correct layout. Stages of this kind must have the connections to the supply rail returned to the same point on the rail, and the connections to ground all connected to the earth bus at the same point. Where direct-coupled stages are used, all leads should be treated as belonging to the same stage.

All resistors in the circuit are half-watt rating, and 10% components may be used. Electrolytics rated at 12 V.W. were used, and the 0.01 and 0.047 capacitors were 25 V.W. miniature ceramic components.

It would be wrong to think that the ceramic cartridge has replaced the crystal unit in the piezoelectric field, as there are many good quality crystal units available. In a "follow-up" article next month, we expect to comment on adaptations of the circuits described in order to use crystal cartridges and pickups.

A Series on Tunnel Diodes

3: SWITCHING

Because of its high frequency capabilities and small power requirements, the tunnel diode exhibits excellent switching characteristics. At the present time, tunnel diodes are being widely used as ultra-fast rise-time pulse generators and in ultra-fast counter circuits. They are also especially promising for digital-computer applications in which they offer speeds up to several hundred times faster than those available with transistor circuits. If speeds equivalent to those of transistors are satisfactory, tunnel-diode circuits can be designed which require only a small fraction of the power consumption of the best transistor micro-power circuit.

Switching Theory

Fig. 24 shows a simple tunnel-diode switching circuit which has a high-impedance input and a constant load resistance. The load line for this circuit and its intersections (points 1 and 3) with the typical tunnel-diode characteristic curve are shown in Fig. 25. When the power is first turned on, point 1 is the circuit operating point. When a step input of current $\triangle I_{in}$ is applied, the load line shifts, as shown by the dashed line in Fig. 25, and the operating point switches from point 1 to point 2. The **forward-current gain** G for the circuit is the ratio between the increase in current to the load ($I_{out2} - I_{out1}$) and the increase in the input current ($\triangle I_{in}$), as follows:

$$G = \frac{I_{out2} - I_{out1}}{\triangle I_{in}} \tag{13}$$

Because the value of $I_{out2} - I_{out1}$ can never be greater than the value of $\triangle I_{in} + I_P - I_V$, the maximum current gain G_{max} is given by

$$G_{max} = 1 + \frac{I_P - I_V}{\triangle I_{in}} \tag{14}$$

Eq. (14) indicates the importance of a large peak-to-valley-current ratio in the determination of the ultimate gain of tunnel diodes.

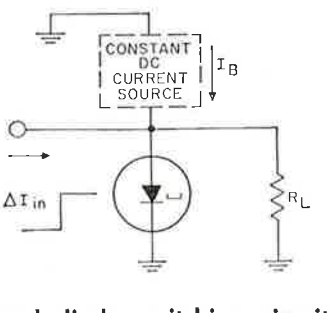


Fig. 24—Tunnel diode switching circuit loaded with linear resistance.

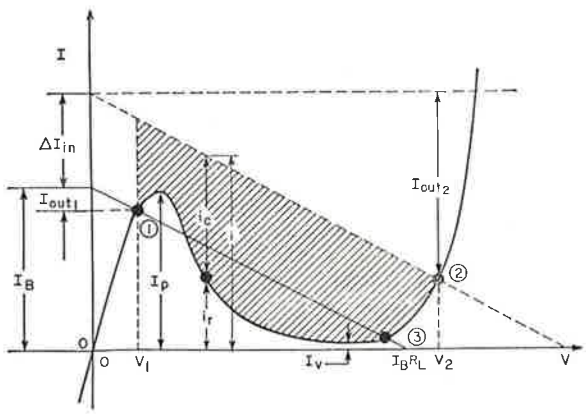


Fig. 25—Tunnel diode characteristic with switching load line.

The **switching speed** of a tunnel diode is determined primarily by its junction capacitance and negative resistance. In the equivalent circuit shown in Fig. 11, the series resistance R_s and the junction resistance R_j determine the static characteristic of the tunnel diode. The series inductance L_s and the junction capacitance C_j together with

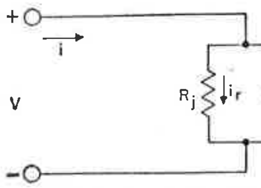


Fig. 26—Simplified tunnel diode equivalent circuit.

R_s and R_j, limit the transient or switching response. The series resistance is generally negligible in switching; if spurious oscillations and recovery time are not considered, the series inductance may also be neglected in the determination of rise time or switching speed. Fig. 26 shows the simplified equivalent circuit for tunnel-diode switching.

The junction capacitance C_j may be considered to be constant during switching. Therefore, the switching speed or rise time of the tunnel diode in switching from point 1 to point 2 of Fig. 25 depends solely on the amount of charging current i_c passing through the junction capacitance. This current is the difference between the input current i to the tunnel diode and the current i_r flowing through R_j. The currents i, i_c, and i_r for a particular point during switching are shown in Fig. 25. The **rise time** t_r of the device during switching from point 1 to point 2 is given by

$$t_r = C_j \int_{V_1}^{V_2} \frac{dV}{i_c} \tag{15}$$

where $i_c = i - i_r = (I_B + \triangle I_{in}) - \frac{V}{R_L} - i_r$

The fastest switching is accomplished when a constant-current lead line is used (i.e., when R_L is very large). Under these conditions, and if the overdrive is small relative to I_P, then $I_B + \triangle I_{in}$ is approximately equal to I_P, and Eq. 15 can be written as follows:

$$t_r \cong C_j \int_{V_P}^{V_F} \frac{dV}{I_P - i_r} \tag{16}$$

Because under these conditions i_r is generally small in comparison to I_P for most of the switching cycle, it can be approximated as a constant I_V. The equation for rise time can then be simplified as follows:

$$t_r \cong C_j \int_{V_P}^{V_F} \frac{dV}{I_P - I_V} = C_j \frac{V_F - V_P}{I_P - I_V} \tag{17}$$

The value of $V_F - V_P$ is approximately 0.5 volt for germanium and one volt for gallium arsenide.

If the small constant current I_V is eliminated from Eq. (17), the following useful relationships can be obtained:

$$\text{for germanium } t_r \cong \frac{C}{2\ I_P} \tag{18}$$

$$\text{for gallium arsenide } t_r \cong \frac{C}{I_P} \tag{19}$$

In both these relationships, constant-current load-line switching is assumed.

Tunnel-diode **fall time,** i.e., switching time from a high to a low state, is calculated in a manner similar to that described above for rise time.

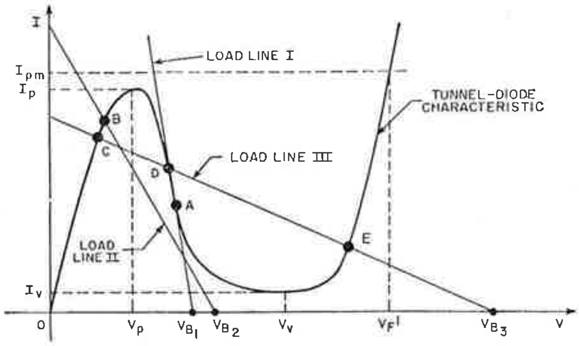


Fig. 27—Tunnel diode characteristic showing three basic types of biasing modes.

The series inductance L_s has very little effect on rise and fall times, especially if constant-current switching is used. However, the series inductance does affect **recovery time** and, therefore, limits the maximum repetition rate of switching. For example, when a tunnel diode is switched from the peak point to the high-voltage region, the current must then be reduced to the valley point before the diode can be switched to the low-voltage region. The time required to reduce the current to the valley point is limited by the time constant L_S/R_D, where R_D is the average value of the diode resistance in the high-voltage region. The recovery time in the low-voltage region is affected in the same manner.

Multivibrator Circuits

When suitable biasing is used, tunnel diodes may be utilised in astable, monostable, or bistable modes of operation. The dc load lines required for these three modes are shown in Fig. 27, and a circuit for obtaining the required biasing is shown in Fig. 28. The dc load line I, which intersects the tunnel-diode characteristic at point A in the unstable negative-resistance region, provides biasing for the **astable** mode. Load line II, which intersects the tunnel-diode characteristics at the stable point B, provides biasing for the **monostable** mode. Load line III, which intersects the tunnel-

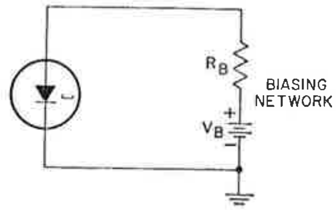


Fig. 28—Tunnel diode biasing circuit.

diode characteristics at the two stable operating points C and E, provides biasing for **bistable** operation (point D is unstable and is not used).

Astable Multivibrator

The design of tunnel-diode astable multivibrators or **relaxation oscillators** uses the stable biasing mode shown by the dc load line I in Fig. 27. A circuit of this type is shown in Fig. 29a; V_{B1} and R_{B1} provide the necessary load line, and inductance L_1 controls the frequency of operation. Three criteria must be met for this circuit to perform properly: (1) the dc load resistance R_{B1} must be less than the minimum negative resistance of the tunnel diode, i.e., $R_{B1} < R_{min}$; (2) the load line must intersect the tunnel-diode characteristic in the negative-resistance region, i.e.,

$$V_P < (V_{B1} - I_D R_{B1}) < V_V,$$

where I_D is the static tunnel-diode current; (3) the total inductance must be large enough so that the biasing point is unstable under ac conditions, i.e.,

$$(L_1 + L_3) > (R_{B1} + R_S) R_{min} C_j.$$

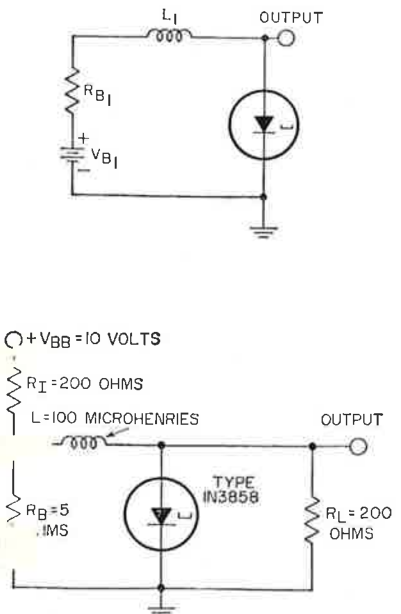


Fig. 29—(a) Basic tunnel diode astable multivibrator, and (b) practical circuit.

The highest frequency of operation for this type of circuit, therefore, is limited by tunnel-diode inductance, capacitance, and negative resistance.

Fig. 29b shows a practical example of a tunnel-diode astable multivibrator circuit. The frequency of oscillation for this circuit is 100 megacycles. The equivalent dc-load resistance R_{B1} is equal to $(R_B) \times (R_I)$ divided by $(R_B + R_I)$, or 4.875 ohms. The equivalent supply voltage V_{B1} is equal to $(V_{BB}) \times (R_B)$ divided by $(R_B + R_I)$, or 244 millivolts.

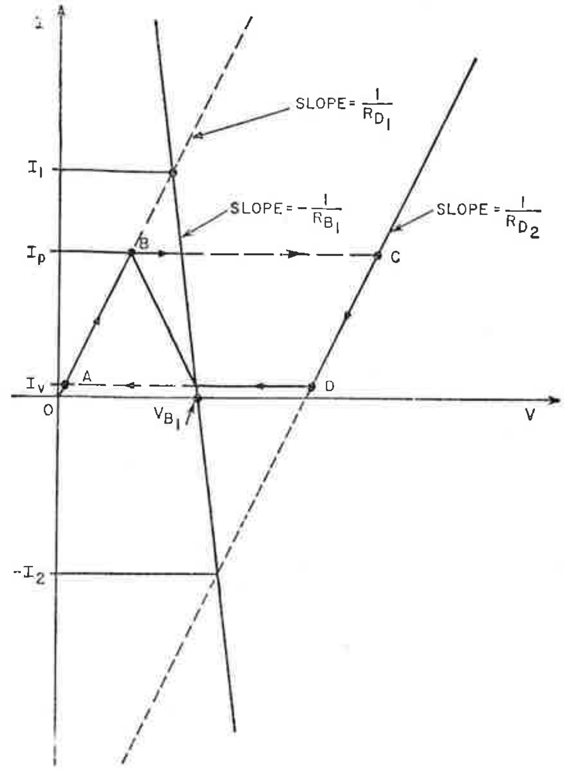


Fig. 30—Astable circuit operation shown on linearized tunnel diode characteristic.

The linearized tunnel-diode characteristic shown in Fig. 30 illustrates the switching trajectory and allows an approximate determination of the **repetition rate.** When L_1 is assumed to be large enough to permit negligible current change during switching, the switching trajectory given by path ABCD in the diagram is obtained. The time during which the tunnel diode is in the low-voltage state AB is determined by the charging of L_1 through R_{D1} and R_{B1}, as follows:

$$t_{AB} = \frac{L}{R_{B1} + R_{D1}} \ln \frac{I_1 - I_V}{I_1 - I_P} \qquad (20)$$

where R_{D1} is the diode resistance in the low-voltage state and I_1 is as defined in Fig. 30. The time during which the tunnel diode is in the high-

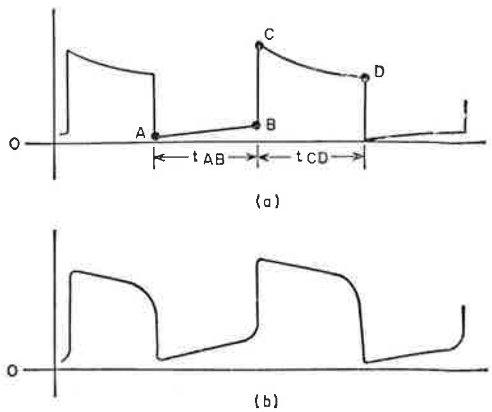


Fig. 31—Astable multivibrator waveforms for (a) linearized tunnel diode, and (b) actual tunnel diode.

voltage state CD is determined by the discharging time of L_1, as follows:

$$t_{CD} = \frac{L}{R_{B1} + R_{D2}} \ln \frac{I_2 + I_P}{I_2 + I_V} \tag{21}$$

where R_{D2} is the diode resistance in the high-voltage state and I_2 is as defined in Fig. 30. Because these times are much longer than the time spent in the negative-resistance region, the repetition rate, or frequency f, of the astable circuit is given by

$$f = \frac{1}{t_{AB} + t_{CD}} \tag{22}$$

Fig. 31a shows the output waveform calculated for the linearized tunnel-diode astable multivibrator. An actual tunnel-diode output appears as shown in Fig. 31b. As shown, the time spent in the negative-resistance region is very short.

Monostable Multivibrator

Tunnel-diode monostable multivibrators may be designed in a manner similar to that described for the astable multivibrator. Monostable operation is obtained when the biasing characteristic or dc load line intersects the tunnel-diode voltage-current curve at only one stable point, either below the peak point or above the valley point, as shown in Fig. 32a. A typical monostable multivibrator circuit is shown in Fig. 32b; waveforms for the two modes of biasing are shown in Fig. 33.

For proper "one-shot" operation, the input pulse width t_{in} must satisfy the following relationship:

$$t_r < t_{in} < t_w \tag{23}$$

where t_w is the **delay time,** or pulse width of the output waveform, and t_r is the tunnel-diode rise time or switching time.

The pulse width is controlled by the size of the inductance L_2, the resistance R_{B2}, and the resistance presented by the tunnel diode. When the idealized tunnel-diode characteristic shown in Fig. 34a is used, the approximate pulse width may be calculated for a condition of biasing below the peak-point, as follows:

$$t_w = \frac{L_2}{R_{B2} + R_{D2}} \ln \frac{I_2 + I_D + I_{in}}{I_2 + I_V} \tag{24}$$

where the currents I_2, I_D, I_{in}, and I_V are as given in Fig. 34a.

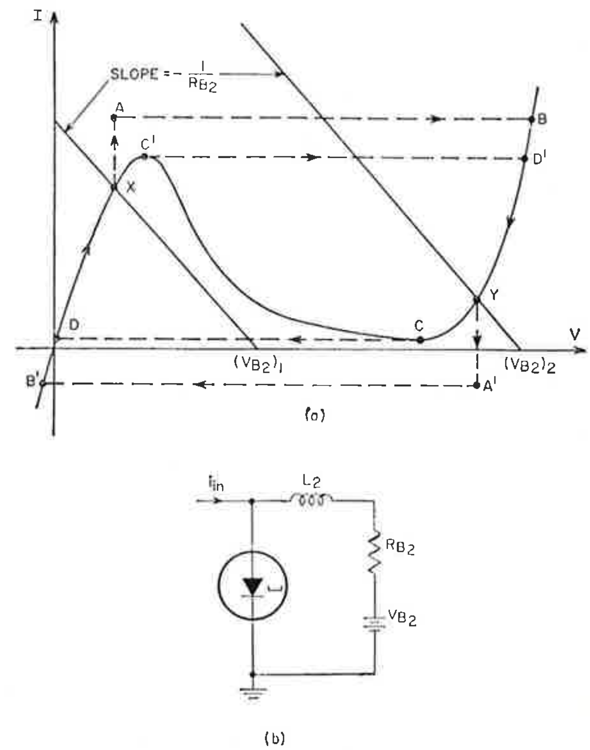


Fig. 32—(a) Biasing characteristic, and (b) monostable multivibrator circuit.

Fig. 34b shows a practical example of a monostable multivibrator circuit which uses biasing below the peak point. This circuit has a pulse width, or delay time, of five nanoseconds and a maximum repetition rate of 70 megacycles per second. The equivalent dc load resistance R_{B2} is 9.75 ohms, and the equivalent supply voltage V_{B2} is 250 millivolts.

Bistable Multivibrators

Because of their inherent memory capabilities, tunnel diodes can be readily used in the design of bistable multivibrator circuits. Fig. 35a shows a simple tunnel-diode bistable **(flip-flop)** circuit

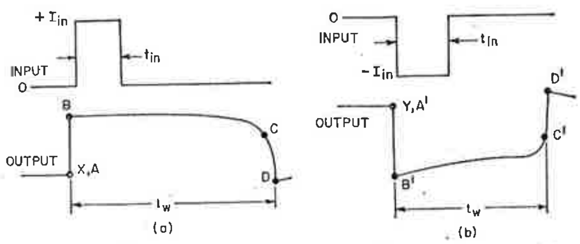


Fig. 33—Waveforms for tunnel diode monostable circuit at (a) biasing below the peak point, and (b) biasing above the valley point.

and its biasing and waveform characteristics. This circuit requires a positive pulse to set the flip-flop and a negative pulse to reset. A practical example of a set-reset bistable multivibrator circuit is shown in Fig. 35b. This circuit has a maximum repetition rate of 200 megacycles per second, an R_B resistance of 29.2 ohms, an equivalent load-line resistance R_{B3} of 27.5 ohms, and an equivalent supply voltage V_{B3} of 542 millivolts.

Fig. 36a shows a flip-flop circuit which can be set and reset with pulses of the same polarity.

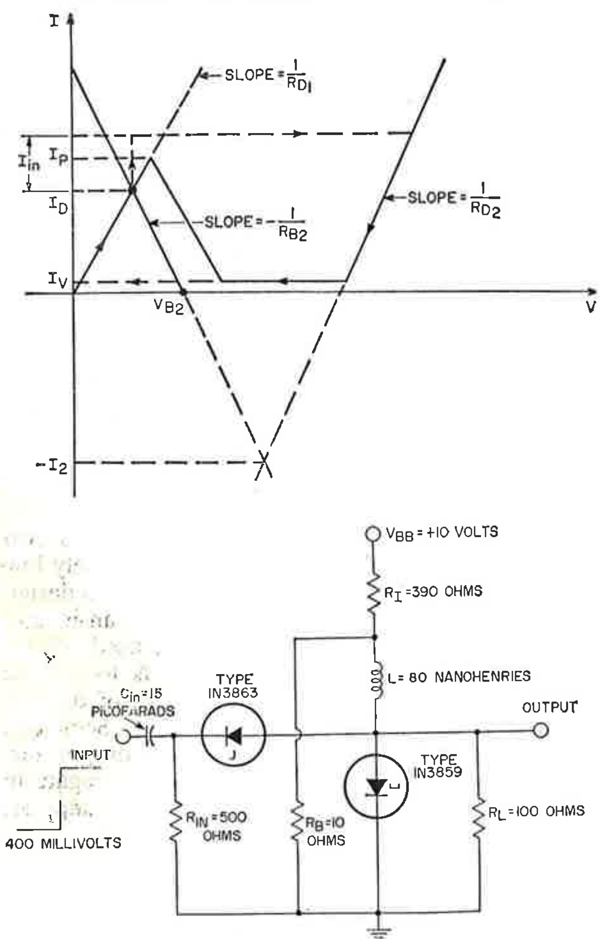


Fig. 34—(a) Monostable operation, and (b) practical monostable multivibrator.

In this circuit, the size of L_3 is determined by a compromise between ease of reset and recovery time of the circuit. Fig. 36b shows a practical example of a set-reset circuit which uses positive triggering. This circuit has a maximum repetition rate of 100 megacycles per second, a load-line resistance of 28.5 ohms, and a load-line voltage of 510 millivolts.

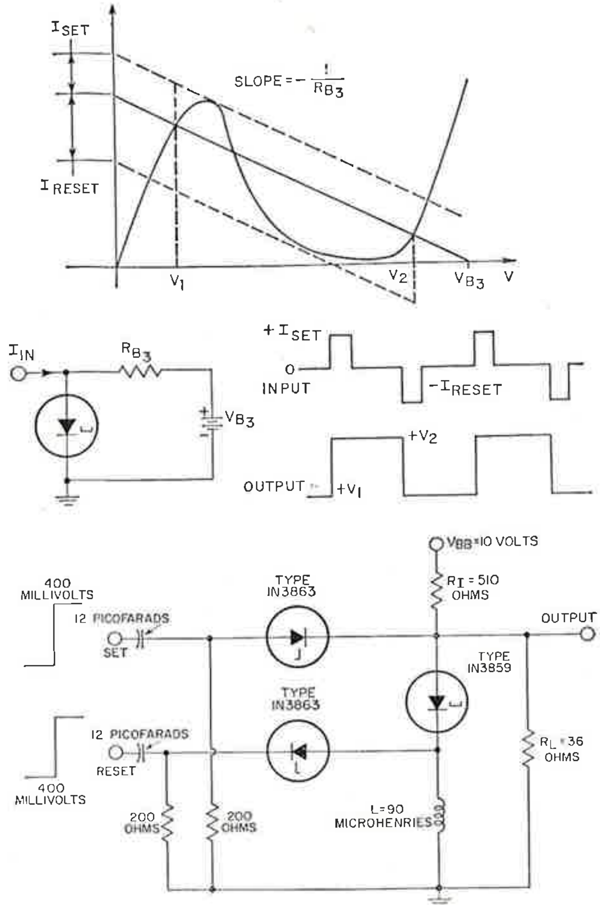


Fig. 35—(a) Simple bistable characteristic, circuit and waveforms, and (b) practical set-reset bistable multivibrator circuit.

The design of a triggered flip-flop circuit[6] using tunnel diodes is somewhat more complicated, as shown in Fig. 37a. Again, the value of the inductance L_3 is determined by a compromise between good triggering and fast recovery time. A practical example of such a circuit is shown in Fig. 37b. This circuit, which has a maximum repetition rate of 100 megacycles per second, may be cascaded with identical stages to form a binary-counter circuit.

Logic Circuits

Tunnel-diode switching is accomplished when a given "threshold" or turning point is exceeded, either the peak point or the valley point of the

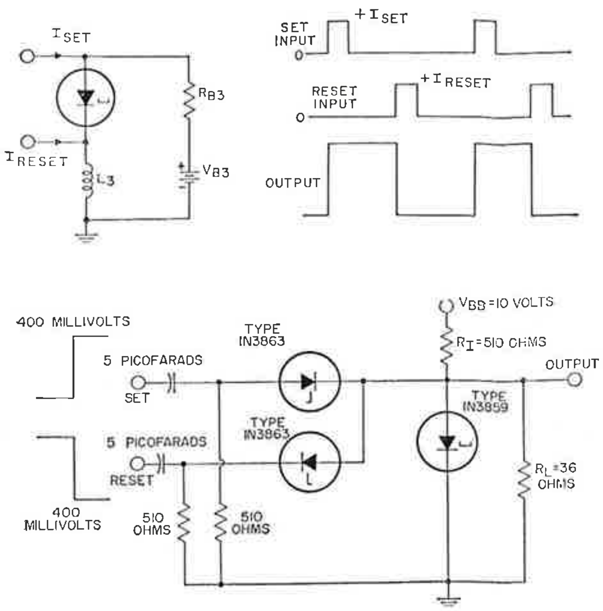


Fig. 36—(a) Set-reset flip-flop circuit and waveforms, and (b) practical set-reset flip-flop circuit having positive triggering.

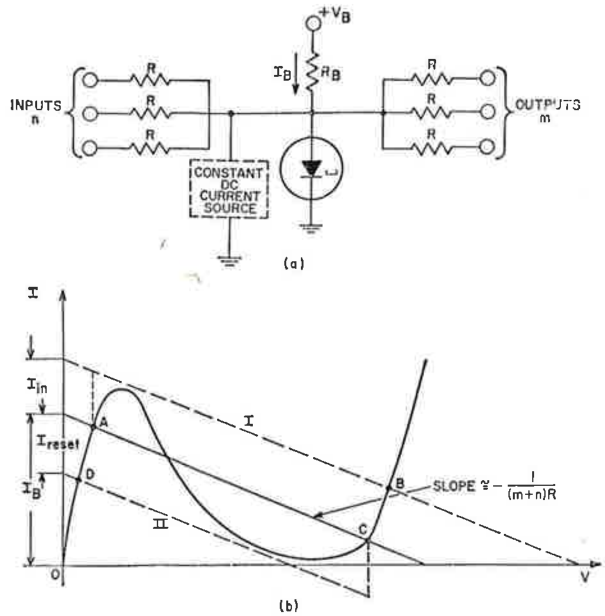


Fig. 38—(a) Tunnel diode bistable OR gate with reset, and (b) operating characteristics.

device. Thus, the word threshold is associated with tunnel-diode logic circuits, or **gates.**

Fig. 38a shows a tunnel-diode bistable **OR gate** with reset; the operation of such a circuit is shown in Fig. 38b. Normal biasing for the tunnel diode is at point A. When any one of the inputs supplies a current I_{in}, the tunnel diode switches to point B. Upon relaxation of the input, the operating point moves to C, a stable point. As a result, the tunnel diode must be reset back to point A before any more logic can be performed. This resetting is accomplished by means of the reset pulse source shown by the dashed load line II.

A bistable **AND gate** is designed in a manner similar to that described for the OR gate, except that all inputs must supply current before the tunnel diode switches to the high state.

The logic circuit shown in Fig. 38a has two disadvantages. First, because of the relative current gain of the tunnel diode, worst-case produces a circuit having impractical f... fan-out when realistic tolerances are ... **in** is the maximum number of inpu... gate; **fan-out** is the maximum numb... that can be driven from a logic gate.) because the tunnel-diode input and ... minals are the same, the signal may pro... either direction; as a result, phased power ... for V_B are required for unidirectional flo... information. This requirement makes the sy... much more complex and costly.

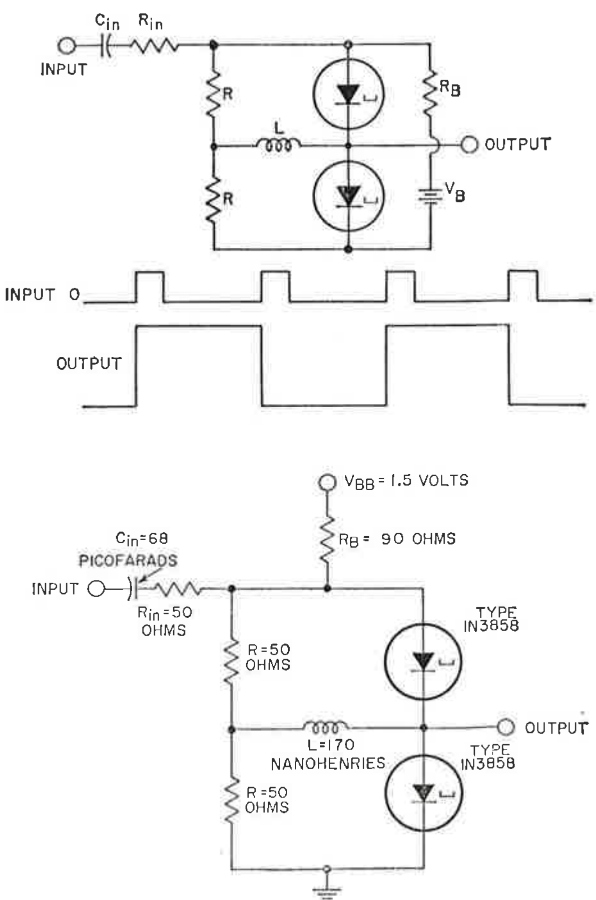


Fig. 37—(a) Triggerable flip-flop circuit and waveforms, and (b) practical triggerable flip-flop circuit.

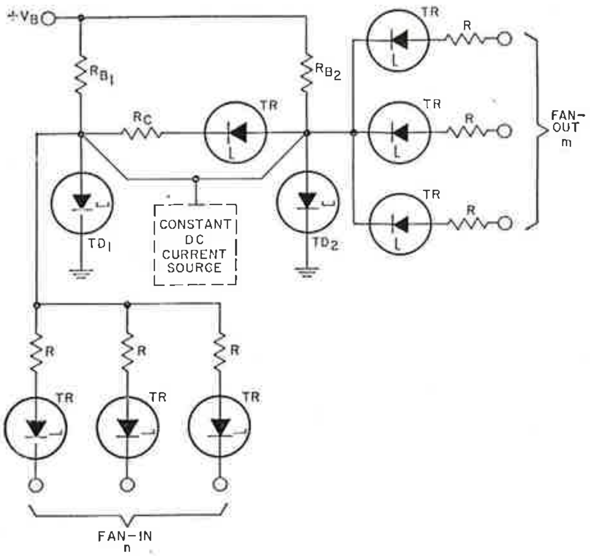


Fig. 39—Cascaded amplifier bistable OR and AND gate using tunnel rectifiers.

Both these disadvantages are overcome by the circuit shown in Fig. 39. This circuit uses a cascaded-amplifier arrangement which produces considerably larger current gain per stage; consequently, reasonable fan-in and fan-out are possible when practical tolerances are used. In this circuit, tunnel diode TD_2 may have a peak current from two to five times that of TD_1, depending upon the tolerances. This higher peak current accounts for the larger amplification.

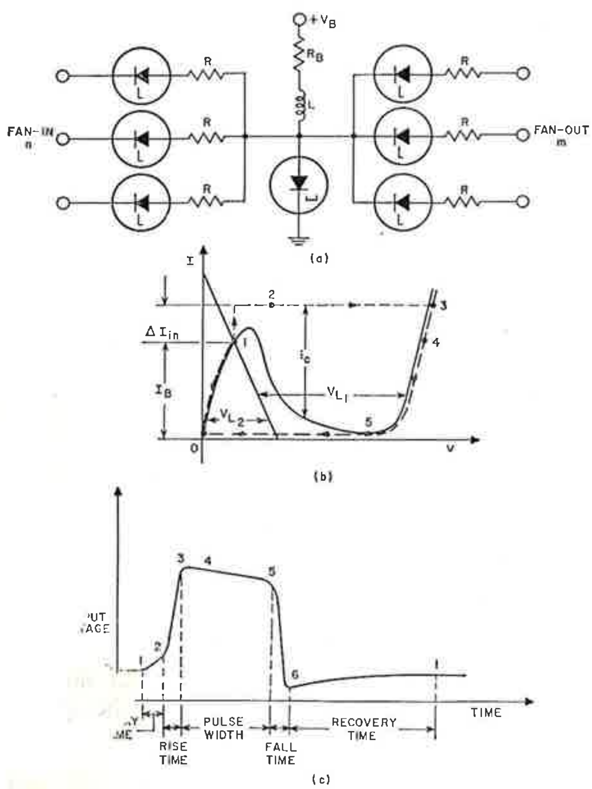


Fig. 40—(a) Monostable logic circuit, (b) operating characteristic, and (c) output waveform.

In addition, the circuit of Fig. 39 does not require phased power supplies because the tunnel rectifiers used in the inputs conduct current in only one direction. Although the reset pulse is still required with this circuit, a low-current pulse can generally be used. This pulse may be considered a **clock source** for synchronous logic circuits. The use of tunnel rectifiers also provides better gain by unloading inputs. In cases where extra-large fan-in and fan-out are needed, cascading may be extended to include three tunnel diodes in tandem. If cascading is extended beyond three diodes, the circuit becomes very costly and, at the same time, the stage propagation delay becomes proportionately longer.

A tunnel-diode logic circuit may be made self-resetting if it is monostable rather than bistable, as shown in Fig. 40a. This circuit also offers faster switching speeds than bistable gates. This advantage is demonstrated in Fig. 40b, which shows that the switching trajectory (dashed line) is constant current during transition from the low to the high state, or vice versa.

As indicated by the output waveform in Fig. 40c, both the pulse width and recovery time are determined by the size of inductance L. The rate of change of current through L is given by

$$\frac{diL}{dt} = \frac{V_L}{L} \qquad (25)$$

As a result, the pulse width is inversely proportional to V_{L1}, and the recovery time is inversely proportional to V_{L2}, as defined in Fig. 40b.

The pulse width and recovery time determine the maximum repetition rate of the monostable circuit. The minimum pulse width is determined by the driving requirements on the inputs to the gate. Therefore, maximum speed is obtained by making the recovery time as small as possible or by making V_{L2} as large as possible. This latter requirement can be readily achieved by the use of **nonlinear** biasing.

Nonlinear biasing or load lines are obtained by the use of a **tunnel rectifier,** as shown in Fig. 41a. In this circuit, the tunnel rectifier in series with the voltage V_{CL} results in the composite static characteristic shown in Fig. 41b. This nonlinear characteristic may be plotted as a load line on the tunnel-diode characteristic in the same manner as a linear load line,[7, 8] as shown in Fig. 41c.

Fig. 42 compares nonlinear and linear biasing, and shows the major advantages of the former. First, nonlinear biasing results in a considerably larger V_{L2} and, consequently, a much faster recovery time than linear biasing. Secondly, nonlinear biasing significantly improves current-gain

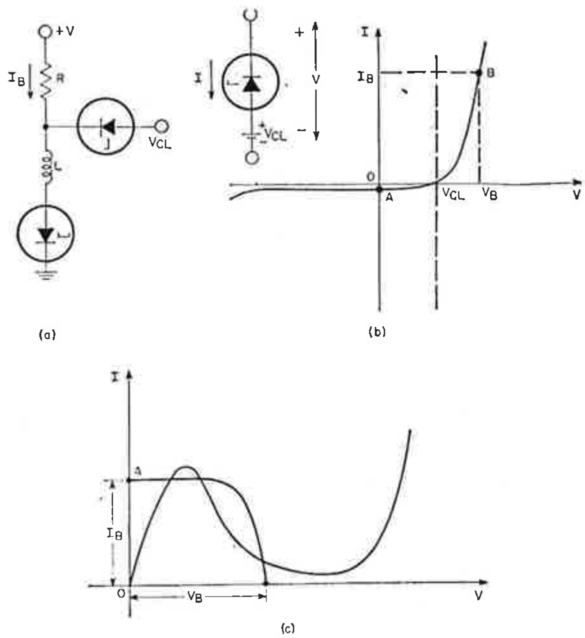


Fig. 41—(a) Tunnel diode and tunnel rectifier circuit, (b) composite characteristic of tunnel rectifier and voltage V_{CL} in series, and (c) nonlinear load line on tunnel diode characteristic.

capabilities, as shown by the much smaller I_{in} required to switch the tunnel diode over the peak at the same percentage of overdrive.

A monostable OR-gate circuit and an AND-gate circuit employing nonlinear biasing are shown in Figs. 43 and 44, respectively. These circuits feature tunnel-rectifier coupling for directionality, as well as cascaded monostable amplifiers for sufficient high-speed gain to obtain reasonable fan-in and fan-out.

Fig. 45 shows a compatible level-producing bistable circuit which can be used with monostable OR and AND circuits. This bistable circuit is used to obtain storage registers, counters, and inversion (through the reset input), as well as to eliminate timing problems in AND gates by gating levels against pulses wherever necessary.

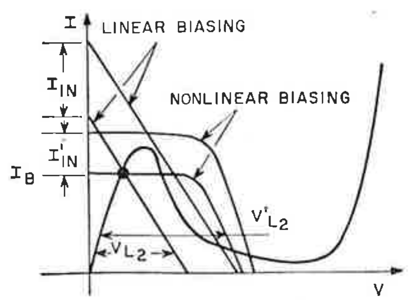


Fig. 42—Comparison of linear and nonlinear biasing.

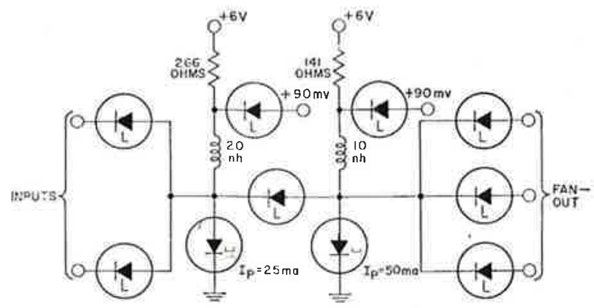


Fig. 43—Complete monostable OR gate using germanium tunnel diodes and rectifiers.

The three circuits shown in Figs. 43, 44 and 45 form a complete set of logic blocks (having one-nanosecond logic delays and 200-megacycles per second repetition rates) for use in a large-scale computer.[9-11] (These circuits as well as that shown in Fig. 49 use tunnel diodes which are presently in the development stages, May, 1963.) These circuits have all been designed for worst-case operation. Logical interconnection of these circuits is accomplished by use of coaxial cable which prevents cross-talk, effect of common-ground paths, and other transient noise problems at high speeds (200 megacycles per second and up).

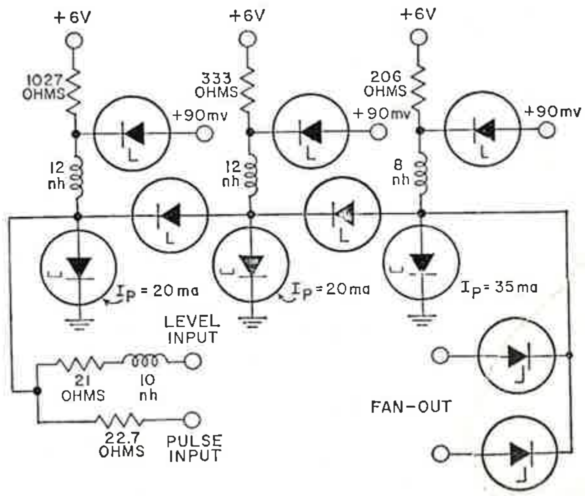


Fig. 44—Complete monostable AND gate using germanium tunnel diodes and rectifiers.

These three logic circuits may be used to build an ultra-high-speed shift register, as show[n in] Fig. 46. The shift register uses two AND ga[tes] and two bistable units in each stage. A count[er] may also be constructed by use of a ... form of the circuit shown in Fig. 46. T... circuit is connected in a ring so that the A_i ... connected to the $\overline{A}_0$ input; the counter the... a scale of 2i where i is the number of stages in the ring. This counter functions properly when all stages are initially set to zero.

When low-peak-current tunnel diodes and tunnel rectifiers are used in the circuits of Figs. 43, 44 and 45, a micro-power computer can be designed which requires several hundred times less power than low-power transistor circuits at no sacrifice in operating speed.

Computer Memories

Because of its voltage-controlled negative-resistance characteristic, the tunnel diode is an ideal element for computer memories. Its ultra-high-speed capabilities make the tunnel-diode memory many times faster than other presently used memory systems.

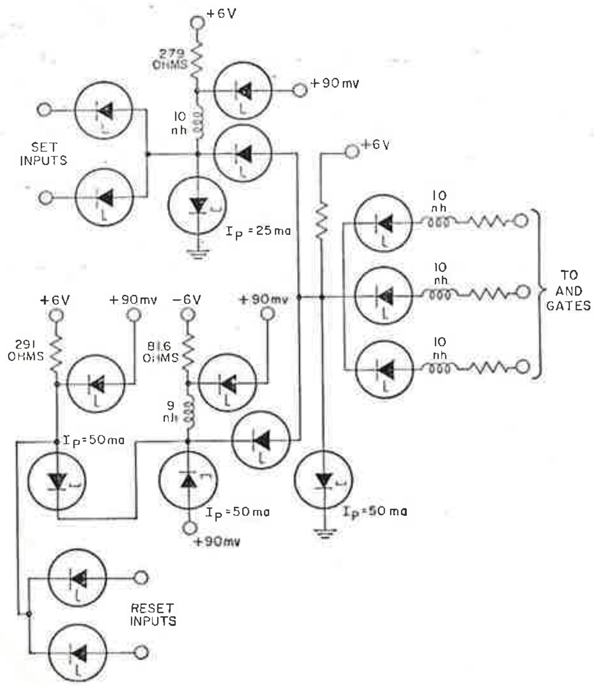


Fig. 45—High-speed bistable circuit having two set and reset inputs.

Fig. 47 shows a 25-nanosecond-cycle-time memory cell which uses a germanium tunnel diode ·'· a gallium arsenide tunnel rectifier.[11, 12] The ·ory cell is read by applying a pulse to the ·ne which causes a sensed output to appear ·digit line. The write operation is achieved ·ying pulses having the indicated polarities the word line and the digit line.

· basic operation of the memory cell is ·rated in Fig. 48. The voltage supply V_B ·tor R_B produce a current which allows ·ium tunnel diode to be either in a low-voltage state (storing a "1") or high-voltage state (storing a "0"), as shown by points A and B, respectively, in Fig. 48a. Reading from the cell is attained by applying a negative word-line voltage to the tunnel diode, as shown in Fig. 48b. If the tunnel diode is in the low state, it is

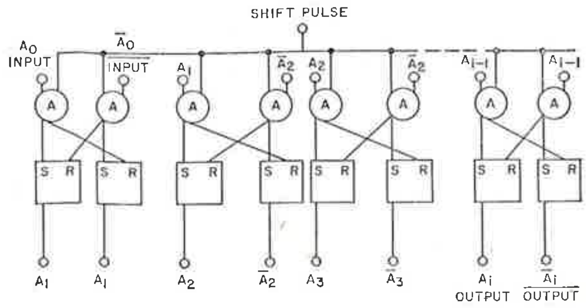


Fig. 46—Block diagram of shift register.

switched to the high state. Because the current which switches the tunnel diode must pass through the tunnel rectifier, a negative output pulse is obtained on the digit line only when the tunnel diode is caused to switch. Writing into the

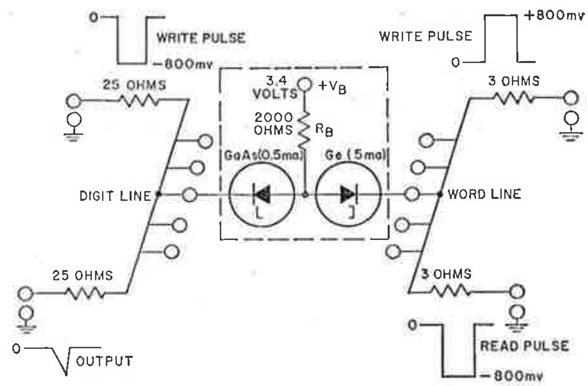


Fig. 47—Basic memory cell.

memory is achieved when a negative voltage appears on the digit line in coincidence with a positive voltage on the word line, as shown in Fig. 48c. This condition switches the tunnel diode to the low state, and a "1" is set into the cell.

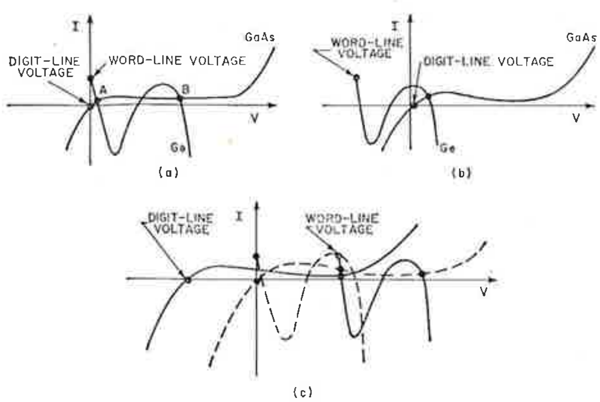


Fig. 48—Operation of memory cell.

The peripheral circuits (selection, sensing, and driver circuits[11, 12]) are shown in Fig. 49. The word switch having X and Y inputs selects the word addressed. The word-switch driver then places a negative-voltage pulse on the word line, which reads the information from the memory cell. When a "1" is stored in the cell, an output is obtained on the digit line which is in turn amplified by the sense amplifier. The sense amplifier feeds the output back to the digit driver, which then places a negative-voltage pulse on the digit line. Because the word-switch driver is

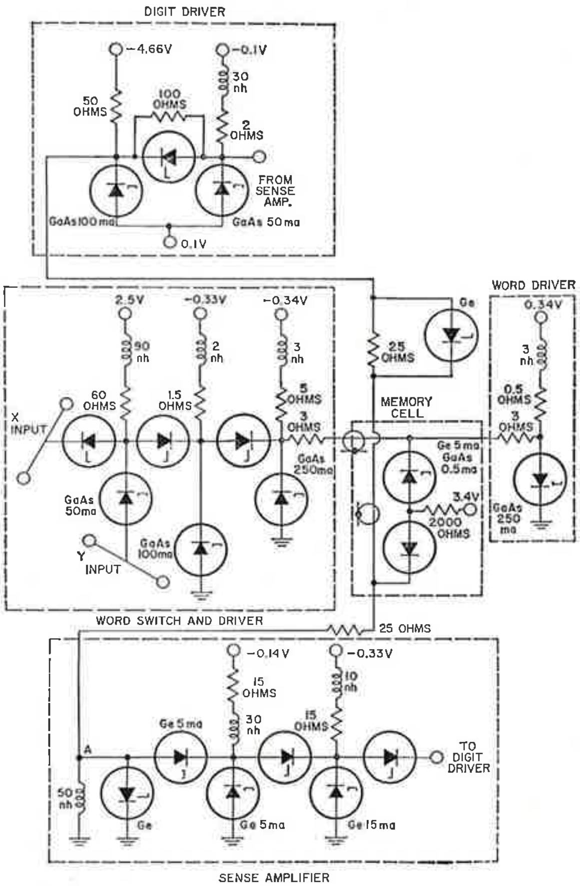


Fig. 49—Peripheral memory circuits.

monostable, its negative pulse terminates; this action in turn causes the word driver to be switched positively. The positive-pulse output of the word driver to the word line and the negative-pulse output of the digit driver to the digit line are made coincident; consequently, the tunnel diode in the memory cell is reset to the low state and the information is regenerated. Obviously, if the memory cell originally contained a zero, there would be no input to the sense amplifier and no output from the digit driver. The memory cell would then not be reset, but would remain in the "0" state.

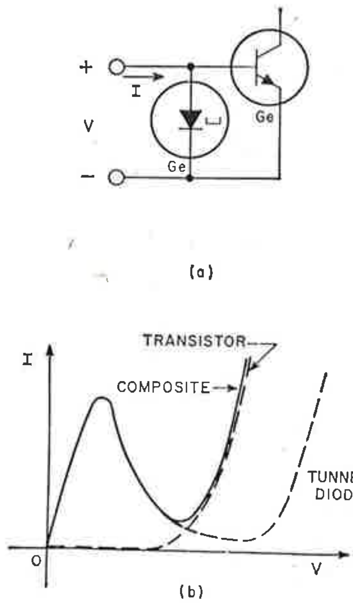


Fig. 50—(a) Basic tunnel diode and transistor combination, and (b) composite characteristic.

Hybrid Circuits

The combination of transistors and tunnel diodes in hybrid circuits offers many advantages. For example, the combination of these two devices in switching circuits offers the possibility of speeds many times faster than those of circuits using transistors only. Accordingly, the speed of present-day computer circuits (particularly those in the arithmetic unit) could be generally improved through the use of hybrid circuits. In addition, hybrid circuits eliminate many of the difficulties of circuits using tunnel diodes only, such as lack of isolation between input and output and the tight tolerances required to achieve reasonable circuit gains. In general, tunnel-diode and transistor hybrid circuits provide a good compromise between high speed and circuit complexity.

Basic Combinations

In the basic tunnel-diode and transistor combination, the tunnel diode is placed in parallel with the base-emitter terminals of the transistor, as shown in Fig. 50a; the resulting composite characteristic is shown in Fig. 50b. This combination can be operated in either the common-emitter or common-base configuration. In **common-emitter** operation, the tunnel diode is used primarily to speed up the input to the transistor. it may also provide some current gain in addition to that obtained from the transistor. In **common-base** operation,[13] the tunnel diode provides all the current gain, as well as high speed. In both modes, the transistor provides both isolation between input and output and voltage gain.

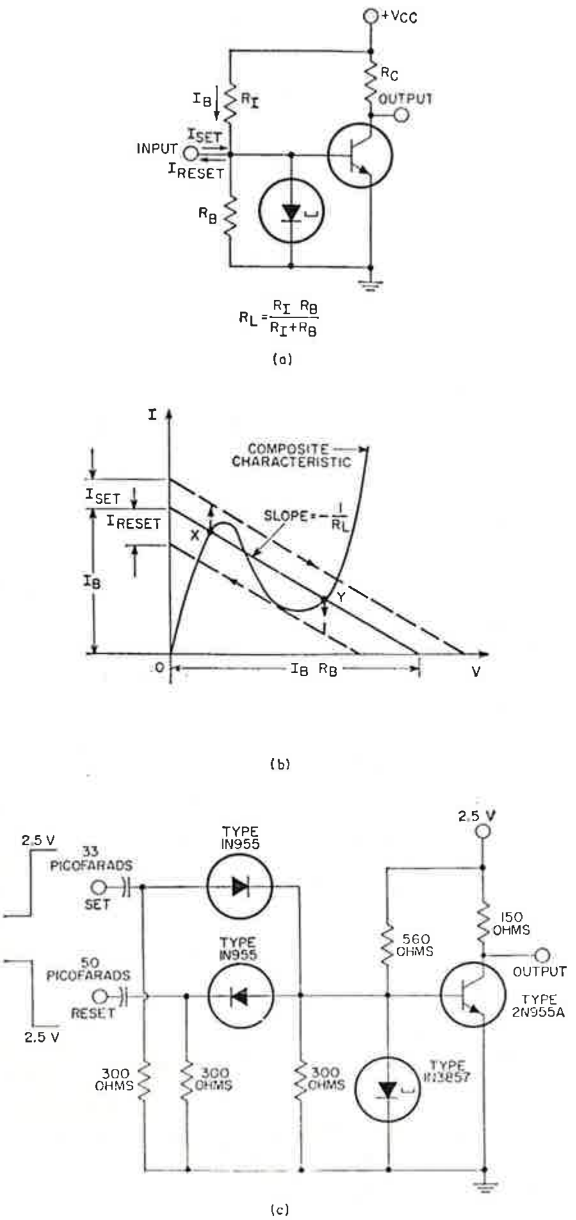


Fig. 51—(a) Common-emitter bistable circuit, (b) characteristic, and (c) practical circuit.

Although Fig. 50a shows the combination of a germanium tunnel diode with a germanium transistor, the use of gallium arsenide tunnel diodes with germanium transistors or the combination of gallium arsenide or germanium tunnel diodes with silicon transistors is also possible. The use of gallium arsenide tunnel diodes with germanium transistors has been found to be the most practical combination because it requires the least control of characteristic-voltage tolerances. The use of silicon transistors requires an offset bias on the emitter even in combination with gallium arsenide tunnel diodes.

Common-Emitter Circuits

The combination of tunnel diodes and transistors in a common-emitter configuration results in a natural bistable circuit, as shown in Fig. 51a. The composite operating characteristic in Fig. 51b shows the two stable points X and Y, and the set and reset currents necessary to accomplish these functions. For proper operation, the current flowing into the base at point B must be sufficient to saturate the transistor. The switching speed of this bistable unit is much faster than that obtained with a transistor alone. Storage and fall time are also fast because the tunnel diode

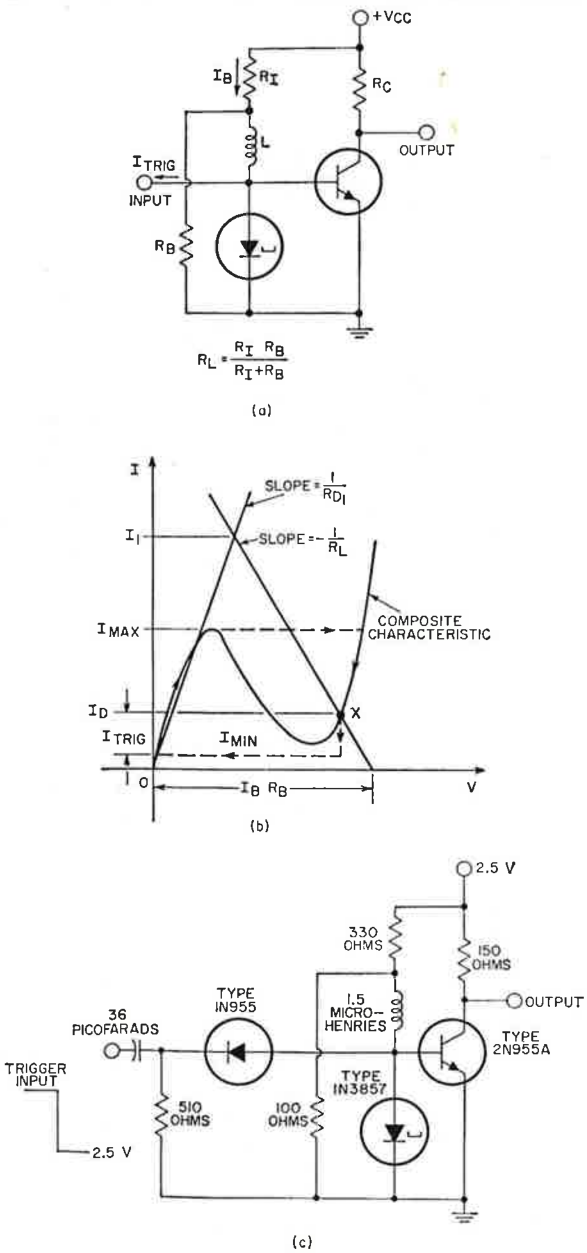


Fig. 52—(a) Common-emitter monostable circuit, (b) characteristic, and (c) practical circuit.

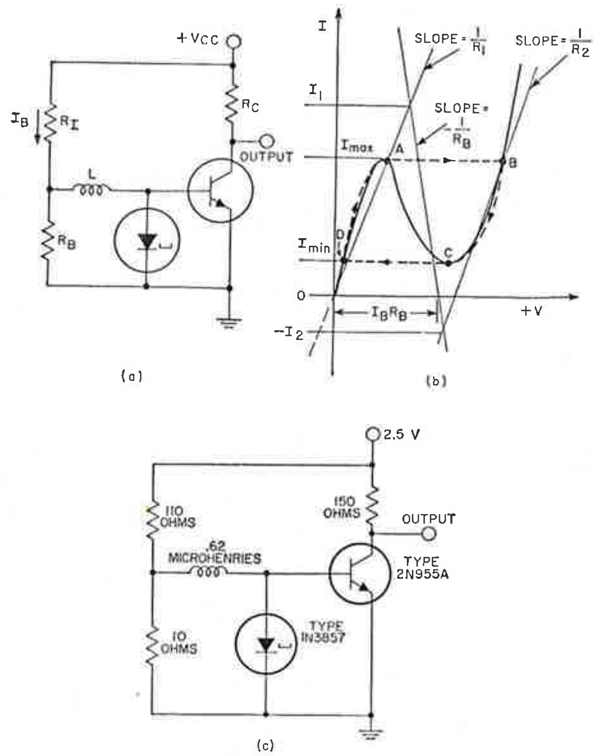


Fig. 53—(a) Common-emitter astable circuit, (b) operating characteristic, and (c) practical common-emitter astable circuit.

presents such a low impedance during turn-off. In addition, these high speeds are achieved without the use of any reactive or speed-up elements because the tunnel diode provides essentially a perfect step input to the transistor. A practical example of a common-emitter bistable circuit is shown in Fig. 51c. This circuit has a maximum repetition rate of 50 megacycles per second.

Monostable operation of the common-emitter combination is also easily obtained, as shown in 52a. The operation of the circuit is illustrated in Fig. 52b, which shows that the stable point on the composite characteristic is established when the transistor is "on" and saturated. A trigger pulse of current momentarily switches the tunnel diode to the low state, and turns the transistor off. The inductance L then allows the current in the tunnel diode to increase towards I_1. When this current, plus the current I_B, exceeds the peak current of the tunnel diode ($I_{max} = I_P$), it switches to the high state, the circuit returns to stable point, and the transistor is turned on and saturated. The pulse width or delay time is the off-time of the transistor and is given by

$$t_{off} = \frac{L}{R_L + R_{D_1}} \ln \frac{I_1 - (I_D - I_{Trig})}{I_1 - I_{max}} \qquad (26)$$

where

$$R_L = \frac{R_I R_B}{R_I + R_B}$$

and I_1, I_D, I_{Trig}, and I_{max} are defined in Fig. 52b.

Fig. 52c shows a practical example of a common-emitter hybrid monostable circuit. This circuit has a maximum repetition rate of 25 megacycles per second and a delay time of 20 nanoseconds.

An astable multivibrator circuit is shown in Fig. 53a; the operation of the circuit is described in Fig. 53b. This circuit is astable because the dc load line formed by R_L intersects the composite characteristic in the negative-resistance region. The inductance L causes the circuit to follow the switching trajectory ABCD. As a result, the transistor switches on and off at a rate determined by L and the slope of the composite characteristic in both the on and off regions.

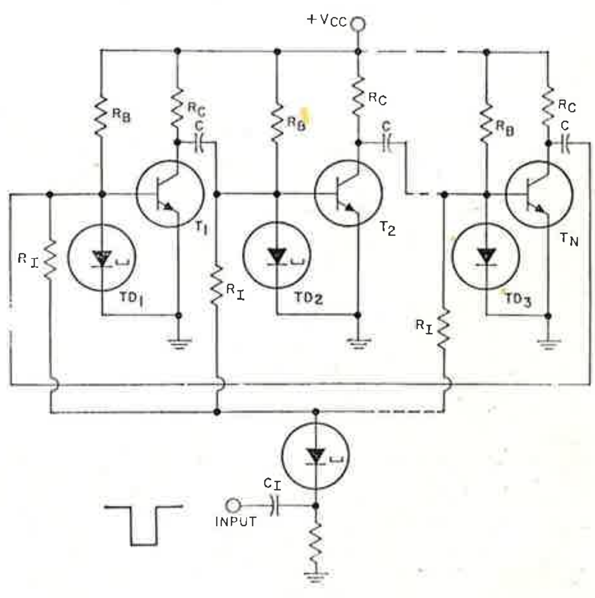


Fig. 54—Hybrid ring counter.

When these slopes are approximated by R_2 and R_1, as shown in Fig. 53b, the on and off times are given by

$$t_{on} \cong \frac{L}{R_L + R_2} \ln \frac{(I_2 + I_{max})}{(I_2 + I_{min})}$$
$$t_{off} \cong \frac{L}{R_L + R_1} \ln \frac{(I_1 - I_{min})}{(I_1 - I_{max})} \qquad (27)$$

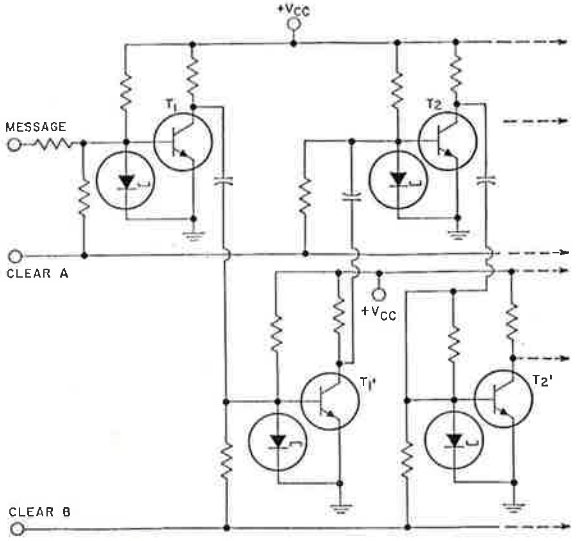


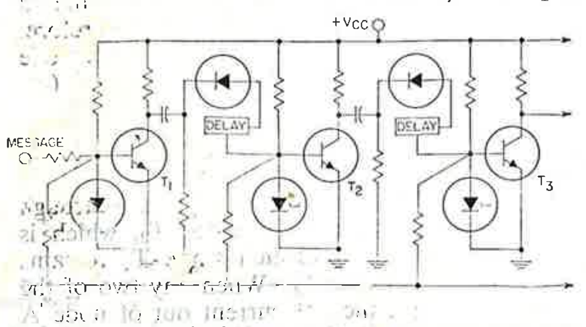
Fig. 55—Hybrid shift register using two bistable units.

where I_1, I_2, I_{max}, and I_{min} are as defined in Fig. 53b.

Fig. 53c shows a practical example of an astable multivibrator circuit. The frequency of oscillation for this circuit is 20 megacycles.

The bistable, monostable, and astable circuits shown have two major advantages over circuits using only transistors. First, they offer speed and repetition rates considerably faster than those possible without the use of the tunnel diode. Second, they are much simpler in construction and require fewer components.

The common-emitter bistable circuit is a storage element and, therefore, is useful in counters and shift registers. A very simple ring counter[14] using the hybrid bistable circuit is shown in Fig. 54. In this ring-counter circuit, transistors T_1 through T_N are normally cut off; as a result T_1 must be set "on" to start the circuit counting. Thus, T_1 is in the "1" or "on" state and all other transistors are in the "0" state. When a negative-going pulse appears at the input, all the tunnel diodes tend to be switched to the low state, and T_1 is



—Hybrid shift register using one bistable unit.

turned off. As the collector of T_1 rises toward $+V_{CC}$, current flows through R_C and C into the base circuit of T_2. This instantaneous current is greater than the sum of the input current and the peak current of TD_2. Therefore, TD_2 is switched to the high state, and T_2 is turned on to the "1" state. Because the current through C decays exponentially as it charges toward V_{CC}, the input current must be terminated before the capacitor current becomes small enough to return TD_2 to its low state. Thus, the following two conditions on input-pulse length must be satisfied for effective operation: (a) input pulse length greater than transistor turn-off time, and (b) input pulse length greater than turn-off time plus current decay time.

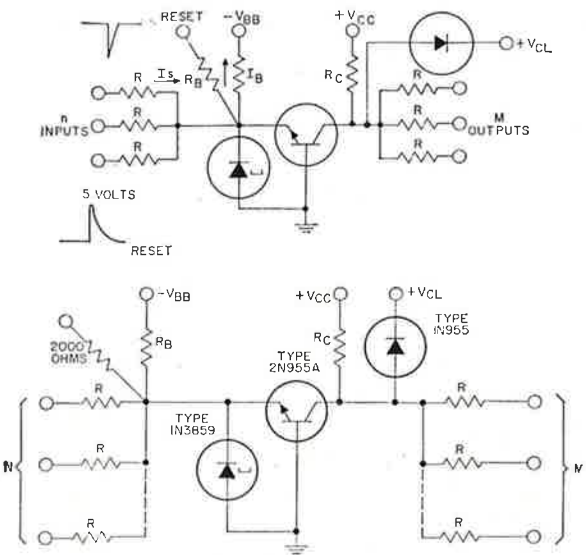


Fig. 57—(a) Basic common-base logic gate, and (b) practical common-base logic gate having propagation delay of 5 ns per stage. OR gate: $R = 1.1K$ ohms, $R_B = 465$ ohms, $R_C = 865$ ohms, $V_{CC} = V_{BB} = 15$ volts, $V_{CL} = 5$ volts, max. $m = 2$, $n = 3$. AND GATE: $R = 1050$ ohms, $R_B = 680$ ohms, $R_C = 890$ ohms, $V_{CC} = V_{BB} = 15$ volts, $V_{CL} = 5$ volts, max. $m = 2$, $n = 6$.

These conditions are achieved by proper selection of C_1 to provide the necessary differentiated waveform from the input.

A shift register using the common-emitter bistable circuit is shown in Fig. 55. This shift register uses two of the basic bistable units to form one stage. The message is read into T_1 and then shifted by a channel-A clear pulse to T_1' for temporary storage so that T_1 can receive the next bit. The channel-B clear pulse follows the channel-A clear pulse so that the first bit is shifted into T_2 before the next clear pulse from A is generated.

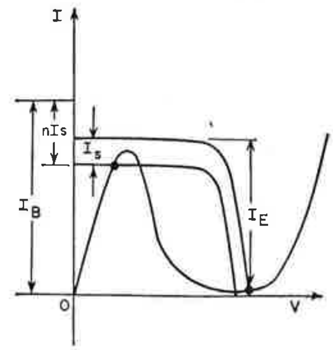


Fig. 58—OR-gate biasing.

Fig. 56 shows a shift register which uses only one basic bistable unit to form a stage. In this circuit, small time delays are necessary to prevent the signal from the previous stage from appearing at the same time as the shift pulse. This shift register therefore uses a dynamic type of temporary storage rather than the static temporary storage used in Fig. 55.

Common-Base Circuits

The common-base configuration is the fastest of the tunnel-diode and transistor combinations, because the tunnel diode provides all the current gain at very high speed. Consequently, common-base operation appears to be the best selection for high-speed logic circuits. However, common-emitter circuits offer much higher current-gain possibilities and, therefore, larger fan-in and fan-out for the same component tolerances. Consequently, if lower speeds are acceptable, common-emitter circuits are the best choice.

The basic common-base hybrid logic gate is shown in Fig. 57a; a practical example of this circuit is shown in Fig. 57b. The operation of this circuit as an OR or AND gate is dependent upon the bias current I_B and the number of inputs n, as shown in Figs. 58 and 59. In both cases, it is assumed that a current I_S flows into each of the inputs when the gate is not energized. The effect of a pulse at any one of the inputs is that

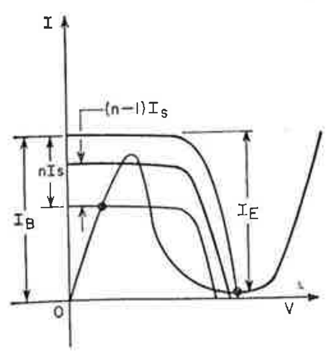


Fig. 59—AND-gate biasing.

the current out of the tunnel-diode node is increased by the amount I_S.

The output rise time of this logic circuit is primarily determined by the current step response of the transistor and the output capacitance from collector to ground. The total signal-propagation delay per stage in this type of circuit is the sum of the tunnel-diode delay, the transistor turn-on delay, and the transistor rise time. This type of circuit offers stage delays as low as a few nanoseconds when fast, low-capacitance transistors are used.

Fig. 60 shows the inversion circuit, or complementary gate, for common-base logic circuits. In this circuit, the state of the tunnel diode determines which of the transistors becomes energized. When the tunnel diode is in the low state, transistor T_1 is not conducting and T_2 is conducting. When the tunnel diode is switched to the high state by the proper input, T_1 is made to conduct, and T_2 is effectively cut off because it receives only the base current from T_1.

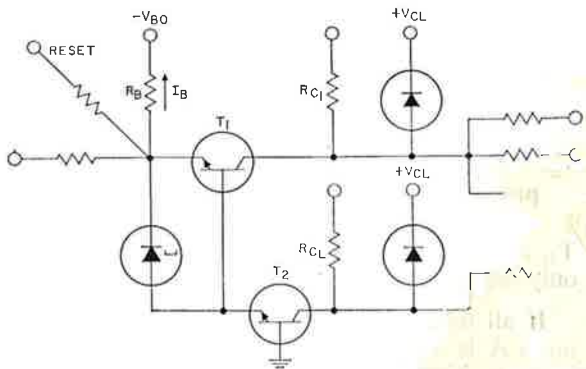


Fig. 60—Basic complementary gate.

The unique current-routing properties of the complementary-gate circuit permit the design of many simple circuits offering both improvements in speed and savings in components. Among these circuits is the binary adder shown in Fig. 6[illegible] The operation of this circuit assumes that "[illegible] input corresponds to the current I_S, and "1" drops the input current to zero. The[illegible] if all inputs are "0", the net current out [illegible] A is zero, and both transistors are cut [illegible] and S_N equal "0").

If any one of the inputs becomes negative, the net current out of node A is I_S. Because I_S is le[illegible] than the peak current I_p, it flows entirely thro[illegible] the tunnel diode and in transistor T[illegible] then turned on and satura[illegible] cut off ($C_N = 0$, S [illegible] inputs go to "1" [illegible] is 2 × I_S; b[illegible] then than I_p, th[illegible] d to [illegible] state, [illegible]ence, [illegible] I_S) [illegible]

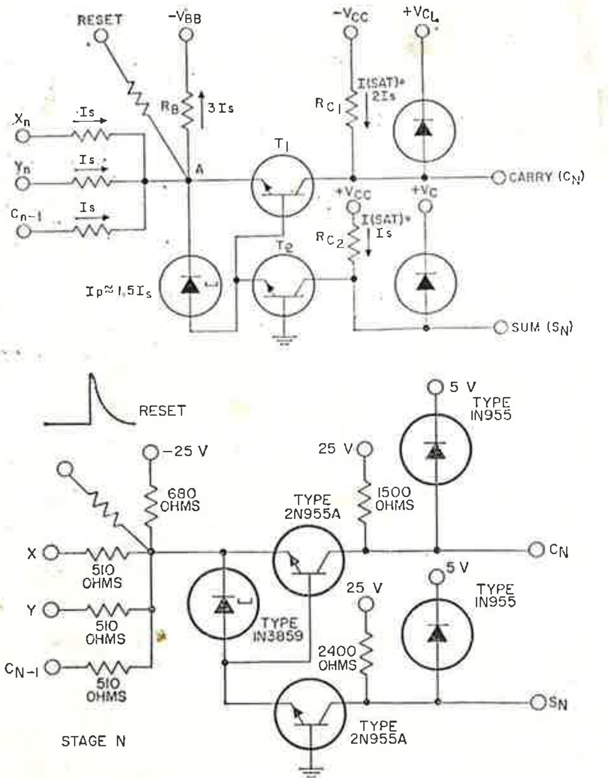


Fig. 61—(a) Hybrid full-adder circuit, and (b) practical example of binary adder circuit.

T_1, and T_2 is essentially cut off because it receives only the base current of T_1 ($C_N = 1$, $S_N = 0$).

If all three inputs are "1", the current out of node A is $3 \times I_S$; as a result, the tunnel diode is in the high state and transistor T_1 is conducting. However, the collector circuit is designed so that T_1 saturates at a current of $2 \times I_S$. This condition means that the net base-current flow of T_1 must be approximately I_S (assuming negligible current through the tunnel diode). Therefore, T_2 is receiving a current I_S and is also on and saturated (and the outputs $C_N = 1$, $S_N = 1$ are achieved).

Fig. 61b shows a practical example of the binary adder circuit. The typical propagation delay for this circuit is eight nanoseconds for "carry", and 16 nanoseconds for "sum". In this example, the C_n output can be used to drive the C_n input of the n + 1 state.

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Radiotronics is published twelve times a year by the Wireless Press for Amalgamated Wireless Valve Co. Pty. Ltd. The annual subscription rate in Australasia is £1, in the U.S.A. and other dollar countries $3.00, and in all other countries 25/-.

Subscribers should promptly notify Radiotronics, P.O. Box 63, Rydalmere, N.S.W., and also the local Post Office of any change of address, allowing one month for the change to become effective.

Printed by CLOISTER PRESS (W. Short), REDFERN, N.S.W.