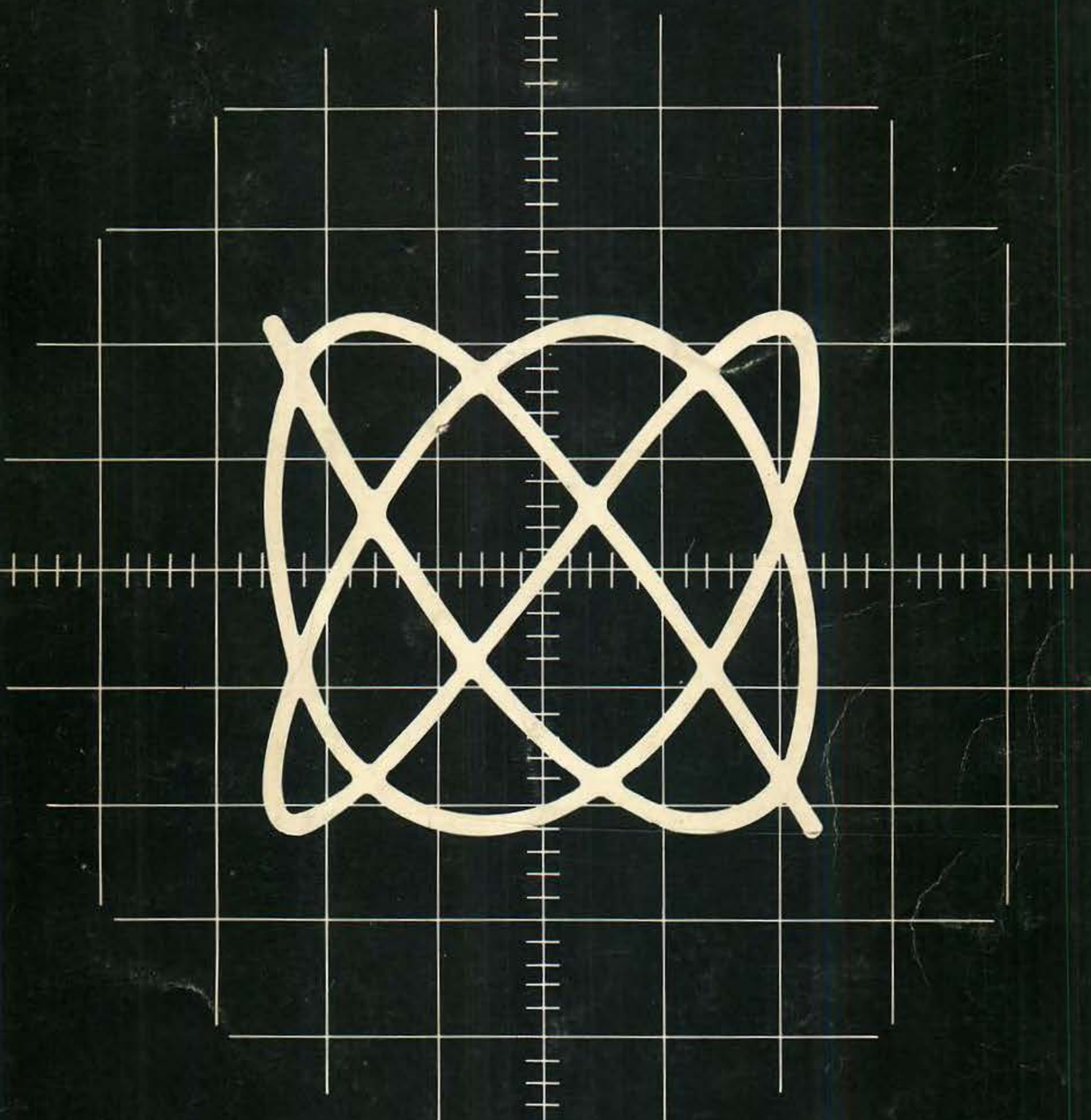

100 IDEAS FOR DESIGN

No. 3

ELECTRONIC DESIGN

Foreword

This is the third volume in the series, "100 Ideas for Design." As in the previous volumes, we have selected 100 of the best items that appeared during the past year in the "Ideas for Design" section of ELECTRONIC DESIGN. These have been grouped by category for easy reference.

The continued popularity of the "Ideas for Design" section is very gratifying, although not surprising. After all, the electronic design engineer is an originator as well as a user of ideas. The almost 1,000 ideas we have published since we started the section some eight years ago have been useful and -- we feel -- have given rise to many new ideas as well.

The volume starts with the $1,000 award-winning "Idea of the Year. Details of the award competition appear on the inside back cover. Why not submit your idea? You might be the winner next year.

Incidentally, Volumes I and II are in their third printing and are available as outlined on the inside back cover.

Edward E. Grazda
Editor

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IDEA OF THE YEAR

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*Indicates winner of "Most Valuable of Issue" Award

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MEASUREMENT & TEST CIRCUITS

COMMUNICATIONS & TELEMETRY

*Indicates winner of "Most Valuable of Issue" Award.

SWITCHING & RELAY CIRCUITS

POWER SOURCES

CIRCUIT TRICKS

*Indicates winner of "Most Valuable of Issue" Award.

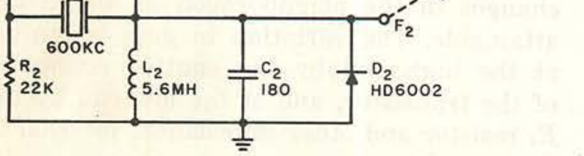
Idea of the Year

ELECTRONIC DESIGN'S 1962 Idea of the Year Award of $1,000 was presented to John M. Tewksbury, principal engineer, Bendix Radio Div., The Bendix Corp., Avionics Products, Baltimore, Md.

His idea, reproduced here, was selected by the editors of ELECTRONIC DESIGN as the most useful and interesting of those published in the Ideas for Design section of the magazine during the year.

Bilateral Crystal Oscillator Has Two-Frequency Output

The bilateral characteristics of the transistor come into play in this two-frequency crystal oscillator. Either of the two frequencies may be selected by applying plus-or-minus voltage to the supply terminal.



Either of two oscillator frequencies can be obtained by reversing polarity of supply voltage.

When plus voltage is applied, the current flows through D_1 to the emitter of the transistor. Tuned circuit L_2C_2 and crystal unit X_2 then become active and an output is derived from F_2. The L_1C_1 network is shorted out by diode D_1 and crystal X_1 is connected between base and emitter where there is no gain to promote oscillation in this unit.

With minus voltage applied to the supply terminal, the transistor operates in the "inverted" mode. Effectively, the collector becomes the emitter and vice versa. In this case, oscillations are generated by L_1C_1 and crystal unit X_1. Output voltage will be delivered at output terminal F_1.

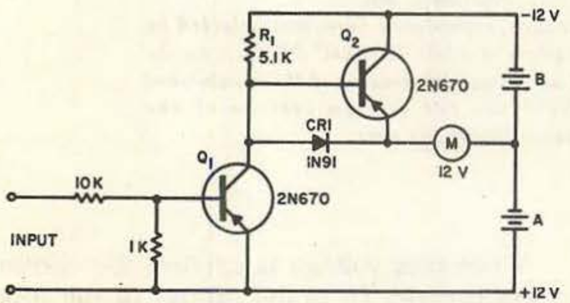
If desired, two additional capacitors (shown by dashed lines) may be added to provide a single output terminal.

Diodes D_1 and D_2 limit the output to 0.7-v rms. Since the transistor is operating in an oscillator circuit the beta of the inverted mode need only be sufficient to produce oscillation. Therefore, it is not necessary to use selected bilateral transitors in this application. This circuit can be used to select frequencies from a remote point by electrical means. Since it uses a single transistor, there is a saving in components.

DC-Servo Amplifier Has Single-Ended Drive

In driving a small dc motor in a servo loop it is sometimes desirable to have a single-ended input. Here is an amplifier that affords high voltage gain and high input impedance.

When the input voltage (or current) is sufficient to cause Q_1 to conduct, it first cuts off Q_2 and then conducts through CR_1. The motor thus is connected to supply A, causing rotation in one direction. When the input is reduced, R_1 will bias Q_2 on and the current through the motor is then in the opposite direction, through supply B. Voltage amplification is provided by Q_1, and Q_2 is an emitter follower that makes possible equal drive torque in each direction. This circuit has been sucessfully used to provide follow-up in a low-voltage power supply to reduce series regulator dissipation.



Input-signal level biases Q_1 or Q_2 on to connect required battery polarity to motor.

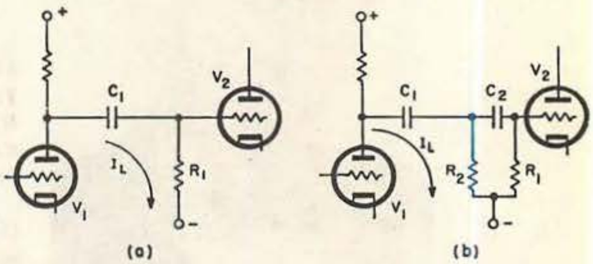
Richard L. Shaum, staff associate, Sandia Corp., Albuquerque, N. M.

Double Coupling Capacitors Avoid Leakage Current

Here's an amplifier coupling scheme that can be quite valuable, particularly when the coupling is over several thousand volts as in cathode ray tubes.

In the commonly used arrangment, Fig. 1a, the leakage current of capacitor C_1 flows through resistor R_1. This can cause an appreciable shift in the bias on the following stage V_2.

But, when C_2 and R_2 are added, as in Fig. 1b, the leakage current of C_1 is largely blocked by C_2. Instead of flowing through R_1, the leakage flows through R_2 and the bias on V_2 is not shifted.



Extra coupling capacitor C_2 in (b) blocks leakage current of C_1 from affecting bias of following stage V_2.

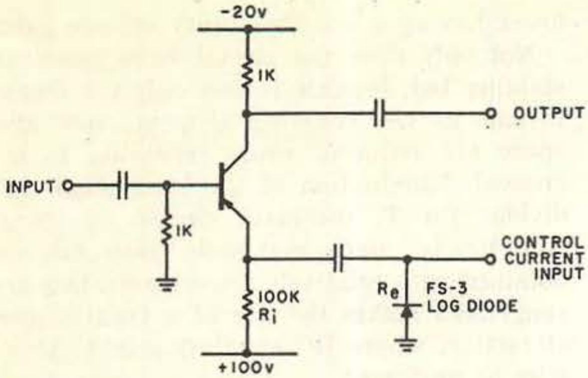
E. R. Roeschlein, electronic engineer, U. S. Naval Avionics Facility, Indianapolis, Ind.

Variable-Gain Amplifier Has 40-Db Range

Gain of the circuit to be described is controlled by the current flowing through a log diode. The diode is the dynamic emitter resistance (R_e) and its dynamic impedance determines the circuit gain. The dynamic resistance of the diode is a linear function of current (the higher the current, the lower the resistance). Thus, the gain of the stage is directly proportional to the diode control current. The R_i resistor derives the operating current for the transistor.

With such a circuit configuration, gain changes in the neighborhood of 40 db are attainable. The variation in gain is limited at the high end by the emitter resistance of the transistor, and at the low end by the R_i resistor and other impedances internal to the transistor.

Such a circuit is limited to small signal operation due to the nonlinear diode characteristic.



Current through log diode controls gain of amplifier stage.

Log diodes can also be placed across the collector resistor and base resistor to affect gain. It is anticipated that the gain of a single video couplet can be varied more than 70 db with triple diode control.

Robert W. Cope, project engineer, Bendix Radio Div., The Bendix Ccorp., Baltimore 4, Md.

Wien-Bridge Amplifier Has Selectivity With Stable Gain

The conventional frequency-selective feedback amplifier having a rejection filter in the feedback loop has the disadvantage that the gain of the amplifier at the center frequency is equal to the forward-path supply voltage, and tube parameters.

The circuit shown in Fig. 1 is an adaptation of the Wien-bridge frequency-rejection network. It is driven by a split-load bridge driver in series with an amplifier. The amplifier is connected between the cathode and grid of the bridge driver. The flow diagram satisfying Fig. 1 is shown in Fig. 2. From this diagram amplifier gain at center frequency can be obtained by $R_2/(R_1 + R_2)$. Gain at frequencies far from the center frequency will be:

$$\frac{R_2}{R_1 + R_2} \; \frac{1}{1 - A\dfrac{R_1}{R_1 + R_2}} \quad \text{or}$$

$$\frac{R_2}{R_1(1 - A) + R_2}$$

where A is the gain of V_2. Simultaneous change of C or R will provide a change of center frequency by $1/2\pi\, RC$.

A simple, inexpensive Wien-bridge frequency-selective amplifier using a triode-pentode tube is shown in Fig. 3. To avoid unbalance of the bridge, a variable resistor is used as R_k. For the adjustment, the feedback loop between the plate of V_2 and the grid of V_1 is opened at a convenient point. The R or C and the pot R_k should be adjusted to the null by applying a desired frequency signal at the input. A CRO or VTVM at the plate of V_2 is used as a detector.

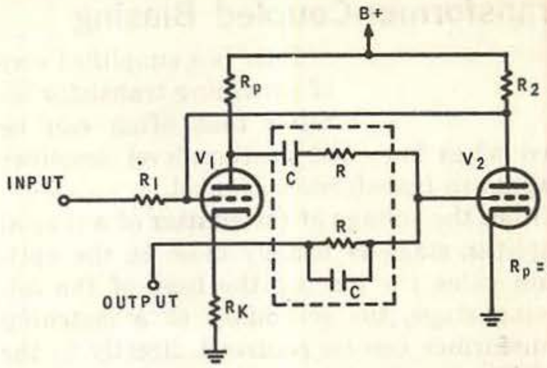


Fig. 1. Basic Wien-bridge frequency-selective amplifier.

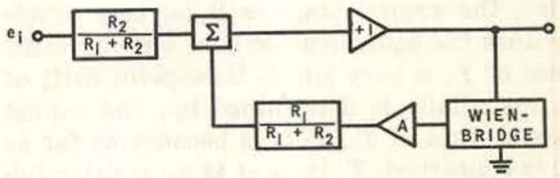


Fig. 2. Gain at center frequency can be obtained from flow diagram.

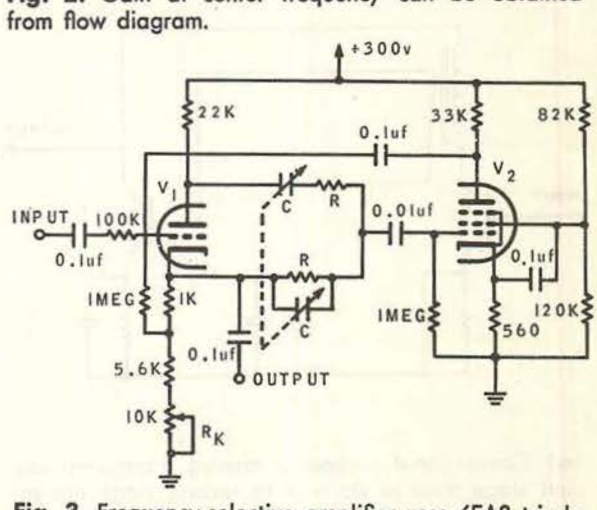


Fig. 3. Frequency-selective amplifier uses 6EA8 triode-pentode. Cathode potentiometer is used for balancing.

The test result indicated that the gain stability of the amplifier at the center frequency is very good despite large changes in the power supply voltage as well as change of tubes. The frequency-selective amplifier described can be operated up to about 500 kc.

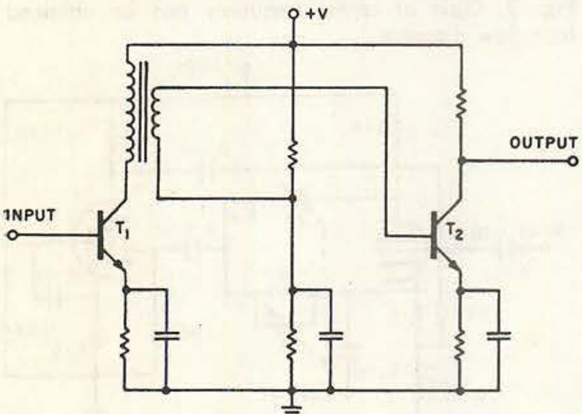
K. H. Liu, engineer, Industrial Nucleonics Corp., Columbus 2, Ohio.

AMPLIFIERS

Short-Cut Connection Simplifies Transformer-Coupled Biasing

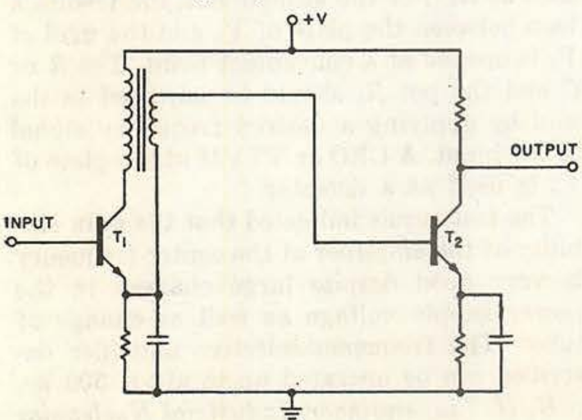
Here is a simplified way of providing transistor biasing that often can be used when low- and medium-level amplifier stages are transformer-coupled.

Since the voltage at the emitter of a typical amplifier stage is usually close to the optimum value for biasing the base of the following stage, the secondary of a matching transformer can be returned directly to the emitter, (b). As long as the emitter of T_1 is bypassed there will be no signal feedback. Also, the second stage will be very stable because the equivalent dc base circuit resistance of T_2 is very small. Base-point drift of T_2 essentially is determined by, and cannot exceed, that of T_1. This is because as far as dc is concerned, T_1 is used as an emitter follower having a less-than-unity voltage gain.



(a)

(a) Conventional method of biasing transformer-coupled stage applied dc bias to second stage through dropping resistors and secondary winding. **(b)** Simplified connection places bias on stage T_2 directly from the emitter of stage T_1.



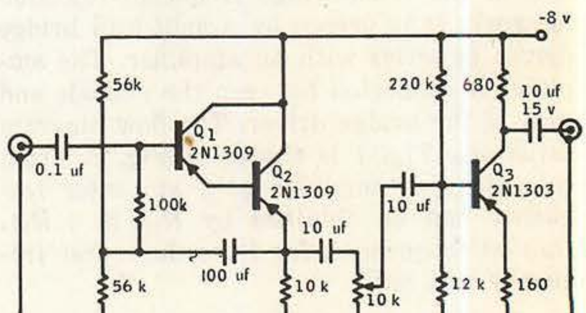
(b)

Not only does the circuit have excellent stability but, because it uses only the transformer as the coupling element, cost and space are reduced, while reliability is increased. Elimination of the biasing voltage divider for T_2 *increases circuit efficiency*. And finally, since maximum gain can be obtained with relatively fewer parts, this arrangement makes the use of a transformer attractive, where RC coupling might otherwise be preferred.

Maxwell Strange, electronic engineer, Goddard Space Flight Center, Greenbelt, Md.

Low-Level Preamplifier Has High Impedance Input

A low-cost preamplifier for oscilloscopes and meters can be a real time saver in the lab. The unit employs the Darlington circuit to obtain a high input impedance. The ac input impedance is approximately equal to β^2 times the emitter resistance of Q_2. The actual measured impedance was in excess of 2.2 megohms. With the input shorted, the noise level is down −78 db as read at the output with a VTVM. Using low-value resistors in the base of Q_1 to establish the operating point provides good dc stability from 25 C to 60 C. Linearity is within 1.5 per cent from 100 μv to 1 mv input.



Darlington low-level preamplifier has 1.5 per cent linearity with input of 100 μv to 1 mv.

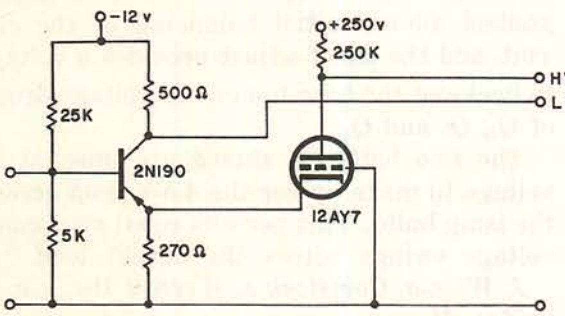
put. Frequency response is ± 2 db from 100 cps to 350 kc. The circuit shown can be readily adapted to other applications.

Edward W. Smith, Senior Engineering Technician, Texas Instruments Inc., Houston, Tex.

Hybrid DC Amplifier Replaces Output Transformer

In a low frequency hybrid amplifier it appeared that an output transformer would have to be used to secure a higher no-load output voltage than was available at the collector of the output transistor. Because very low frequencies were involved, a transformer would have been large and expensive.

It was found simpler and better to use a grounded grid vacuum tube output stage. This gave direct coupling and unusually simple circuitry for the additional stage of amplification.



Grounded grid vacuum tube output stage in low frequency hybrid amplifier takes the place of bulky, expensive output transformer.

Laurence G. Cowles, Electronic Design Engineer, The Superior Oil Co., Bellaire, Tex.

Amplifier Design Provides 20-Megohm Input

A high-impedance transistor amplifier was needed in a metering circuit. Impedances approaching those of vtvm amplifiers were desirable to keep circuit loading to a minimum. An output voltage of 1-v rms across a 3300-ohm load and a frequency response from 10 cps to 200 kc was necessary. Figures 1, 2 and 3 show the evolution of an amplifier that more than met the requirements.

In Fig. 1, bootstrapping was used on a basic emitter-follower circuit to eliminate the shunting effect of the base-bias resistors. Using a transistor with a current gain of approximately 100, the input impedance was measured at 200 K with a 3300-ohm load.

A significant increase in input impedance

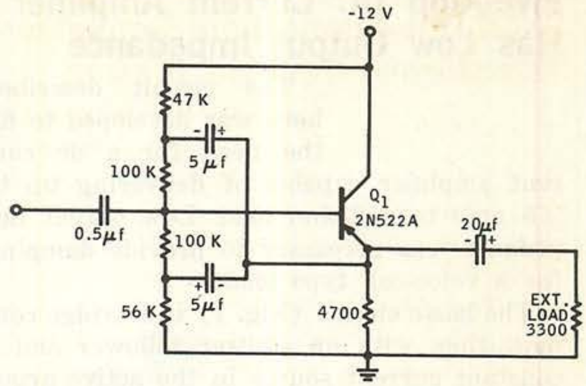


Fig. 1. Basic amplifier uses bootstrap capacitors to eliminate shunting effect of base bias resistors.

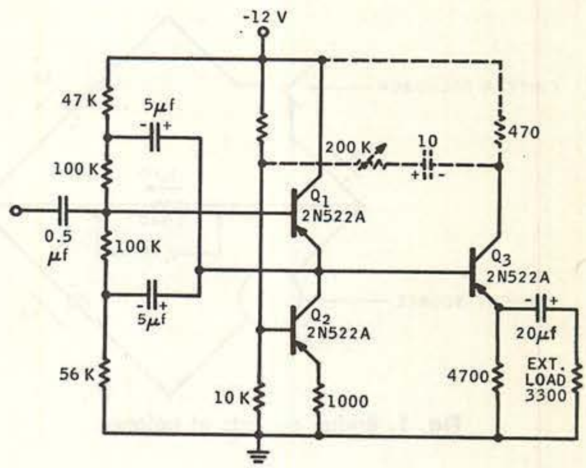


Fig. 2. Transistor in emitter leg of Q_1 and positive feedback (dashed lines) increase input impedance.

was obtained by replacing the emitter resistor of Q_1 with the collector resistance of a grounded-base transistor Q_2, as shown in Fig. 2. To keep the loading as light as possible on the emitter of Q_1, an emitter follower Q_3 was used to couple the load. An impedance of slightly over 1 megohm was measured at the input with the load connected.

Input impedance can be greatly increased by the addition of the components shown in the dashed lines. This, of course, is positive feedback and if overdone will result in oscillation. However, if the feedback adjustment is set with care, the input impedance can be raised as high as 20 megohms before instability occurs.

Gordon D. Svendsen, engineer, Ampex Corp., Redwood City, Calif.

Five-Amp DC Current Amplifier Has Low Output Impedance

The circuit described here was developed to fill the need for a dc current amplifier capable of delivering up to ±5 amp to a 2-ohm load. Low output impedance was necessary to provide damping for a voice-coil type load.

The basic circuit (Fig. 1) is a bridge configuration, with an emitter follower and a constant current source in the active arms. The current source arm of the bridge provides lower average dissipation than a straight resistive arm. This source is composed of transistor Q_4, whose base is held at a fixed potential by a Zener diode. Emitter current from Q_4 flows through a resistive load composed of several paralleled 6-v bulbs. The number of bulbs used determines the amount of current required to raise the emitter of Q_4 to a stable point near the base voltage. Straight resistors could be used instead of the bulbs, but the positive temperature coefficient of the bulbs provides additional current regulation.

The remainder of the circuit consists of three compound emitter followers, Q_1, Q_2 and Q_3. These provide large current gain and moderate input impedance. A zero adjust control allows initial balancing of the circuit, and the offset adjust provides a voltage to buck-out the base-to-emitter voltage drops of Q_1, Q_2 and Q_3.

The two batteries shown are unequal in voltage to make up for the 4.5-v drop across the lamp bulbs. This permits equal maximum voltage swings across the output load.

J. Wisnia, Comstock & Wescott Inc., Cambridge, Mass.

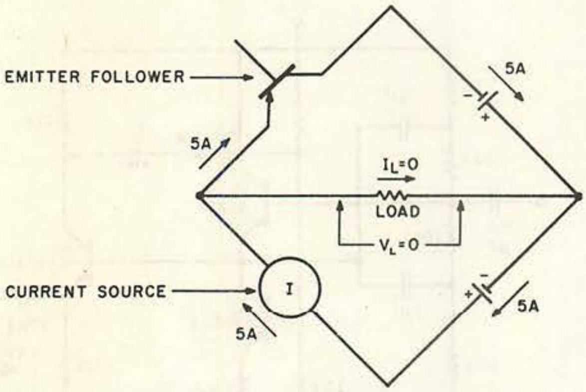


Fig. 1. Bridge currents at balance.

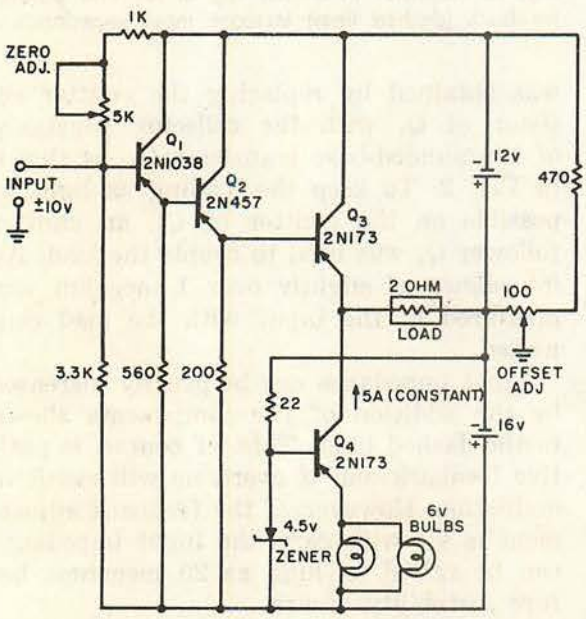


Fig. 2. Current amplifier was designed to drive 2-ohm voice-coil type load with maximum damping.

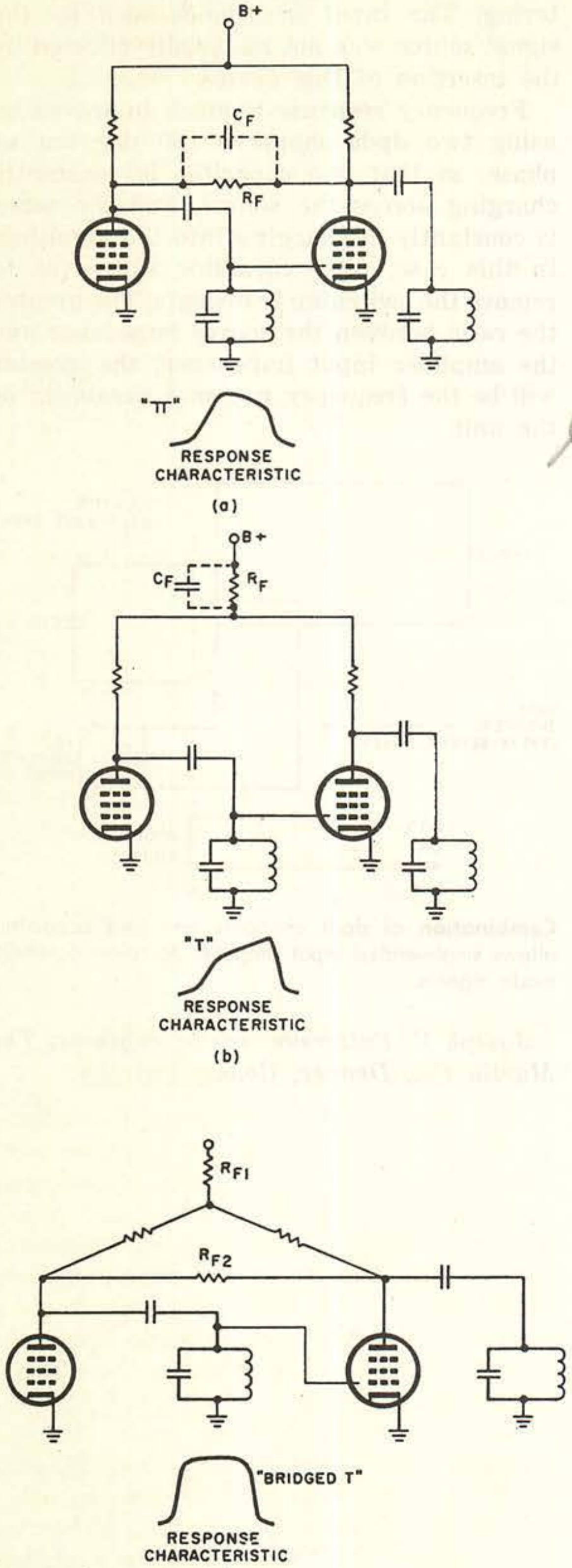
Bridged-T Feedback Yields "Maximally Flat" Response

The "maximally flat" response characteristic obtainable with staggered amplifier pairs can be synthesized by using plate-to-grid degenerative feedback in "feedback pairs." This eliminates the problem of staggering the individual responses. The circuit arrangements of (a) and (b) are most commonly used. Either circuit can be made essentially equivalent to the other by using π to T transformation equations.

In practice, it has been found that the inherent shunt capacitance across R_F—the feedback element in each of the circuits—affects the response by providing either too much or too little feedback at the high frequency end of the band. Thus, the response of the circuit of (a) tends to have a negative slope, while the response of the circuit of (b) tends toward a positive slope.

By combining both feedback configurations in one "bridged-T" network, (c), the effects of shunt capacity across the feedback elements tend to cancel. If the resistors are chosen correctly so that the total feedback remains the same as for either circuit alone, the response will be "maximally flat" as in a corresponding staggered pair.

Ernest I. Fox, engineer, Raytheon Co., Sudbury, Mass.



By combining the amplifier-pair π and *T* coupling configurations of (a) and (b), left, p. 46, the effects of shunt capacity tend to cancel and the response of (c), above, is "maximally flat."

Audio Frequency Amplifier Responds over 10-Cps Bandwidth

Many applications require a band-pass amplifier in the audio-frequency range. The circuit shown in Fig. 1 is intended for such an application.

Transistors Q_1 and Q_2 are connected in the compound (or Darlington) connection. The output current is the sum of the collector currents of the two transistors. By using a bridged-T network in the negative-voltage feedback loop between collector and base of Q_2, the required bandwidth can be ascertained. It is known that the null frequency

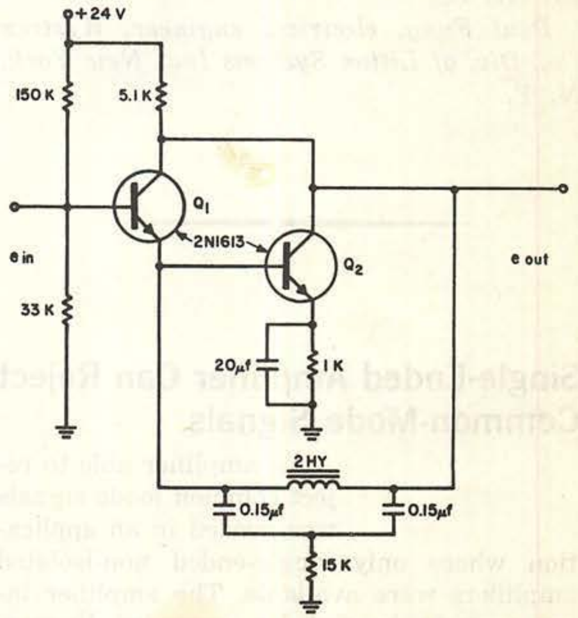


Fig. 1. Darlington circuit combined with bridged-T has 10-cps bandwidth at 400 cps.

ω_o of the bridged-T network is:

$$\omega_o = \frac{2RC}{Q}, \omega_o = \sqrt{\frac{2}{LC}}$$

The fundamental relationship between the gain A of the amplifier with a feedback coefficient of β is

$$A = \frac{\alpha}{1 - \alpha\beta}$$

Consequently feedback occurs at all frequencies except at ω_o where $\beta = 0$. Thus the circuit has a gain of α at ω_o.

The input impedance of the amplifier is very high, for a frequency of 400 cps and a

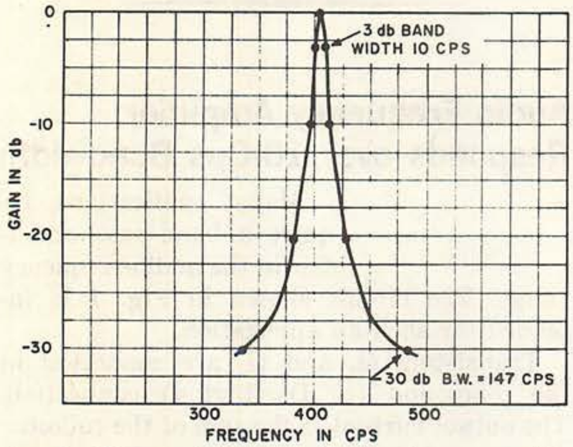


Fig. 2. Characteristic curve for band-pass amplifier has 10-cps bandwidth at 3-db point.

bandwidth of 10 cps. The amplification is over 30 db over a temperature range of −20 to +85 C.

Paul Fung, electrical engineer, Westrex Co., Div. of Litton Systems Inc., New York, N. Y.

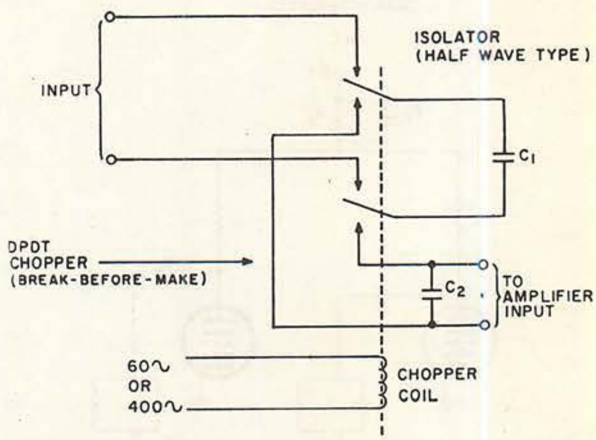
Single-Ended Amplifier Can Reject Common-Mode Signals

An amplifier able to reject common mode signals was needed in an application where only single-ended non-isolated amplifiers were available. The amplifier input was to be fed by grounded thermocouples.

Using a dpdt chopper and two capacitors the circuit shown in the figure was designed to provide the common mode rejection. Also, it was able to isolate completely the thermocouple from the amplifier.

Capacitor C_1 charges up across the signal source, and discharges into the amplifier. Capacitor C_2 provides the desired filtering. The input impedance seen by the signal source was not materially affected by the insertion of this device.

Frequency response is much improved by using two dpdt choppers 180 deg out of phase, so that one capacitor is constantly charging across the source, and the other is constantly discharging into the amplifier. In this case, filter capacitor C_2 serves to remove the switching transients. The greater the ratio between the source impedance and the amplifier input impedance, the greater will be the frequency response capability of the unit.

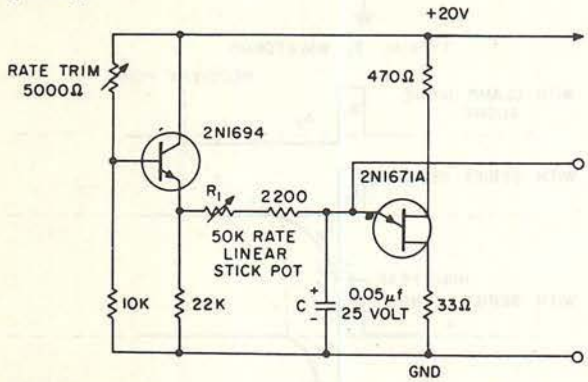


Combination of dpdt choppers and two capacitors allows single-ended input amplifier to reject common-mode signals.

Joseph V. Patterson, senior engineer, The Martin Co., Denver, Colo.

Unijunction Oscillator Has Frequency Trim Control

The unijunction relaxation oscillator, because of its frequency stability and wide frequency range (variable with a single potentiometer), invites uses where pot shaft position accurately describes a given frequency.



Low-interaction trim control for frequency of unijunction relaxation oscillator uses common-collector variable regulator for timing-circuit voltage supply.

In practice, however, variation of unijunction intrinsic stand-off ratio and timing-capacitor variations can destroy the pot shaft/frequency characteristic if simple series resistance trim is employed in the timing network. Further the variation in Thevenin equivalent resistance of a variable divider may be intolerable if divider current must be held to a low value.

The circuit described provides a very low interaction trim control for frequency by employing a common-collector variable regulator in the timing-circuit voltage supply.

The source impedance is low and relatively constant over a wide range with little additional current drain.

Production tolerances of aluminum electrolytic timing capacitors can be absorbed by this circuit, and the least expensive unijunction transistors can be employed.

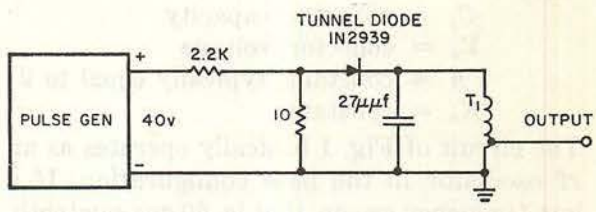
The 5000-ohm rate trim potentiometer provides frequency set that is independent of the main timing potentiometer R_1. The center frequency is approximately 1000 cps. R_1 provides at least a decade variation in frequency and the trim control provides at least one octave variation.

John H. Phelps, manager application engineering, General Electric Co., Semiconductor Products Dept., Syracuse, N. Y.

Gated Tunnel-Diode Oscillator

A pulsed 30-mc oscillator was needed to test radar IF strips. Our commercial signal generator had excessive 30-mc signal feedthrough when the pulse was off. Starting phase of the 30-mc signal varied from pulse to pulse. In addition turnoff time was unsatisfactory in the commercial unit.

The test requirements were met by using the output of a fast-rise pulse generator to power a simple tunnel-diode oscillator. The results were excellent. Approximately 100 mv peak-to-peak were delivered to a 100-ohm load. Changing the pulse amplitude accomplished electronic tuning, varying the frequency from about 28 mc to 35 mc cycles. There is no feedthrough problem since the oscillator is completely off in the absence of a pulse. The oscillator is coherent in that it always starts with the same phase when pulsed.



Radar IF test oscillator provides 100 mv into 100-ohm load.

This device readily may be converted to a sweep-frequency oscillator. Tilt on the input gating pulse will produce fm within the output wave packet. Pulse-to-pulse fm may be achieved by amplitude-modulating the input gating-pulse train.

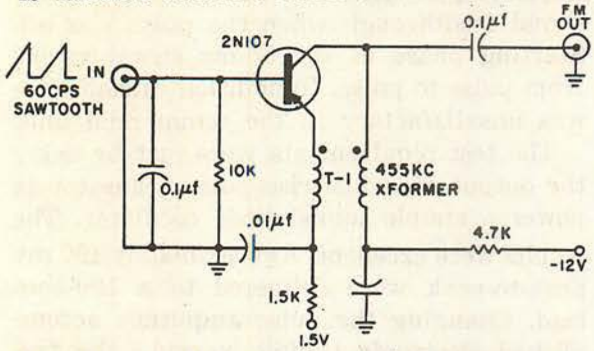
If frequency modulation or electronic tuning is not required, a crystal-controlled tunnel-diode oscillator is desirable, since frequency calibration is inherent in crystal choice.

Paul E. Harris, research associate, Syracuse University Research Corp, Syracuse, N. Y.

Sweep-Frequency Oscillator Operates on Collector Capacity

Low frequency if amplifier design often calls for a sweep oscillator whose frequency can be changed electronically. When the ratio of frequency deviation to center frequency is small, the circuit shown can be used to achieve surprising linearity.

The circuit's mode of operation depends upon the inherent collector capacitance, C_c, of the transistor. This collector capacitance is variable and is a function of the collector voltage, V_c. The analytic relationship can be expressed as:



Operation of electronically-swept oscillator depends upon the inherent collector capacitance, C_c, of the transistor.

$$C_c = \frac{K_o}{(V_c)^{1/n}} \quad (1)$$

where:

C_c = collector capacity
V_c = collector voltage
n = constant, typically equal to 2
K_o = constant

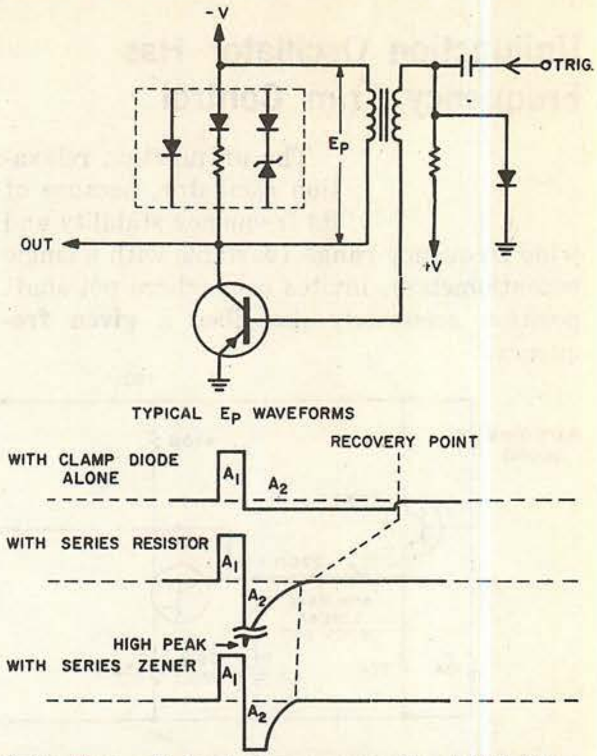
The circuit of Fig. 1 basically operates as an rf oscillator in the base configuration. If a low-frequency sweep, that is, 60 cps sawtooth, is applied to the base, the collector capacitance changes according to Eq. 1 and the desired frequency change is obtained.

Joseph R. Kotlarski, member of technical staff, Hughes Aircraft Co., Culver City, Calif.

Clamp Circuit Improves Blocking Oscillator Duty Cycle

Blocking oscillators find limited application because conventional circuits are unable to achieve a duty cycle much greater than 0.2.

This limitation is introduced by the standard procedure of putting a clamping diode across the transformer primary to protect



Series Zener diode shortens recovery time of blocking oscillator without endangering transistor.

the transistor. After each pulse, the energy stored in the transformer field must be discharged before the oscillator can fire again. The transformer attempts to do this by the well known fly-back pulse.

The clamping diode prevents the fly-back pulse from building up to a significant voltage level, and appears as a very small resistance to the pulse. This lengthens the L/R time constant of the discharge path.

The duty cycle may be improved by increasing the equivalent resistance of the discharge path to allow faster discharge of the stored energy. A resistor in series with the clamping diode will help considerably, but is somewhat unpredictable. A better solution is to use a Zener diode in series with the clamp diode.

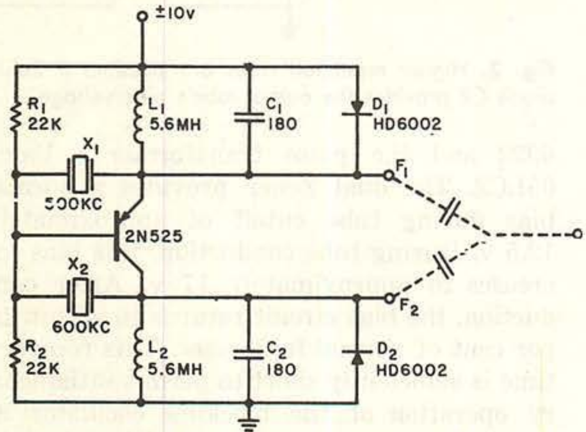
This allows the fly-back pulse to build up to a significant voltage and presents the equivalent of a larger path resistance while still precisely controlling the fly-back amplitude to protect the transistor. In any case, the transistor has to withstand a higher peak voltage, but this is normally no problem.

Roy P. Foerster, group engineer, Martin Co., Baltimore, Md.

Bilateral Crystal Oscillator Has Two-Frequency Output

The bilateral characteristics of the transistor come into play in this two-frequency crystal oscillator. Either of the two frequencies may be selected by applying plus-or-minus voltage to the supply terminal.

When plus voltage is applied, the current flows through D_1 to the emitter of the transistor. Tuned circuit L_2C_2 and crystal unit X_2 then become active and an output is derived from F_2. The L_1C_1 network is shorted out by diode D_1 and crystal X_1 is connected between base and emitter where there is no gain to promote oscillation in this unit.



Either of two oscillator frequencies can be obtained by reversing polarity of supply voltage.

With minus voltage applied to the supply terminal, the transistor operates in the "inverted" mode. Effectively, the collector becomes the emitter and vice versa. In this case, oscillations are generated by L_1C_1 and crystal unit X_1. Output voltage will be delivered at output terminal F_1.

If desired, two additional capacitors (shown by dashed lines) may be added to provide a single output terminal.

Diodes D_1 and D_2 limit the output to 0.7-v rms. Since the transistor is operating in an oscillator circuit the beta of the inverted mode need only be sufficient to produce oscillation. Therefore, it is not necessary to use selected bilateral transitors in this application. This circuit can be used to select frequencies from a remote point by electrical means. Since it uses a single transistor, there is a saving in components.

John M. Tewksbury, principal engineer, Avionics Products, Bendix Corp., Towson, Me.

Diode-Resistor Pair Improves One-Shot Multi Fall Time

In a collector-coupled one-shot multivibrator, Fig. 1, the trailing edge of one of the pulses always has a poorer fall time than the other. This easily can be overcome by adding a resistor and a diode as shown in Fig. 2.

In Fig. 1 when the trailing edge of a pulse occurs, the collector of Q_1 cannot rise to E immediately because C has to be charged. The result is a slow trailing edge, which depends on R_c and C.

If the circuit is designed as shown in Fig. 2 diode D will be reversed biased when the trailing edge occurs. The voltage at the collector of Q_1 will now rise to $ER/(R_c + R)$. This voltage practically will be E if R is chosen much larger than R_c. It can be seen that the fall time now is independent of C. As a result, both trailing edges will have the same fall time.

The diode D has been added to short out the resistor R from the time the one-shot is triggered until the trailing edge occurs. The operation during this period is that of a conventional one-shot.

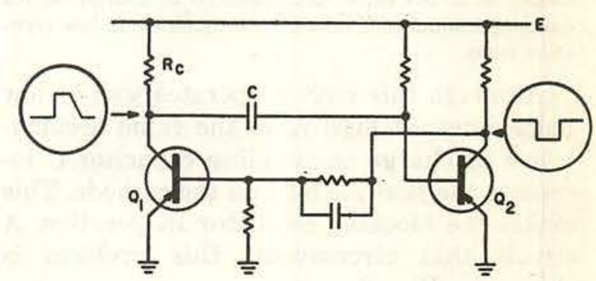


Fig. 1. In conventional one-shot, one of the pulses always has a poorer trailing edge than the other.

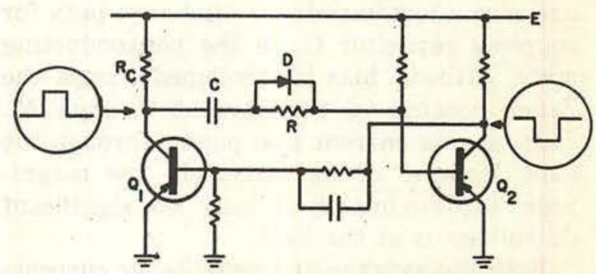


Fig. 2. Both pulses can be squared-up by adding a resistor and diode to the collector with the poorer pulse.

Erik Rosenbaum, engineer associate, The Bendix Corp., Baltimore, Md.

Zener-Biased BO Has High Repetition Rate

The usual method of self-biasing the output tube of a monostable blocking oscillator is to apply a positive voltage to its cathode. But if the output pulse is to be taken from this cathode, it usually is desirable to block this dc bias from the load by a suitable coupling capacitor, as shown in Fig. 1.

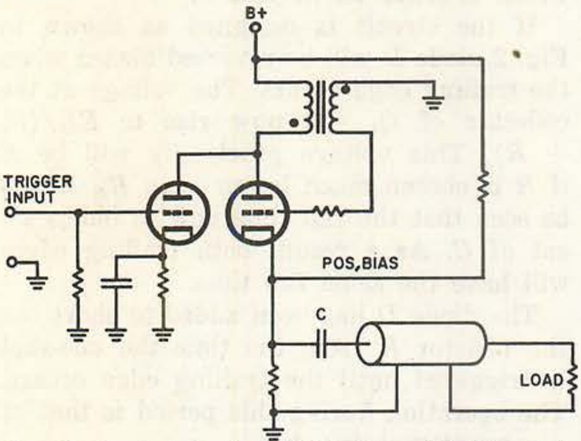


Fig. 1. Usual method of self-biasing a blocking oscillator applies a positive voltage to the cathode of its output tube. But rapid accumulation of charge on the coupling capacitor C restricts the oscillator to low repetition rates.

Although this circuit operates well at low pulse rates, at high rates the rapid accumulation of charge on coupling capacitor *C* increases the positive bias on the cathode. This makes the blocking oscillator inoperative. A circuit that circumvents this problem is shown in Fig. 2.

In this circuit, the Zener diode, *CR*, provides the bias voltage for the output tube and also a low-impedance discharge path for coupling capacitor *C*. In the nonconducting state, cathode bias is developed across the Zener because of the current through R_1. This biasing current also passes through the load. Because of its extremely low magnitude (approximately 2 ma) no significant dc voltage is at the load.

Both the average and peak Zener currents must, of course, be limited to the maximum ratings of the particular diode used. Average Zener current is equal to the average load current. Maximum Zener current is limited by resistor R_2, which usually has a value in the order of a few hundred ohms.

In the circuit of Fig. 2, the Zener is actually two 1N429 Zeners in series. The tube is a

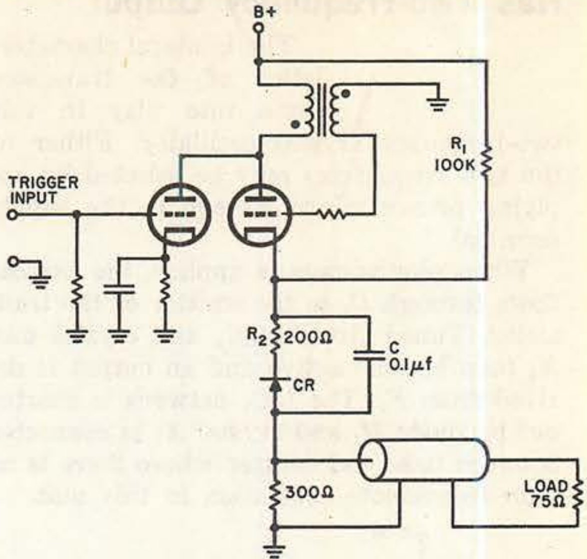


Fig. 2. Higher repetition rates are possible if Zener diode *CR* provides the output tube's bias voltage.

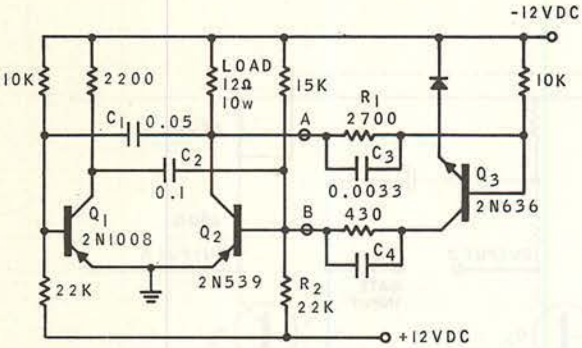
6922 and the pulse transformer a Valor 05LC2. The dual Zener provides a normal bias during tube cutoff of approximately 12.5 v. During tube conduction, this bias increases to approximately 17 v. After conduction, the bias circuit returns to within 10 per cent of normal in 25 μsec. This recovery time is sufficiently short to permit satisfactory operation of the blocking oscillator at pulse repetition rates up to 50 kc.

Edward E. Godin, Sandia Corp., Sandia Base, Albuquerque, N. M.

Regenerative Stage Enhances Flip-Flop Power Output

An inherent deficiency of the astable flip-flop as a source of ac is its load limitations. As the collector load is increased, a point is reached where the available base current is inadequate to drive the transistor into saturation. The circuit shown here makes use of a regenerative amplifier stage to supply this needed base drive.

The circuitry to the left of *A-B* is a typical astable flip-flop, except that the 12-ohm load would prohibit proper operation. As Q_1 starts turning off, the charge on C_2 goes negative causing Q_2 to start turning on. As the collector of Q_2 goes positive (toward ground) Q_3 is forced to turn on, thus supplying another 25 ma (approximately) of base current to Q_2. The peak power delivered to the 12-ohm load will be nearly 10 w, the average power will be 3 w for a 50 per cent duty cycle (square wave output). The current delivered to the load when Q_2 is in saturation is 1 amp. Capacitors C_3 and C_1 increase the output rise time.



Regenerative amplifier boosts flip-flop output to supply 10 w to 12-ohm load.

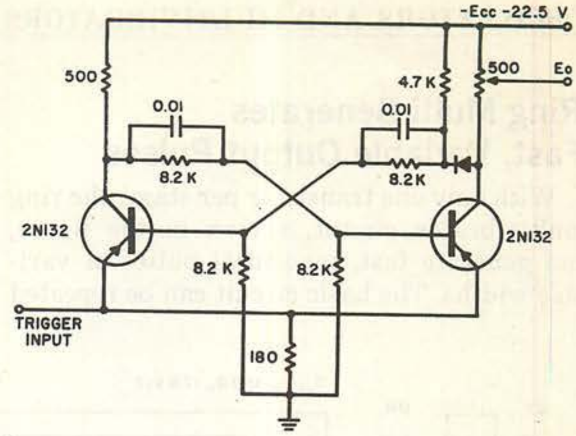
This basic idea easily can be extended to the stable and monostable flip-flops as well. The astable circuit as shown can be used as a dc-to-ac or dc-to-dc converter, by replacing the load resistor with a transformer primary, or as an oven control circuit, by using a heater element as a-load and putting a thermistor in series with R_2 to vary the duty cycle with temperature. The monostable could be used to drive high-current relays, or to switch a larger number of logic stages.

Robert A. Durand, electrical engineer, Bendix Systems Div., The Bendix Corp., Ann Arbor, Mich.

Obtaining Zero Reference For Flip-Flop Square-Wave Output

A transistor flip-flop driven by a blocking oscillator produces excellent symmetrical square waves, but lacks a zero voltage reference. For certain test procedures, such as testing the response time of a potentiometric recorder, a square wave with negligible zero offset is convenient.

As shown in the diagram, this is easily accomplished by obtaining the base bias for one transistor through a separate resistor, isolating the resultant voltage drop by means



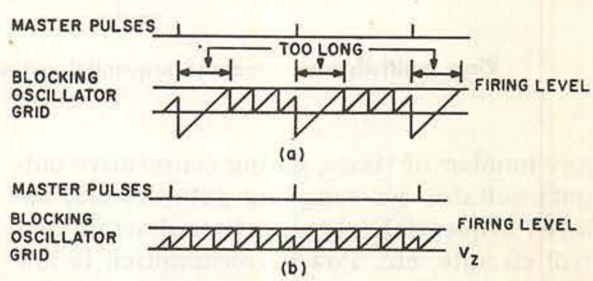
Zero reference in flip-flop is obtained by isolating base bias of one transistor through diode.

of a diode. The remaining zero offset is only that due to the I_{co} of the cut-off transistor—usually at least two orders of magnitude lower than the normal base-biasing current residual in the collector load resistor.

Roy A. McCarthy, engineering designer, Beckman Instruments, Fullerton, Calif.

Zener Diode Helps Set Blocking Oscillator Synch

When a free-running blocking oscillator is to generate timing pulses, but must be started by a master pulse, the master pulse itself can cause the blocking oscillator to give incorrect timing of the first interval. However, if a Zener diode is placed across the blocking oscillator capacitor, the capacitor is always driven down to the Zener voltage and the timing will not be disturbed.



Pulse intervals of blocking oscillator output are made equal by adding Zener diode across circuit capacitor. (a) Circuit without Zener; (b) circuit with Zener.

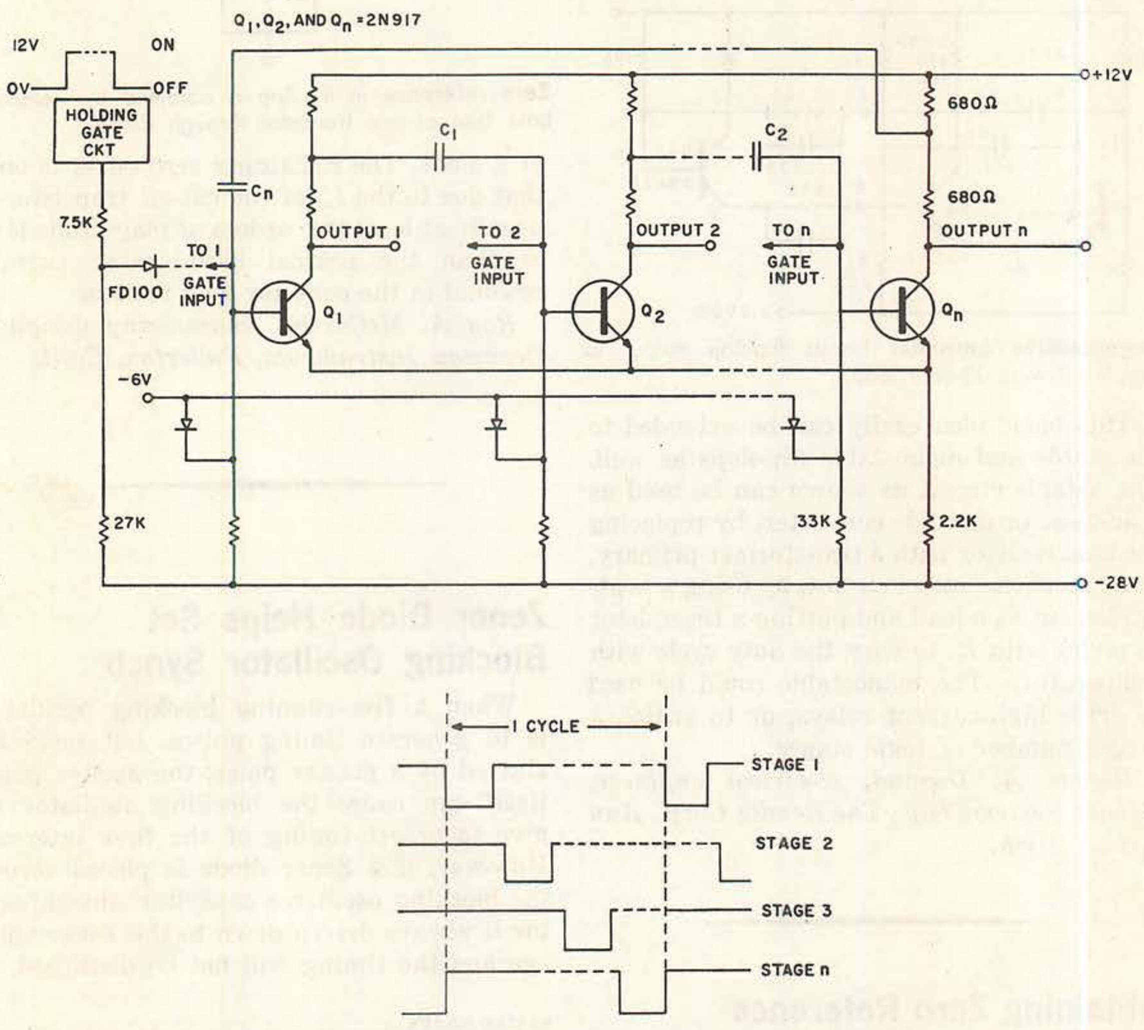
E. R. Roeschlein, electronic engineer, U. S. Naval Avionics Facility, Indianapolis, Ind.

Ring Multi Generates Fast, Variable Output Pulses

With only one transistor per stage, the ring multivibrator circuit, shown in the figure, can generate fast, sequential pulses of variable widths. The basic circuit can be repeated any number of times, giving consecutive outputs suitable for sampling gate drivers, delayed sequential triggers, time-sharing control circuits, etc. Power consumption is low because only one stage conducts at a time.

Each stage is basically a constant-current, nonsaturating circuit providing negative output pulses. The RC combination in the transistor base determines pulse width. Pulse width may vary as much as 3 to 1 for consecutive outputs, without any noticeable affect on the collectors. Fast transistors in the circuit can yield speeds of less than 100 nsec.

Circuit operation is as follows: As each transistor finishes its conducting period and returns to the 12-v level, its collector produces a positive transient. This positive swing, coupled to the following base, turns on the next stage. The sequence can be made repetitive by coupling the last stage back to the first. Various load conditions can be accommodated by adjusting the emitter and collector resistors. By keeping the gated transistor on, the individual holding gates can stop the cycle without prematurely interrupting the sequence.



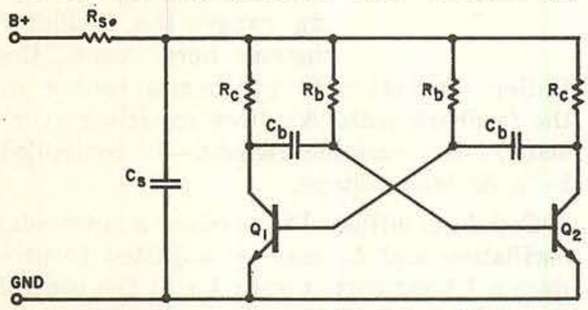
Ring multivibrator produces sequential output pulses whose widths are independently variable.

John A. MacIntosh, Application Engineer, Fairchild Semiconductor Corporation, Mountain View, Calif.

Voltage Control Starts Free-Running Multivibrator

The circuit shown in the illustration provides a simple method of starting a free-running multivibrator. If the multivibrator is used where its output cannot be shorted, it can stall only at the moment power is applied. Both transistors then may saturate.

At low E_s, I_b is reduced more than I_c because V_{BE} becomes significant. In addition, h_{FE} is low at low currents. If E_s rises slowly when power is first applied, neither transistor can saturate for a while. Meanwhile, the multivibrator will oscillate be-



Slow voltage rise assures nonsaturation of transistors in free-running multivibrator.

cause of positive feedback around the loop. Once oscillating, E_s can rise to normal operating level. A large decoupling capacitor, C_s, will cause E_s to rise slowly. Since good equipment design requires a liberal use of decouplers, there is no penalty for additional components.

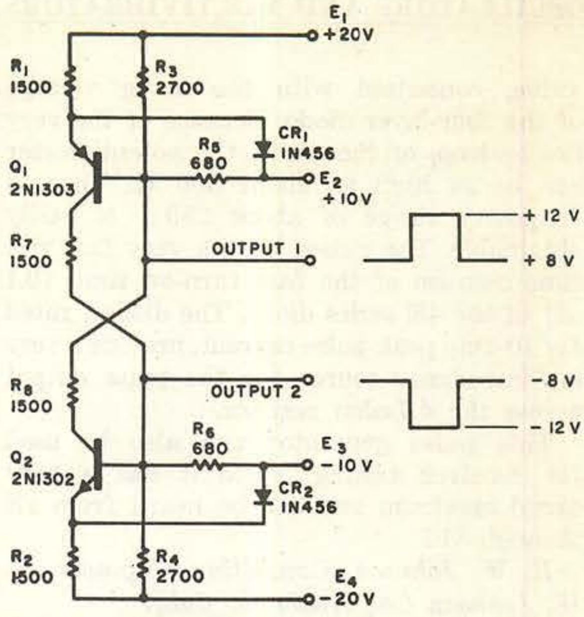
The value of R_c depends upon load and optimum operating point for the transistors. Ratio R_c/R_b should be less than h_{FE} min, and R_bR_c should be approximately f. R_sC_s should be greater than $10R_bR_c$.

Cameron Burley, research engineer, Lockheed Missile and Space Co., Sunnyvale, Calif.

Current-Mode Flip-Flop Has Dual Voltage Outputs

The main advantage of the vertical current-mode flip-flop shown is that two outputs can be obtained at widely separated voltage levels.

In essence, the vertical current-mode flip-flop is one half of a symmetrical current-mode flip-flop.[1] The transistors operate out



Positive and negative voltage levels are available at dual-mode flip-flop output.

of saturation, and are either both on or both off. Resistors R_7 and R_8 serve to reduce the power dissipation in the transistors. They are unnecessary if power is not a problem.

Means for triggering are not shown since there are many possibilities, and the choice depends on the particular application.

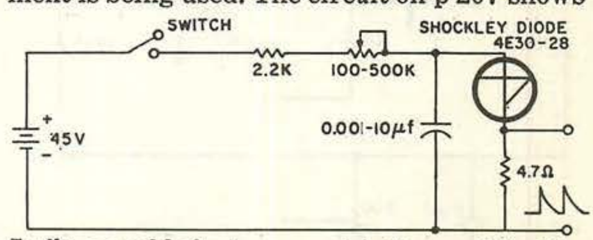
Richard W. Hofheimer, project manager, Non-Linear Systems, Inc., Del Mar, Calif.

Reference

1. Richard W. Hofheimer, "Symmetrical Current-Mode Flip-Flop," *Electrical Design News*, June, 1961.

Shockley Diode Modernizes Neon-Tube Oscillator

An easily-assembled source of pulses sometimes is needed when available test equipment is being used. The circuit on p 207 shows



Easily-assembled pulse source is built around Shockley four-layer diode.

a simple, but by-no-means new circuit that we found useful in such cases. It is a modernized version of the classic neon-tube oscillator, built around a Shockley four-layer diode.

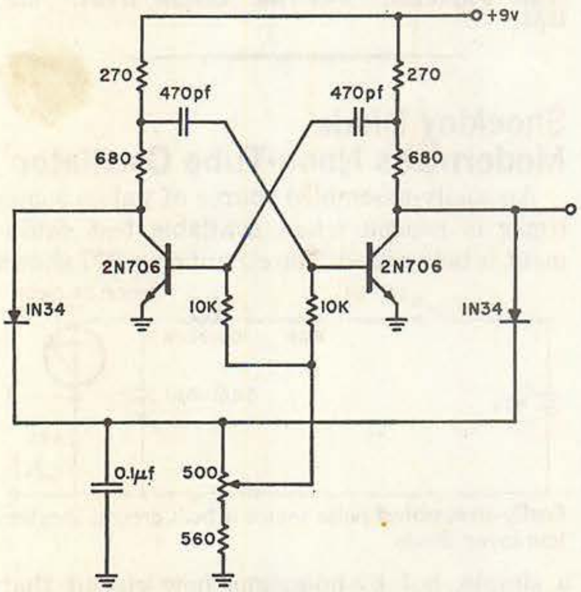
The battery voltage can be any convenient value, consistent with the firing voltage of the four-layer diode. Because of the very low leakage of the diode, the potentiometer can be as high as about 500 K. Thus, a frequency range of about 250:1 is easily obtainable. The pulses have a very fast rise time because of the fast turn-on time (0.1 μs) of the 4E series diode. The diodes, rated for 10-amp peak-pulse current, provide a very low-impedance source for the pulse output across the 4.7-ohm resistor.

This pulse generator can also be used for receiver testing, since it has a very broad spectrum and can be heard from vlf through vhf.

R. W. Johnson, Consulting Engineer, R. W. Johnson Co., Anaheim, Calif.

Tunable Multivibrator Starts Reliably

The circuit shown here is an oscillator tunable from 200 to 300 kc. The most troublesome feature of transistor multivibrators is an occasional failure to start, because both transistors reach saturation simultaneously. Deriving the base supply voltage in the manner shown prevents this and,



Simultaneous saturation of transistors is prevented by obtaining base bias through diode network.

as a small bonus, clips the top portion of the collector waveform, where the voltage would otherwise be rising rather slowly. Tapping the cross-coupling capacitors up on the collector load resistors yields full benefit from the clipping, but is otherwise nonessential. It should be possible to stabilize the frequency against supply voltage change by shunting a Zener diode across the potentiometer circuit, but this was unnecessary in the present application.

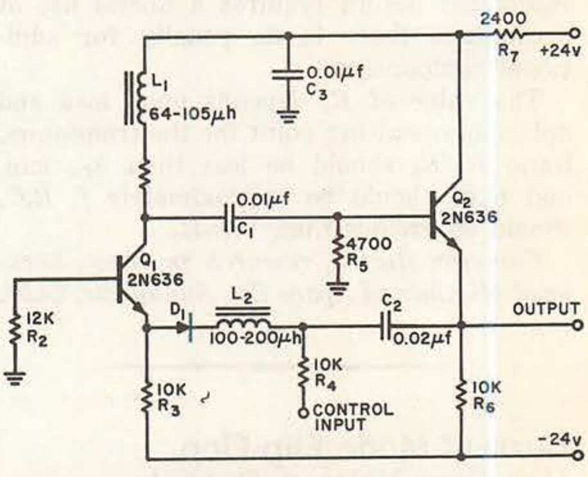
M. W. Egerton, Jr., Towson Laboratories, Inc., Towson, Md.

Voltage-Controlled Oscillator Exceeds 2-to-1 Frequency Range

To obtain a wide tuning range, the oscillator shown here uses the Butler connection to put series tuning in the feedback path. A silicon capacitor (varactor)—the variable element—is controlled by a dc bias voltage.

Coil L_1 is adjusted to produce a sinusoidal oscillation and L_2 may be adjusted to produce a 1.6-mc output with 4 v at the control input. Under these conditions, the frequency deviation is from 1.1 mc when the control voltage is zero to 2.4 mc with 30 on the control input.

Advantages of this oscillator include a low output impedance and simple adjust-



Using component values shown, output frequency can be varied from 1.1 mc to 2.4 mc.

ment. The control impedance is limited primarily by the required response time. Frequency sweep rates longer than 100 μsec could be obtained through a higher value of input resistor.

Charles Turner, electronic engineer, Sanders Associates, Inc., Nashua, N. H.

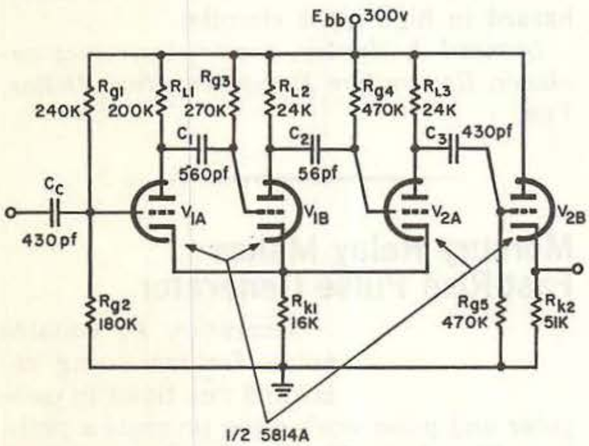
Delayed-Pulse Generator Uses Fewer Components

It often is necessary to produce a pulse of given width, delayed a specified length of time from a reference pulse. The usual circuitry for doing this requires two one-shot multivibrators followed by an isolation-amplifier output, a total of five stages.

It is possible to build such a two-multivibrator combination using only three stages. The isolation amplifier would require a fourth stage. The circuitry has the form shown in Fig. 1, and the switching sequence is given in the table.

Switching Sequence

Step	V_{1A}	V_{1B}	V_{2A}
1	OFF	ON	ON
2	ON	OFF	ON
3	ON	ON	OFF
4	OFF	ON	ON



Using component values shown, output of generator will be a 15-μsec pulse delayed 8 μsec from trigger.

Stages V_{1B} and V_{2A} have identical plate resistors and the current in them is about the same. The value of R_{L1} is much larger than R_{L2} and R_{L3}. The common cathode voltage, then, is determined by the current in tubes V_{1B} or V_{2A} or both. The current through the common cathode resistor (R_{K1}) during steps 2 and 3 when only V_{1A} or V_{2B} is conducting is about two-thirds the current during steps 1 and 4 when both V_{1A} and V_{2B} are conducting. Grid voltage of V_{1A} is held constant at a value that cuts off V_{1A} when both V_{1B} and V_{2A} are conducting, but allows V_{1A} to conduct when only V_{1B} or V_{2A} alone is conducting. The combination $R_{g3}C_1$ and $R_{g4}C_2$ give the desired delay and pulse width.

A positive input pulse will cause stages V_{1A} and V_{1B} to switch. The plate of V_{1B} rising to a more positive voltage only drives V_{2A} temporarily into harder conducting. With V_{1A} and V_{2A} conducting, common cathode voltage is about two-thirds of its initial value. After time-delay capacitor C_1 has discharged to a point where V_{1B} begins to conduct, the plate voltage of V_{1B} falls and stages V_{1B} and V_{2A} switch.

The common cathode voltage does not change during the latter switch and stage V_{1A} remains in a conducting state. After pulse-width capacitor C_2 has discharged to a point where V_{2A} begins to conduct, stages V_{1A} and V_{2A} switch and the circuit resumes its initial conductance state. During this last switch there is feedback from V_{1A} through V_{1B} which opposes V_{2A} turning on. This effort can be reduced by making the value of R_{g3} somewhat smaller than normal.

Stage V_{2B} is an isolation amplifier. The output will be a pulse of width determined by $R_{g4}C_2$ and will be delayed in time by an amount determined by Rg_3C_1.

W. L. Lassetter, engineer, Sperry Piedmont Co., Charlottesville Va.

SCR Drives Cold Cathode Counter Tube

Miniature silicon-controlled rectifiers can be used to produce the high-voltage guide pulses necessary to drive a cold-cathode counter tube. This method simplifies the counter-tube drive circuits and requires less power and space than vacuum-tube drive circuits. The circuit used is shown in Fig. 1.

For reliable operation of the counter tube, a 140-v negative double pulse must be applied to the guides. The pulse widths must be at least 60 μsec, however; for lower counting rates (below 50 kc) longer pulse widths can be used.

A positive 3-v pulse applied to the gate turns the SCR on. Resistor R_1 limits the current through the SCR to a value well below the holding current so the SCR regains its open state upon the termination of the input pulse. The output pulse is applied to the two guides through resistors R_3 and R_4. Capacitor C_2 and resistor R_4 delay the pulse to guide 2. The guide pulses are as shown. Resistors R_6, R_7, R_8, and R_9 supply the guide bias.

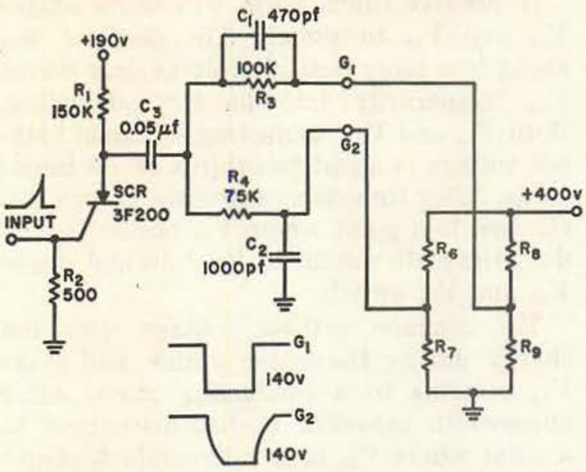


Fig. 1. Double pulse for counter-tube drive is obtained from SCR output. Output waveform is as shown.

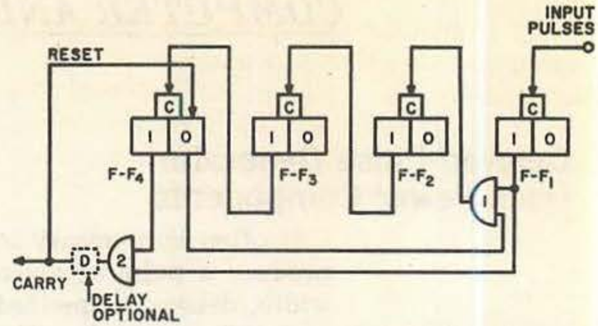
M. W. Egerton, Jr., engineer, Towson Laboratories, Inc., Towson, Md.

Simple Gates Provide Binary Scale-of-Ten Counter

Problems attendant to scale-of-ten counters have been widely discussed and many solutions have been proposed. All of the remedies known to the author have one of the following disadvantages:

(1) The counter is prematurely "forced" to a higher count so that the final state is equivalent to a binary 15. The problem here is that while only 10 states are allowed, the count does not follow the normal binary sequence.

(2) The 11th binary state (1010) is sensed briefly and used to provide a reset to the affected decade as well as a "carry" to the next decade. The problem here is the transient 11th state which may be highly undesirable.



Steering gates provide binary output to count of ten. Carry to next flip-flop occurs on 1-to-0 change of state.

The addition of two simple "steering" gates, as illustrated, solves both problems. Gate *1* allows the binary count to proceed normally to the count of eight at which time flip-flop *4* goes to the "one" state, and Gate *1* is disabled while Gate *2* is enabled. The count of nine occurs normally. When the 10th input pulse occurs, flip-flop *1* returns to the "zero" state causing propagation of a pulse through gate *2*, thereby generating both a reset for flip-flop *4* and a carry to the next decade. A delay element may be inserted on the output of Gate *2* to prevent a timing hazard in high-speed circuits.

Leonard J. Nunley, senior electronics engineer, Recognition Equipment, Inc., Dallas, Tex.

Mercury Relay Makes Fast-Rise Pulse Generator

Generation of suitable pulses for measuring delay and rise times in computer and pulse work often presents a problem. The fact that some mercury relays have a "make before break" characteristic, can be used to generate a fast rising pulse with little overshoot or tilt.

The amplitude versatility and choice of

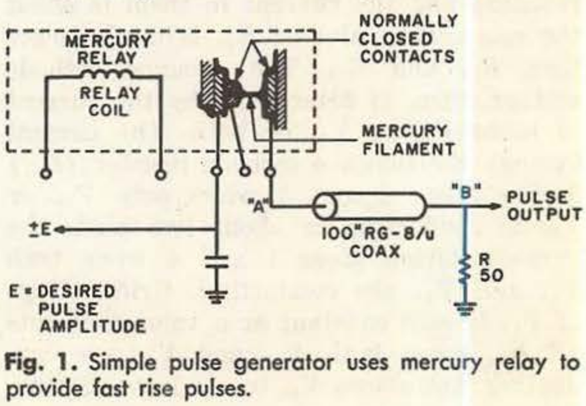


Fig. 1. Simple pulse generator uses mercury relay to provide fast rise pulses.

polarity, plus the excellent wave shape, outweigh the fact that the pulse width is determined by the geometry of the relay. In the units tried (Clarė HG 1005 and Western Electric 275C), pulse width is about 250 μsec. Coil-driving current can be adjust-

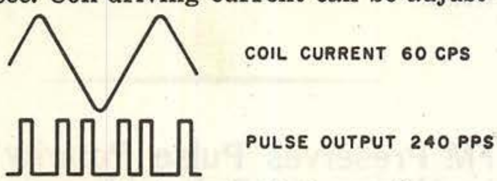


Fig. 2. Pulse output obtained with 60-cps coil current.

ed so the "make" pulse width equals the "break" pulse width.

The 50-ohm resistor also may be returned to *A* voltage if desired to determine the base line of the pulse.

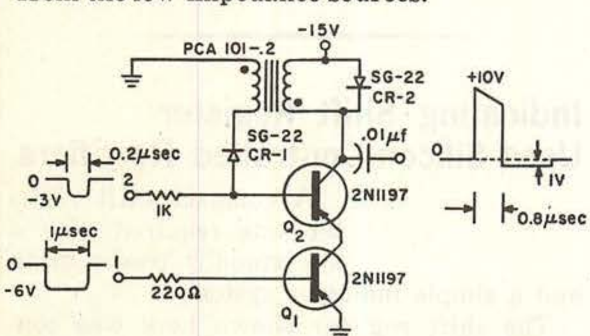
J. R. Bowers, Applied Mathematics Div., Argonne National Laboratory, Argonne, Ill.

Blocking Oscillator-And Gate Produces Standard Output Pulse

Using only two transistors, the circuit shown checks for pulse coincidence and then produces a standard output pulse. Only when signals are applied simultaneously at both inputs will a pulse appear at the output.

A trigger pulse at input *2* is sufficient to trigger the blocking oscillator circuit. Since Q_1 is not in the regenerative path, the gate at input *1* must be as wide or wider than the blocking oscillator pulse to maintain Q_1 in a conducting state until the BO pulse is terminated.

Diode CR-1 prevents trigger signal feedthrough from input *2* into the transformer. The two resistors serve to isolate the circuit from the low-impedance sources.

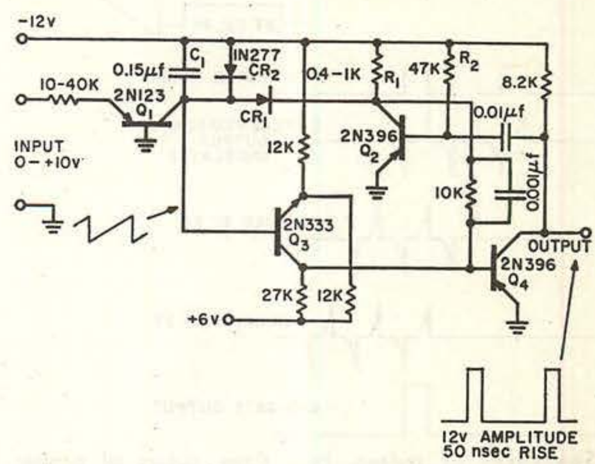


Blocking oscillator-and circuit checks for pulse coincidence, then produces a standard output pulse.

Alfred W. Zinn, project engineer, Farrand Optical Co., New York, N. Y.

Transistorized Voltage-Frequency Converter Operates Linearly

Developing a servo system using a digital motor, we needed a pulse generator that would produce a pulse rate directly proportional to a varying dc input voltage. Conventional generators would not do because (a) they produced an output even though there was no input and (b) they did not respond linearly to the voltage amplitude over a large enough output frequency range.



Pulse generator produces output pulses proportioned to a varying dc input voltage. With input of 10 v dc output is 300 pps.

With a voltage at the input, Q_1 acts as a constant-current source for C_1. This produces a linear slope until the base of Q_3 becomes forward-biased. The negative-going collector of Q_3 then starts the single-shot cycle of Q_2 and Q_4.

Transistor Q_2 is cut off and its collector goes negative. This allows C_1 to discharge through R_1. When the single-shot has completed its cycle the collector of Q_2 goes to ground and the input current again charges through Q_1. This restarts the circuit cycle.

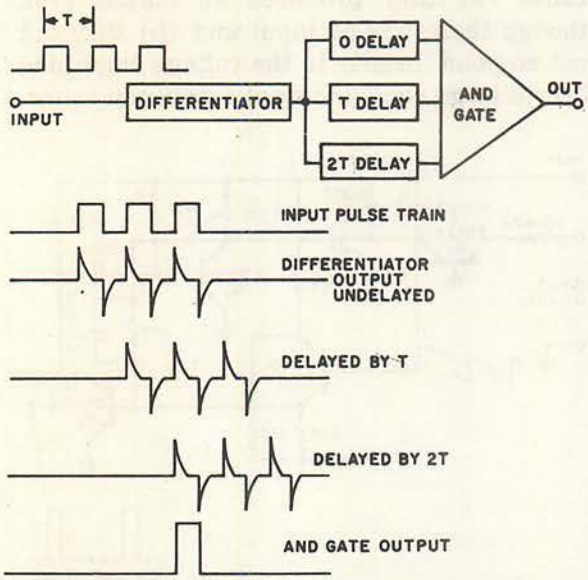
Diode CR_2 passes the leakage currents of CR_1, Q_1 and Q_3. This prevents cycling at low frequencies when there is no input current, and allows the output to reach zero cps.

Modifications can be made to extend the frequency range, alter the input sensitivity and improve the low-end linearity. These steps depend, of course, on how the circuit will be used.

James M. Howe, electronic development, Navy Electronics Laboratory, San Diego, Calif.

Delay-Line Discriminator Detects Sequences of Pulse

A sequence of equally spaced pulses obscured by random or recurrent signals of period different from the period of



Sequences of pulses, here three pulses of proper width and repetition rate, are detected by combination of delay lines and AND gate. Time relation between pulses is also indicated.

the desired pulse sequence, can be detected with the circuit shown in the figure. It extracts the three pulses and generates a trigger when the third pulse is received.

The signal first is differentiated and then fed through parallel delay lines (with delays of *O*, *T*, and *2T*, respectively) to a 3-input "AND" gate designed to respond to positive signals. A trigger pulse is produced when three positive pulses (the differentiated positive peaks corresponding to the leading edges of the original pulses) are presented to the gate simultaneously. The time relationship between the various signals also is shown in the figure.

The circuit discriminates against both narrow and wide random pulses. Increasing the number of pulses in the sequence and the number of delay lines (with delays of *O*, *T*, *2T*, . . *nt*) reduces the possibility of response to spurious signals to an insignificant level.

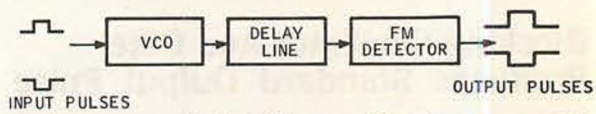
Admittedly a single, tapped delay line might be used; however, pulse attenuation and rise time complications might result. Untapped delay lines are generally easier to obtain, and they allow operation at low impedance levels.

R. Michel Zilberstein, sr. project engineer, Andersen Laboratories, Inc., West Hartford, Conn.

FM Preserves Pulse Polarity In Ultrasonic Delay Lines

When unipolar pulses are used, amplitude modulation can be applied to a delay-line carrier signal with no problem. However, when both positive and negative pulses must be delayed, a different technique is needed, since a dc reference cannot be propagated in this type of delay line.

A simple solution to this problem is to use frequency modulation, as shown in the



Voltage-controlled oscillator and fm detector provide polarized outputs from delay line.

figure. A positive pulse will cause the voltage-controlled oscillator frequency to increase while the negative pulse will cause the frequency to decrease. The fm detector then will yield a pulse of polarity dependent on the input pulse polarity.

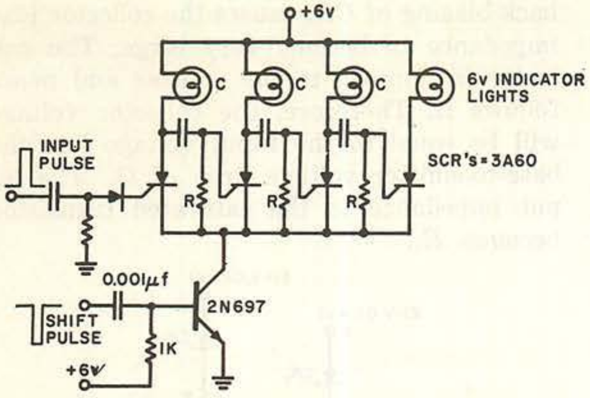
The VCO must have a linear frequency vs voltage characteristic over the dynamic range of interest to preserve accurate pulse amplitude information.

Robert A. Durand, electrical engineer, The Bendix Corp., Bendix Systems Div., Ann Arbor, Mich.

Indicating Shift Register Uses Silicon-Controlled Rectifiers

A compact shift register was required with a low standby power drain and a simple indicator system.

The shift register shown here was constructed using small silicon-controlled rectifiers. Only one SCR per stage of memory is required. Standby power (with register empty) is only a few microwatts per stage. Op-



Use of SCRs provides high speed with little power drain in shift register.

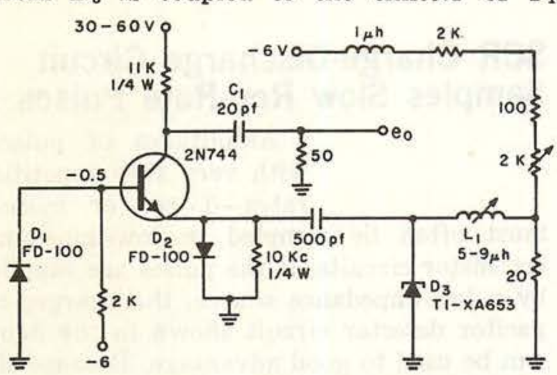
erating speed is high because of the rapid switching of small SCRs. In the four-stage shift register shown, values of R and C are determined by input pulse length.

James M. Loe, circuit design engineer, General Electric Co., Philadelphia 1, Pa.

Tunnel Diode Triggers Avalanche Pulse Generator

An ultra-fast pulse generator utilizing the fast rise-time of a tunnel diode to trigger a transistor in the avalanche mode of operation is shown in the diagram.

To get a nanosecond trigger pulse of about 1 v, a TI-XA653 gallium-arsenide tunnel diode was used in a pre-pulse trigger circuit. This is done because the output pulse rise-time, to some degree, is dependent on the trigger rise-time. The fast negative trigger from D_3 is coupled to the emitter of T_1.



Tunnel-diode pulse generator provides 0.7-nsec rise time.

T_1 normally is biased off by the drop across D_1. This negative trigger pulse is sufficient to drive T_1 into its avalanche breakdown mode. The output pulse, as a result of this breakdown, has a rise-time of 0.7 nsec, pulse width of 2 nsec. Pulse amplitude is 20 v into a 50-ohm load with a prf from 10Kc to 2 Mc. The pulse width may be varied by a transmission-line pulse-forming network.

The circuit must be layed out with strict attention to rf considerations. It should be packaged in a strip-transmission-line configuration. T_1 is a 2N744, which will exhibit avalanche breakdown tendencies with an E_{cc} under 24 v when properly biased.

James R. Williams, design engineer, Sanders Associates, Inc., Inglewood, Calif.

Fixed Interval Timer Gates Random Pulse Stream

A simple method was needed to derive a gate signal whose duration equalled that of a complete random stream of pulses, plus a fixed period of time. It is assumed that the time between any two pulses in the stream does not exceed this fixed period. A short-term stability was all that was necessary.

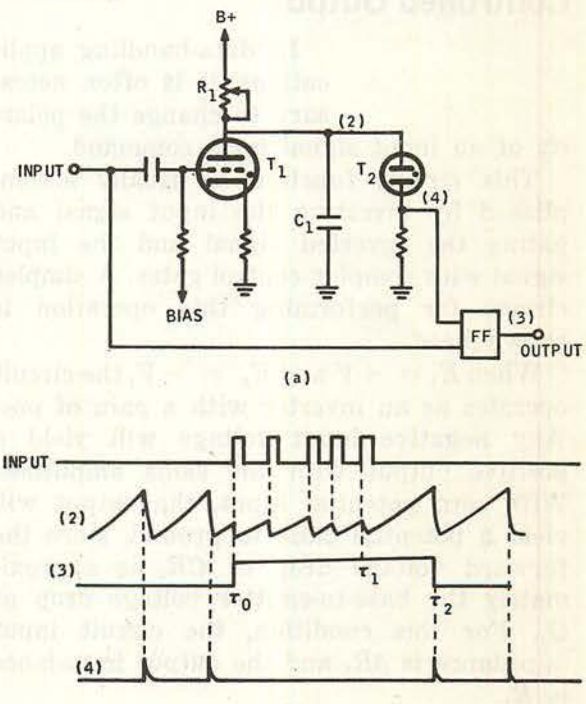


Fig. 1. Flip-flop pulse gate (a) provides gating waveform (b) for random pulse train.

Fig. 1a shows a simple method to accomplish the above. The operation is as follows: T_1, R_1 and C_1 comprise a free-running sawtooth generator. The breakdown voltage of T_1 is higher than that of T_2 so that the output at the cathode of T_2 in the quiescent condition (previous to time t_o in waveform (4) of Fig. 1b) is a series of pulses. When the first input pulse arrives it fires T_1 before its plate voltage reaches the breakdown potential of T_2. Initially the flip-flop was reset by the pulses at point (4), so that the initial input pulse sets the flip-flop at time t_0. At time t_1 the last pulse ends and permits C_1 to charge to the breakdown potential of T_2. The flip-flop is now reset and waveform (3) results, where $(t_2 = t_1)$ is the desired fixed period of time. By varying R_1 the designer can compensate for any long-term changes in T_2.

Irving Bayer, senior member technical staff, Radio Corp. of America, New York, 13, N. Y.

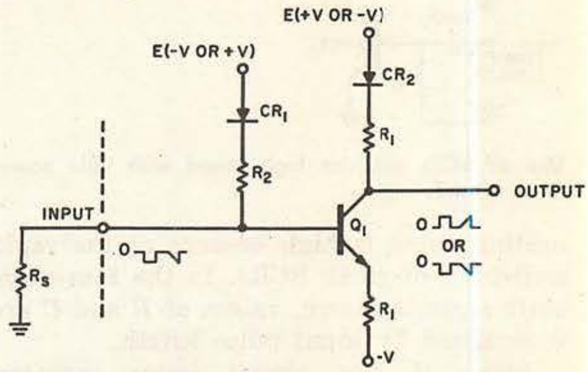
Transistor Stage Yields Polarity-Controlled Output

In data-handling applications, it is often necessary to change the polarity of an input signal upon command.

This circuit function is usually accomplished by inverting the input signal and gating the inverted signal and the input signal with complex control gates. A simpler circuit for performing this operation is shown here.

When $E_1 = +V$ and $E_2 = -V$, the circuit operates as an inverter with a gain of one. Any negative input voltage will yield a positive output with the same amplitude. With zero potential input, the output will yield a potential close to ground, since the forward voltage drop of CR_2 is approximately the base-to-emitter voltage drop of Q_1. For this condition, the circuit input impedance is βR_1 and the output impedance is R_1.

When $E_1 = -V$ and $E_2 = +V$, the circuit acts as a direct path between the input and output. Transistor Q_1 saturates, since the back-biasing of CR_2 causes the collector load impedance to become very large. The collector is clamped to the emitter and hence follows it. Therefore, the collector voltage will be equal to the input voltage less the base-to-emitter voltage drop of Q_1. The input impedance to the saturated transistor becomes R_1.



Reversing supply voltage polarities provides positive and negative outputs.

To keep the output quiescent level at ground, R_2 biases the base to $+V_{BE}$. The value of R_2 depends on the value of R_1, the forward voltage drop of CR_1 and the source impedance R_s. The input impedance to the circuit in the saturated condition becomes R_1 in parallel with R_2. The output impedance is R_1 in parallel with R_2 and R_s.

Control for this circuit can be derived from either a switch or a pnp multivibrator. Inputs can be any unidirectional signal. Refinements of this circuitry technique can adapt the circuit for special applications.

Frank A. Rappolt, engineer, Airborne Instruments Laboratory, Melville, L. I., N. Y.

SCR Charge-Discharge Circuit Samples Slow Rep-Rate Pulses

Amplitudes of pulses with very slow repetition rates—1 sec or more—must often be sampled in low-impedance transistor circuits. If the pulses are supplied by a low-impedance source, the charged-capacitor detector circuit shown in the figure can be used to good advantage. Because silicon controlled rectifiers are used to charge and discharge the capacitor, this unit can be made very large.

At time t_o, capacitor C is discharged to ground by the recycling pulse gating 'on'

SCR 1. This pulse occurs just before the sampling pulse.

When the current through *SCR 1* drops below the holding current, the SCR will turn itself off. *SCR 2* is then fired by the sampling pulse and the generator charges up the capacitor to the pulse voltage. Here, again, the SCR will turn off, when the current through it drops below the holding current, and the capacitor will then discharge through the load.

The maximum current through the SCR will be determined primarily by the turn-on time of the device, the generator impedance, and the amount of capacity used in the circuit. Current through the SCR is given by:

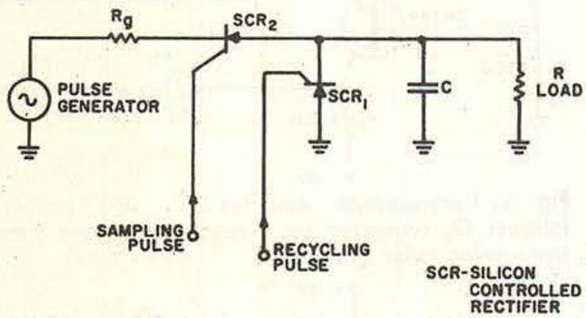
$$i = C\frac{dv}{dt}$$

where C is the capacitance, dv the amount of voltage in which the capacitor is to be charged, and dt is determined by the SCR turn-on time or the R_gC time constant (whichever is greater).

The turn-off point of the SCR depends on when the current through this device drops below the holding current, i_h. The current through the capacitance is determined by:

$$i = \frac{E_g}{R_g} e^{-t/R_gC} \text{ (neglecting } R_L\text{)}$$

When this is below the holding current the SCR will turn off.



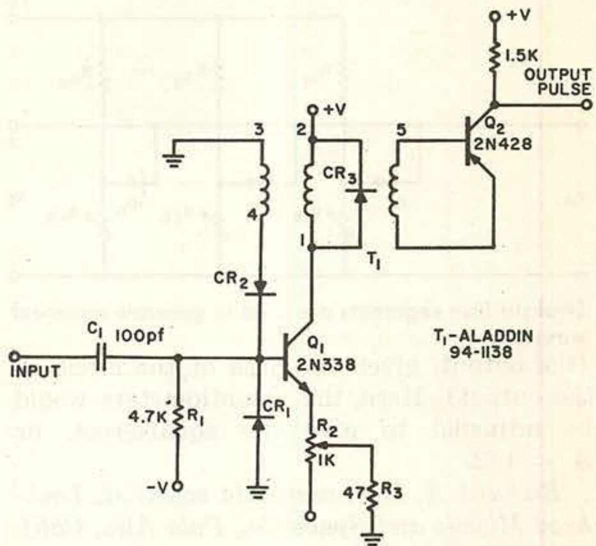
Amplitudes of slow repetition rate pulses are stored on capacitor C which is charged and discharged through SCRs 1 and 2.

Vito Del Guercio, electrical engineer, International Telephone and Telegraph Laboratories, Nutley, N. J.

Variable-Width Pulse Generator Provides Fast Rise/Fall Times

Stable, low-impedance, variable-width pulses suitable for shift register clear, shift or read-in pulses can be obtained from the generator shown here. The pulse width is adjustable from 2 through 15 μsec.

The generator is triggered by a positive pulse or transition which saturates Q_1. Transformer coupling provides a positive regenerative pulse to the base of Q_1. The amplitude of the feedback is controlled by R_2. When R_2 is minimal the feedback sustains an output for approximately 15 μsec when



Pulse width of generator output varies less than ±10 per cent over temperature range for airborne equipment.

R_2 is maximum the output is approximately 2 μsec.

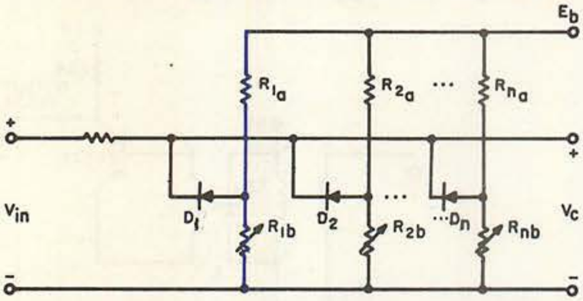
The transformer-coupled output is utilized to cut off a pnp stage and provide a positive output pulse. Diodes CR_1, CR_2 and CR_3 are utilized to prevent negative transitions from causing false triggering. Rise and fall times are from 0.1 to 0.2 μsec.

David W. Gray, Electronic Communications, Inc., St. Petersburg, Fla.

Root Taker Using Biased Diode Networks

Often, a given quantity must be raised to a fractional or a decimal power. This circuit performs such an operation by taking an analog voltage input V_{in} and producing an output $(V_{in})^a$, where A is any decimal number less than one.

Output voltage V_c is a nonlinear function of the input. It is adjusted by means of the variable potentiometers R_{1b} through R_{nb}. The required nonlinear characteristic is generated by straight-line segments approximating the desired fractional power curve. An application for the device would be, for instance, solving for the diameter of a circle



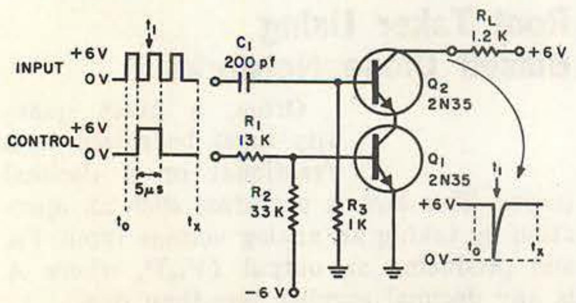
Straight line segments are used to generate nonlinear waveform.

(the output, given the area of the circle as the output). Here, the potentiometers would be adjusted to give the square-root, or $A = 1/2$.

Richard A. Dye, associate scientist, Lockheed Missile and Space Co., Palo Alto, Calif.

Gate Circuit Inhibits Pulses on Command

We required a gate circuit for our logic system that would pass or inhibit positive-going pulses on command. The input (signal to be gated was a Gv square-wave train of pulses that could be at any frequency from dc to 200 Kc. The control (gating signal) was to have two levels; +Gv open, and 0v closed. The circuit decided upon is shown in the diagram. When Q_1 is saturated by a control signal greater than +4 v, the gate is open and an input pulse will, after being differentiated by C_1 and R_3, saturate Q_2. It thereby presents the inverted Q_2 base signal at the collector of Q_2, as shown. In practice, R_L is usually the load in a flip flop or monistable.



Simple gate is opened by saturation of Q_1 by pulse larger than 4 v.

R. Rehfieldt and V. Gavoiosi, electronic engineers, Computer Systems, Inc., Monmouth Junction, N. J.

Two-Transistors, Feedback Produce Free-Running Pulser

A regenerative loop consisting of a common-base amplifier, Q_1, driving an emitter follower, Q_2, produces a simple, free-running pulse generator, Fig. 1. Pulse width is controlled by R_2C and pulse repetition rate is controlled by R_1C.

With only a slight modification the circuit can be triggered as shown in Fig. 2.

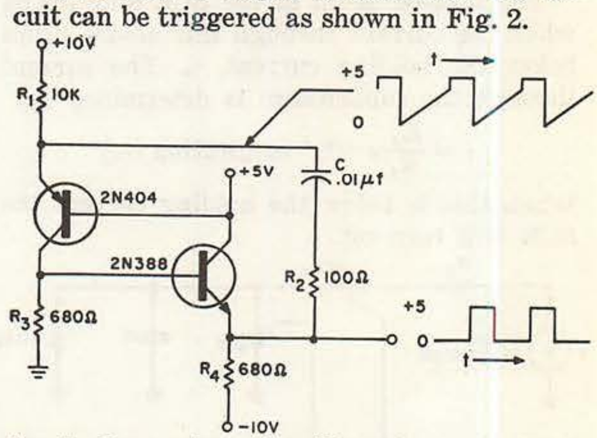


Fig. 1. Common-base amplifier Q_1 and emitter-follower Q_2 connected by a regenerative loop from free-running pulse generator.

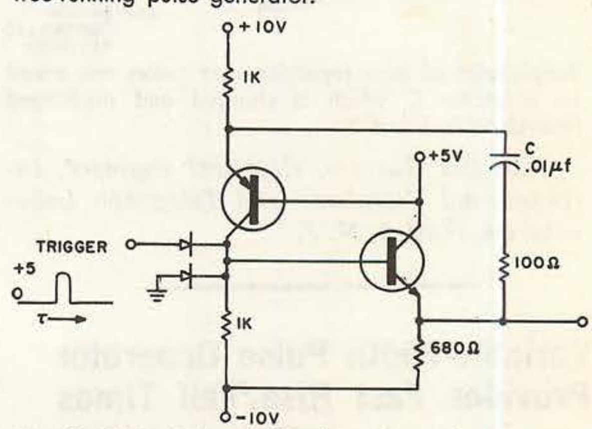


Fig. 2. With slight modification circuit of Fig. 1 may also be externally triggered.

Paul Lucas, design engineer, Adage, Inc., Cambridge, Mass.

Modified Detector Adds Markers For Frequency Response Tests

The frequency response of tuned circuits is usually observed on an oscilloscope by using a swept-frequency generator and an am detector. One or more fixed-frequency sources, called markers, help to identify specific frequency points on the response curve.

By slightly modifying the usual detector circuit, two additional markers can be created for each actual one. These are located symetrically on either side of the original marker.

The arrangement is useful for the production alignment or the checking of band-pass circuits. Here, the original marker is set to the desired center frequency and the additional markers indicate the desired 1 db or 3 db bandwidth points.

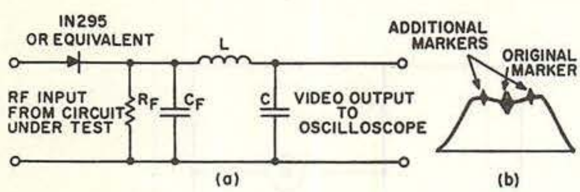


Fig. 1. (a) Detector, modified by *LC* combination at output, produces additional marker pips for frequency response tests of tuned circuits. (b) Typical oscilloscope pattern circuits such as tuned If or rf amplifiers.

Shown in the figure are the modified detector circuit and a typical frequency response pattern that can be obtained with a single marker source. The detector circuit is conventional except for the series *LC* combination on the output side.

The central "pip" in the response curve is the usual one caused by the swept-frequency zero-beating the marker generator frequency. Beat frequencies significantly higher than zero are normally bypassed by the detector filter capacitor C_F. But, at the frequency for which L is series resonant with C and C_F, there is an enhancement of the beat frequency component in the output.

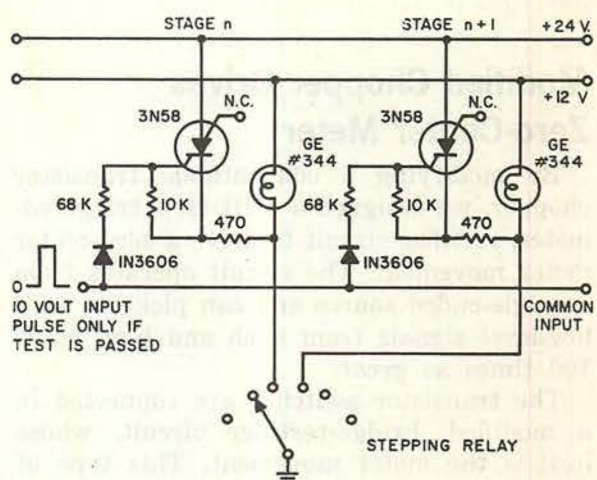
Because of this series resonance, the additional "pips" are created on either side of the zero-beat point. The frequency separation of the side markers from central marker is $1/2\pi LC_T$, where C_T is the series combination of C and C_F. For best results, a high impedance oscilloscope input should be used. However, any capacitance associated with the input simply will be added in parallel with C.

In equipment to be shipped, it has been found useful to connect the modified detector to a test point to aid in-the-field troubleshooting.

John T. Zimmer, staff engineer, Raytheon Co., Wayland, Mass.

Visual Readout for Tester Uses Silicon-Controlled Switch

Stepping relays are frequently used to sequence a series of tests on electronic components such as transistors and diodes. The circuit shown in the diagram will provide a visual indication of which test is being performed and if the test has been passed.



Silicon-controlled switches and lamps indicate those tests in a sequence that have been passed.

The stepping relay position is indicated by a lamp connected from ground to +12 v. If the device passes the test, a 10-v common input pulse turns on only the silicon-controlled switch at the relay location. When the relay sequences to the next location, the grounded terminal of the lamp is returned to +24 v through the silicon-controlled switch and 470-ohm resistor in series. Consequently the lamp remains on at reduced intensity, indicating the test was passed. If the device fails the test, no input pulse occurs and the silicon-controlled switch is not turned on. The lamp will then turn off as the relay sequences to the next position.

If a low-impedance input line is used, the 1N3606 diodes can be eliminated since their purpose is to isolate the line from changes in load.

The lamps are reset by opening the +24-v supply momentarily. If voltage is reapplied too rapidly the lamps may relight. This can be avoided by replacing the type 3N58 silicon-controlled switches with 22-K resistors connected between the anode gate (N.C.) and a 24-v supply which is not interrupted. An alternate solution is to connect a 0.005-μf capacitor across the switches.

Erwin Pacia, General Electric Co., Semiconductor Products Dept., Syracuse, N. Y.

Modified Chopper Drives Zero-Center Meter

By modifying a conventional transistor chopper, we designed a drift-free, triggered-meter, rectifier circuit to drive a zero-center meter movement. The circuit operates from a single-ended source and can pick out very low-level signals from hash and hum levels 100 times as great.

The transistor switches are connected in a modified, bridge-rectifier circuit, whose load is the meter movement. This type of circuit has several inherent advantages over other triggered-meter rectifier circuits.

Since the transistors are used only as on-off devices, and no dc amplification is required at the output, there is essentially no dc drift. Zero adjustment is not needed.

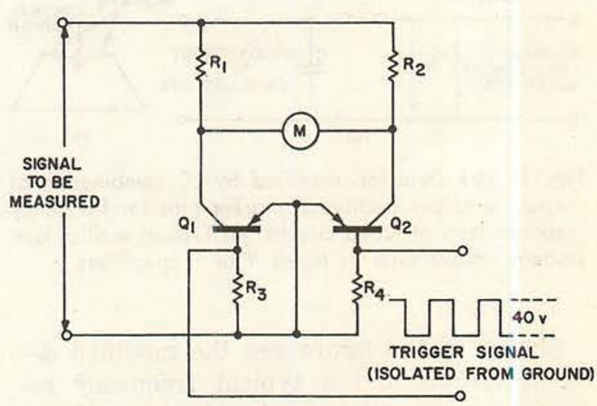
Also, since it is a bridge circuit it operates at high efficiency from a single-ended signal source. This eliminates the need for transformers or phase inverters. The circuit can operate directly from a signal source of at least 5 v rms. And, it can easily be isolated from ground.

The circuit is particularly adaptable to meter amplifier circuitry using current feedback for linearization, since the ac return is completely floating. Designed with the components listed, the largest detectable error was 0.3 per cent of full scale.

The trigger signal need not be a perfect square wave. It should be at least 40 v peak-to-peak and must be isolated from the system ground. In this circuit, the trigger was derived from a 35 v rms transformer winding with a 10-K series resistor and two 20-v Zener diodes back-to-back.

Input impedance of the circuit is approximately 20 K. Input voltage should be approximately 4.4 v rms (in phase with the trigger signal) for full scale deflection.

A balanced dc output voltage for driving recorders or external control circuits is available across the meter terminals.



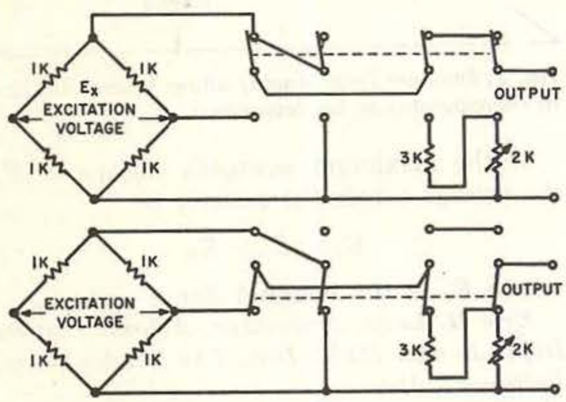
Transistor chopper connected in modified bridge rectifier circuit, drives zero-center meter with essentially no dc drift. Circuit values are:
R_1, R_2 —33 K, 1/2 w, 5%
R_3, R_4 —27 K, 1/2 w, 10%
Q_1, Q_2—2N220 transistors
M—100-0-100 μa, 1,000 Ω, movement

Don Gephart, Staff Engineer, Ohio Semiconductors, Div. of Tecumseh Products Co., Columbus, Ohio.

Technique For Simulating Strain-Gage-Transducer Signal

It is generally agreed that shunt-resistance signal simulation is not suitable for the semiconductor strain-gage transducer. This is due to the relatively large change in bridge resistance with applied input and the high temperature sensitivity of the semiconductors. However an alternate method could be used. The components needed are shown in the diagram.

When a simulation signal is required, the output terminals of the transducer are shorted out, and one of the output connections is transferred to one of the input terminals. The voltage across the output leads will be close to half of the excitation voltage,



Attenuation section of strain-gage signal simulator could be incorporated in common signal conditioner.

dependent on the matching of bridge resistors.

The ratio between this voltage and the rated output voltage is slightly more than 2 to 1. To bring the simulated signal down to a voltage suitable for recording or transmission, an attenuator, or voltage divider is used. Using a 3-K and 2-K resistance in series will provide a voltage across the 2-K resistance of about 2/5 x $E_x/2$. With 25-v excitation this will produce a signal fairly close to that of rated output. The described resistance combination will also provide an input impedance to the signal conditioner or recorder, similar to that of the transducer proper.

Any malfunction of the strain-gage bridge will show up as a drastic change in the simulated signal. The latter is in a fixed ratio relationship with the rated output and also directly proportional to the excitation voltage.

The temperature effect on the simulated signal is negligible, amounting to less than 0.2 per cent of rated output over a temperature change of 200 F.

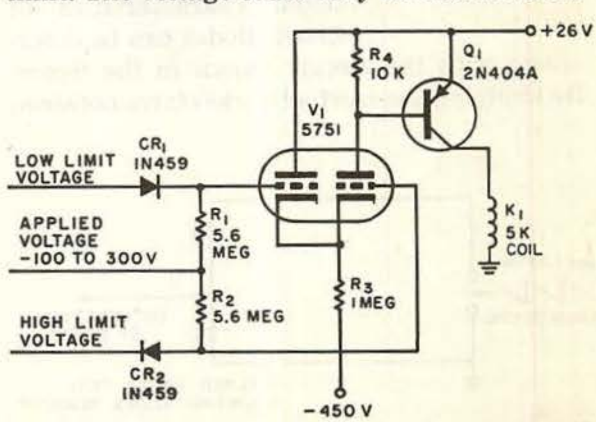
The difference in simulated signal with no pressure or rated pressure applied to the transducer, is about 0.5 per cent of rated output.

This form of simulation also is applicable to the standard wire-type strain-gage transducer. For a 500-ohm bridge with a 10-v excitation and 30 mv rated output the attenuation would be about 200 to 1, which could be accomplished with a 100-K and a 500-ohm resistor in series. Again the impedance presented to the recording or readout device basically would be unchanged. In a telemetry system this method will have the advantage of the elimination of the calibration resistor. The attenuator will be common for all similar transducers. Shorting the output terminals and switching over the output lead easily can be done on a stepping switch, which also will introduce the attenuator into the output circuit when signal simulation is required.

Sigmund Meieran, engineer, The Boeing Co., Seattle, Wash.

Voltage Band Comparator Uses Tube-Transistor Circuit

We require a simple circuit that would energize a relay as long as an applied voltage remained within a high and low limit. The voltage could vary several hundred



Tube-transistor circuit opens relay when applied voltage is outside of preselected limits.

volts, but a maximum of 1 μa could be drawn from the source. The circuit was required to respond within about one volt of the limits, which could be anywhere in the range of applied voltage.

The circuit designed to meet these requirements is shown in the diagram. As long as applied voltage is within limits, neither of the diodes CR_1 or CR_2 conducts, and both grids see the same potential. Current is drawn through the left section of the tube, turning transistor Q_1 on, which in turn keeps relay K_1 energized.

If the applied voltage drops below the low limit, diode CR_1 conducts, and a voltage develops across R_1. If the applied voltage rises above the high limit, diode CR_2 conducts, and a voltage develops across R_2. In both cases the voltage drop appears as a potential difference between the two grids, shutting off the left section. Thus, outside the limits, Q_1 is turned off, and relay K_1 is deenergized.

G. Richwell, staff engineer, Reflectone Electronics, Inc., Stamford, Conn.

Test Set Displays Zener Diode Characteristics

Several important functional characteristics of Zener diodes can be determined with the circuit shown in the figure. By applying the sawtooth waveform obtained from the scope, the resultant display of the Zener diode ramp allows:

1. The Zener voltage to be measured within the accuracy limits of the scope.
2. The Zener voltage change over the applied current range to be determined.
3. Some measure of the transient response characteristic to be obtained if timebase of the scope is varied.
4. The characteristics of the "knee" to be observed.

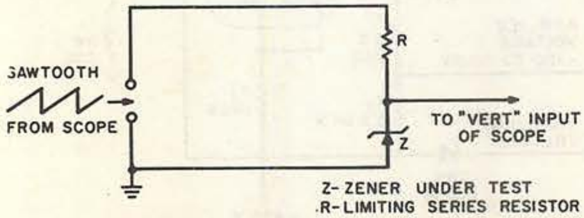


Fig. 1. Sawtooth wave form from scope is applied to Zener diode circuit.

The value of the current-limiting resistor is determined from the manufacturers ratings of maximum current or power and the maximum voltage applied to the circuit.

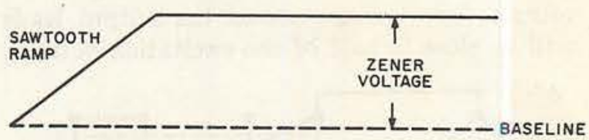


Fig. 2. Resultant Zener display allows several functional characteristics to be determined.

If the maximum sawtooth voltage is E_p, the voltage across the resistor is:

$$E_R = E_p - E_Z$$

where E_Z is the nominal Zener voltage.

Eric H. Levy, supervisor, Avionic Service Dept., Bendix Radio Div., The Bendix Corp., Baltimore, Md.

Automatic Continuity Detector Performs Double Test Function

This simple and reliable circuit will check the continuity of a conductor and simultaneously test that conductor against all other conductors in the system under high-voltage conditions.

The automatic continuity tester, of which this circuit is a part, can test 500 circuits at the rate of 10 circuits per second with an adjustable high-potential voltage from 100 to 1000 v dc. The components utilized in the detector consist entirely of resistors and a Western Electric type 280R differential relay.

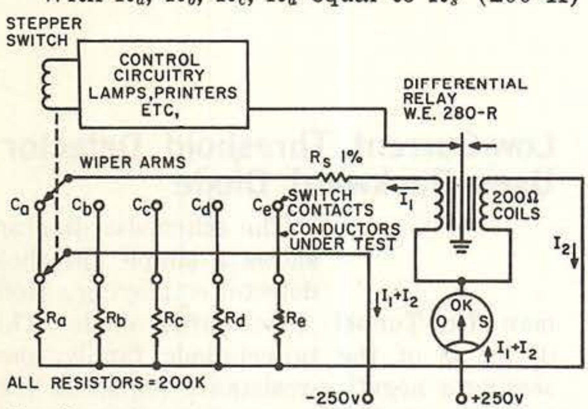
A relay of this type is constructed with a coil consisting of at least two electrically identical windings and a magnetically biased armature for sensing the direction of current through the windings. As shown in the schematic diagram, the continuity testing function consists of a summation of currents $I_1 \pm I_2$ through the two windings of the relay. When $I_1 = I_2$ the relay remains inactive, but when the balance is upset by either the opening of C_a or shorting of C_a to any other conductor, the relay operates.

This statement can be verified by starting at the +250-v point and tracing the circuit through both sides of the relay to the −250-v point.

The hi-pot function, performed each time a conductor is checked for continuity, is achieved as follows:

While C_a is being tested, a potential of −250 v is applied through the stepper switch arm; at the same time, +250 v is applied to all other conductors through their associated resistors, R_b, R_c, etc. Hence, the difference of potential existing between the conductor under test and all other conductors is 500 v.

With R_a, R_b, R_c, R_d equal to R_s (200 K)



Simultaneous continuity and high-voltage tests can be performed by means of a differential relay.

the circuit can detect 300-f K shorts between conductors. This sensitivity is made possible by the mechanical adjustment capabilities of the differential relay, which permit reliable and consistant operation down to 0.5 ma. Since these relays are factory-adjusted, only a slight modification is necessary and consists of manipulating the pole pieces until a current of 0.0005 amp operates and holds the armature against the contact associated with the energized winding.

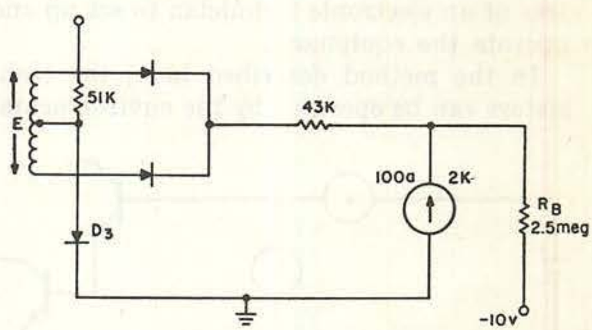
One pole piece then is adjusted to keep the armature on one side, as indicated on the schematic, while maintaining a 0.0006-amp operating sensitivity. This adjustment insures that the relay will remain balanced or inactive when no current or equal currents flow through the two windings.

To obtain the maximum amount of information from a reject circuit, a milliammeter was inserted in the total current line. Full scale was calibrated to twice $I_1 + I_2$ so that a good circuit was indicated on the meter as half-scale reading. When an "open" reject (less current) was encountered, the meter read less than mid-scale and when a "short" reject was encountered the meter read more than mid-scale. With the meter labeled "open," "good" and "short", valuable troubleshooting information is gained.

Frank L. Egenstafer, project engineer, Jerrold Electronic Corp., Hatboro, Pa.

Biased Diode Reduces Non-Linearity of AC Meters

Ac metering circuits are non-linear because of the non-linear impedance of rectifier diodes. The effect of the non-linearity can be reduced by making the excitation voltage very large compared with the diode voltage drop of approximately 0.5 v. However with transistorized circuitry, large voltages are troublesome.



Biased diode D_3 in ac metering circuit overcomes the effect of 0.5 v diode voltage drop; circuit operates with greater linearity.

An alternate approach is to add the dc biased diode, D_3, and bleeder resistor, R_B as shown in the figure. Sufficient dc bias current is passed through D_3 to exceed the maximum (peak) current required by the meter. The best linearity, it has been found, is achieved with the bias current through D_3 approximately twice the peak current.

The circuit values indicated in the schematic yielded 1 per cent linearity with a 100 μa, 2,000-ohm meter movement and an applied voltage, E, of 10 v rms. To achieve similar results without D_3, required voltages greater than 50 v.

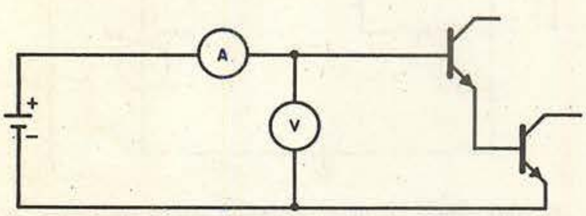
Carl L. Bose, senior engineer, Nortronics Div. of Northrop Corp., Hawthorne, Calif.

Burned Out Transistors Save Time and Money in Heat Tests

In design of circuits using power transistors, heat-dissipation tests frequently are run with the transistors mounted on heat sinks. To be valid, these tests should be run using mockup heat sinks and equipment as close to the final package design as practical. The same type of power transistor should be used in the mockup and the final package.

With some types of silicon power transistors costing $100 or more a piece, it is expensive to risk these valuable components in these tests. Also, the control of dissipation in multiple transistor circuits requires the services of an electronic technician to set up and operate the equipment.

In the method described here, the transistors can be operated by the environmental



Simple series circuit provides controlled dissipation for heat tests.

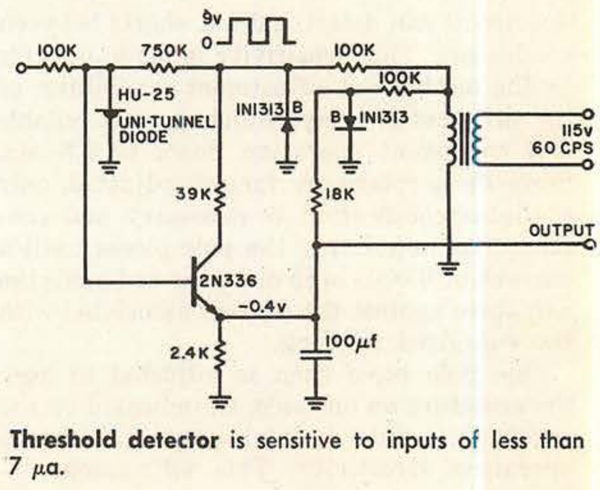
test technicians because a simple, stable series circuit is used.

We have found that burned-out (shorted collector-to-emitter) transistors of the same case type as those to be used in the equipment make good resistors for these heat tests. (Burned-out transistors are usually in good supply around development labs.) The burned-out transistors are mounted on the heat sinks and a low voltage high-current source is connected base-to-emitter or collector-to-emitter. The base-to-emitter resistance usually is higher. (Typical resistance for a 2N1016 was 1.8 ohms for a current of 10 amp.) By monitoring current and voltage, the wattage desired can be dissipated in the exact mechanical configuration of the final equipment. In most burned-out transistors, the base diode is still intact so the current must be passed in the forward direction of the base diode.

Frederick W. Paget, advanced development engineer, Sylvania Electric Products, Inc., Sylvania Electronic Systems, Needham, Mass.

Low-Current Threshold Detector Uses Backward Diode

The schematic diagram shows a simple threshold detector employing a Hoffman Uni-Tunnel (backward) diode. This device is of the tunnel-diode family, possessing a negative-resistance region at very low current. Peak current for this diode is at about 20 μa.



Threshold detector is sensitive to inputs of less than 7 μa.

The dc input signal is applied to the HU-25, together with a square-wave current obtained from one of the 1N1313 regulators. Below the trip point the diode acts essentially as a very low impedance, and the square-wave voltage across it is on the order of 10 mv or less. As the dc input cur- at which this ac voltage will suddenly jump to about 0.4-v peak. In the circuit shown, this occurs for an input current of slightly less than 7 μa. Input power is less than 5 μwatt.

The 2N336 is supplied with a 9-v square wave supply to the collector through the 39 K resistor and an emitter bias of about 0.4 v. Under these conditions the collector voltage is a square wave of about 7-v peak. When the HU-25 switches the 0.4-v signal on the base is sufficient to saturate the transistor and the output drops to zero.

Richard F. Shea, consulting engineer-electronics, General Electric Co., Schenectady, N. Y.

Comparator Circuit Simplifies Integrator Time-Constant Check

A conventional method for checking the time-constant accuracy of electronic integrators is shown in Fig. 1. This procedure is costly and elaborate in that it involves an accurately controlled reference voltage, a precision timing mechanism and a digital or balancing-type dc voltmeter. By this method an output voltage (e_o) measurement is obtained after a known input voltage (e_i) has been applied to the integrator for a controlled period of time (t). The time constant (RC) then is determined from the integration equation:

$$\frac{e_o}{e_i} = -\frac{t}{RC}$$

The method shown in Fig. 2 provides a simpler and less costly approach to the problem.

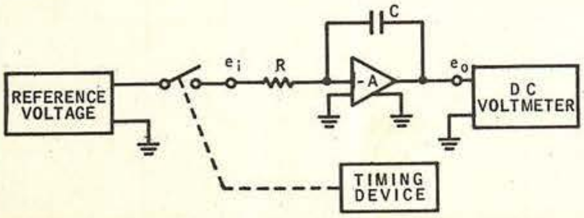


Fig. 1. Conventional measuring technique requires high-accuracy instrumentation.

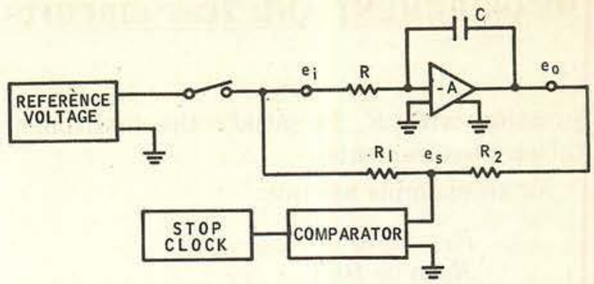


Fig. 2. Stop-clock-comparator gives voltage ratio without determining actual voltage.

Precision resistors R_1 and R_2 are chosen to provide a summing voltage (e_s) equal to zero when the output voltage is some desired multiple of the input voltage. The summing point is connected to the input of the dc voltage comparator which, in turn, operates to stop an electric stop clock (or electronic counter) when the summing voltage is zero.

Application of the input voltage will start the clock. When $e_o/e_i = R/R_1$, the summing voltage will be zero and the clock will stop. Thus, the elapsed time t is found, the voltage ratio e_o/e_i is known (without knowing the actual voltages) and the time constant may be determined from the integration equation as before.

Richard F. Cutler, project engineer, Sperry Utah Co., Salt Lake City, Utah.

Reducing Effect of Intermittent Instrumentation Loads

Five identical sensor elements are used in a control system. However, the summing network equivalent impedances are different, necessitating the addition of resistor R_c to provide a constant 10-K load to the sensor secondary, as in Fig. 1a. When selected on this basis, the intermittent connection of the instrumentation load reduces the equivalent impedance to $10 \cdot 105/115 = 9.13$ K. This change in impedance results in a 1-v drop in the sensor output voltage. This creates difficulties in calibration and acceptance.

The problem was solved by using the circuit in Fig. 1b. Here, R_1 and R_2 are selected so that the sum of their values compensates R_L to the desired 10 K equivalent. Resistor R_1 is made larger than R_2 so that the effect of connecting R_g is minimized. The final ratio of R_1 to R_2 must be compromised to provide sufficient voltage at their junction with R_g to satisfy the instrumentation requirements.

As an example assume:

$R_L = 18$ K
R_e to be 10 K

then

$$R_1 + R_c = 22.5 \text{ K} \left(10 = \frac{18R_c}{18 + R_c}\right)$$

let

$R_1 = 16$ K
$R_2 = 6.5$ K

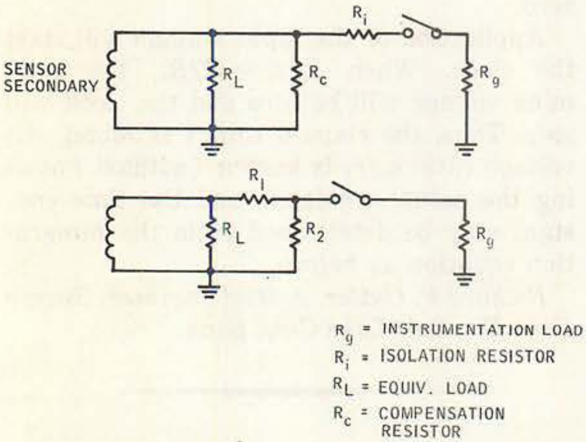


Fig. 1. Modification of sensor circuit reduces effect of loading.

with R_g connected

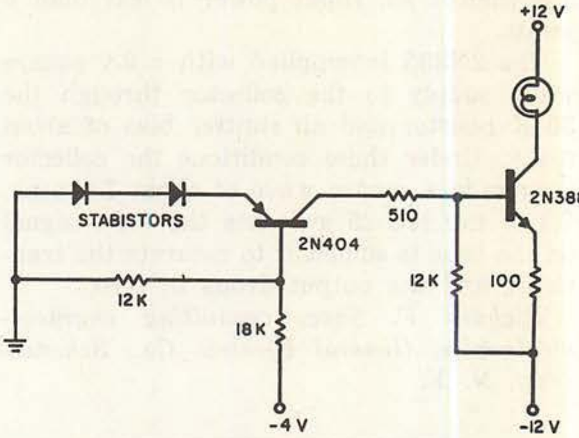
$$R_c' = 16 + \frac{6.5(30)}{36.5} = 21.35$$

$$\text{Required: } \frac{18(21.35)}{39.35} = 9.75 \text{ K}$$

R. L. Stoval, project engineer, The Bendix Corp., North Hollywood, Calif.

DC Presence Indicator Checks Three Voltages

The circuit shown here will indicate presence of all three voltages in a typical pnp digital system with negative collector and clamp supplies, and a positive bias supply. An npn system could use the complementary version of the circuit. Failure of any one of the voltages will extinguish the "dc-presence" lamp.



All three voltages must be present in modified AND gate to have bulb light.

The circuit is a modification of the typical transistor AND circuit. The conventional circuit will not indicate properly because of voltage incompatibility. The pnp-npn AND circuit shown in the diagram uses the three voltages as inputs.

The pnp-npn circuit detects in a true AND function. Absence of any one voltage will extinguish the lamp. The stabistors provide a tight control over the error level of the clamp voltage. Resistor R_4 limits the initial current surge present during turn-on of a cold bulb.

Daniel Chin, development engineer, Computer Control Co., Inc., Framingham, Mass.

Band Switcher Uses Light-Sensitive Resistors

We have devised a bandswitching circuit using light-variable resistors that has been successful in a number of multirange, low-power oscillating devices. The circuitry is simple and rugged. And, it eliminates a costly rotary switch that has low rf leakage, because our switch does not carry any rf at all.

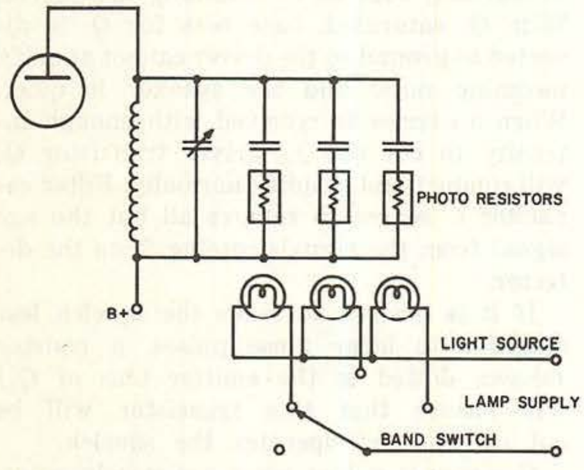


Fig. 1. Bandswitching is accomplished by shining light on photosensitive resistors and shorting out sections of the oscillator coil.

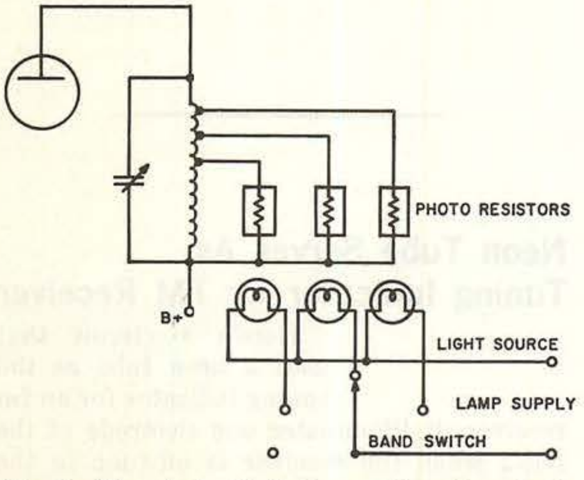


Fig. 2. Alternate bandswitching configuration also is feasible.

Light-variable resistors, connected across successive portions of an rf inductor, Fig. 1, are used for the switching. Illuminating any of the photoresistors with an essentially saturating light source effectively shorts out the portion of the coil bridged by the resistor. This changes the tuning characteristics of the LC circuit.

This scheme could be the solution to a multitude of problems associated with receiver and oscillator design, and in remote-control and telemetering applications.

An alternate configuration is shown in Fig. 2.

R. M. Zilberstein, project engineer, Andersen Laboratories, Inc., West Hartford, Conn.

Forward Biased Diode Gives Protection for Crystal Filters

Electrical characteristics of crystal filters may be permanently damaged if large input voltages occur, even for short intervals. In some receivers, for example, the amplifiers before the crystal filter are controlled by a relatively slow-acting agc circuit, and it is possible for very large signals to be momentarily impressed upon the filter input. The simple circuit shown in Fig. 1 effectively prevents large signals from reaching the filter.

Resistors R_2 and R_3 are for dc path and termination purposes, respectively. A path is provided through R_1 for a constant current to bias CR_1 in the forward direction. The maximum peak-to-peak voltage that may exist across the filter input is determined primarily by the dc current through R_1 multiplied by the parallel combination of the crystal input Z and R_2. For normal signal amplitudes CR_1 is always conducting and transfers all of the signal to the filter. When a large input signal exists, C acquires a charge as shown and biases CR_1 in the reverse direction so that it conducts for only a portion of each cycle, limiting the crystal

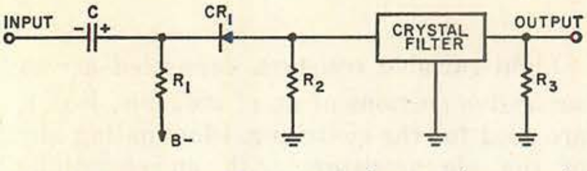


Fig. 1. Crystal-protection circuit shunts voltage peaks through R_1.

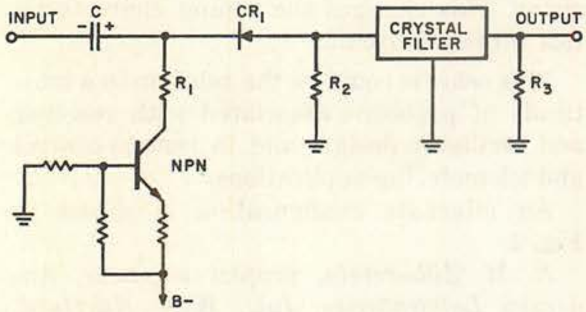


Fig. 2. Use of transistor with R_1 maintains constant current.

filter input. The polarity acquired by C forces a larger current through R_1, which should be reasonably constant for good protection. The B− voltage may be increased to provide a more constant current or R_1 may be replaced in part by a transistor, as shown in Fig. 2.

These circuits also may be used for amplitude limiting.

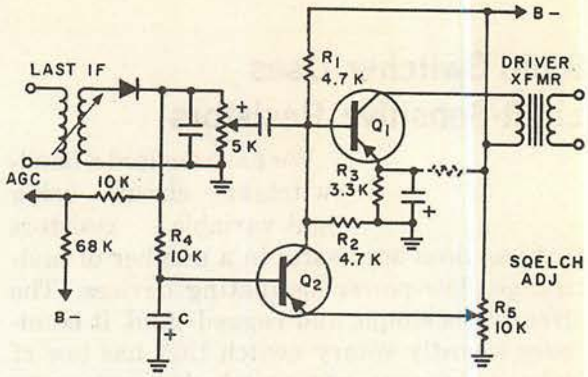
M. F. Feller, engineer, General Dynamics Corp., San Diego, Calif.

Computer Circuit Provides Simple Transistor Squelch

Adapting transistor-computer circuits to radio work can provide simple control circuits capable of handling complex jobs. A good illustration is the transistorized squelch circuit shown in the diagram.

This simple, but positively operating circuit calls for minor component value changes in the audio circuit and only four additional components, R_4, C, R_5 and the blocking transistor, Q_2. In a short-wave transistor radio, the circuit will make the set function as a monitor receiver; or, inserted into a inexpensive citizens band transceiver, the circuit will up-grade its performance.

Without an input signal, normal IF amplifier forward bias flows in the agc bus. A



Switching circuit is adapted to audio squelch system.

portion of this bias voltage is applied to the base of Q_2, biasing this transistor into full conduction, with R_5 determining the degree. With Q_2 saturated, base bias for Q_1 is diverted to ground so the driver cannot amplify incoming noise and the speaker is quiet. When a carrier is received with enough intensity to cut off Q_2, driver transistor Q_1 will conduct and amplify normally. Filter capacitor C serves to remove all but the agc signal from the signals coming from the detector.

If it is desired to make the squelch less sensitive to large noise pulses, a resistor (shown dotted to the emitter bias of Q_1) will ensure that this transistor will be cut off until rf operates the squelch.

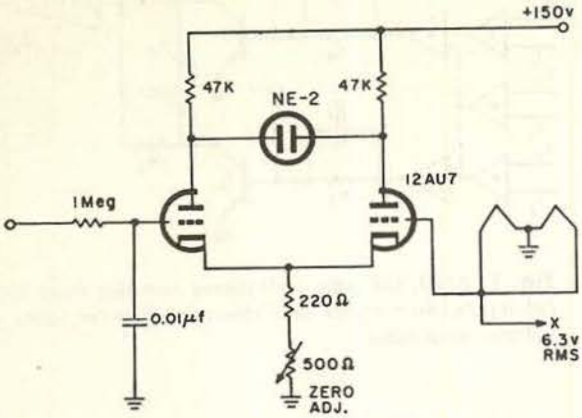
Component values given are merely representative and should be arrived at by cut-and-try methods in each case.

Leonard E. Geisler, senior applications engineer, Apollo Industries, Ltd., Tokyo, Japan.

Neon Tube Serves As Tuning Indicator for FM Receiver

Here's a circuit that uses a neon tube as the tuning indicator for an fm receiver. It illuminates one electrode of the NE-2 when the receiver is off-tune in the high direction, and the other electrode when off-tune in the low.

On-tune (zero volts input), the 60 cps part of the plate voltage keeps both electrodes



Neon-tube fm tuning indicator can be added to circuit of fm receiver.

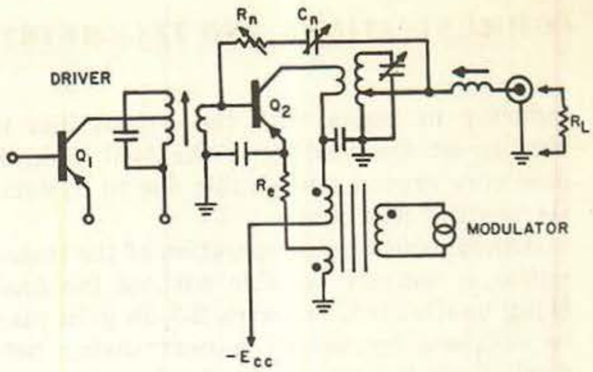
fired, so the tuning transition is smooth. A ± 1 v swing from the discriminator is ample for clear indication. This compares favorably with the performance of the GAL7 indicator tube.

M. W. Egerton, Jr., engineer, Towson Laboratories, Towson, Md.

Linear Modulation Of Transistor Power Amplifiers

Separate modulation of the collectors of both the driver and the final of am transistor power amplifiers has been recommended as the only way to obtain 100 per cent modulation without excessive distortion.

The approach described here yields highly satisfactory results with only the final stage modulated. The basis of the system is simultaneous modulation of both the collector and emitter by application of audio signals that are 180 deg out of phase. The base drive to the final is not varied. As the result, demodulated output of a transmitter using this system has a characteristic distortion-free, "punchy" quality, which greatly extends the range of operation by comparison with simple collector modulation. The circuit of a typical class C final is shown in the diagram.



High level modulation of transmitter final is obtained by applying modulation to both emitter and collector.

When modulation is applied in series with the collector supply of a class-C transistor final, the height of the collector-current pulses varies within the transistor limits. If a lower amplitude of modulation, shifted by 180 deg, is simultaneously fed in series with the emitter return, the transistor operating point will be moved along the h_{FE} curve and the power content of the collector current pulses will be greatly increased on modulating wave crests.

Since driver loading by the final decreases during troughs of the modulating waveform, the drive tends to rise just enough to prevent downward carrier modulation on large "negative" modulation peaks.

Oscilloscope inspection of the driver waveform shows no appreciable variation at any normal level of modulation.

It was found, pragmatically, that overcoupling between driver and final must be avoided, or the final will depart from true class-C operation with consequent inefficiency and distortion.

The value of R_e is arrived at by experiment. The emitter resistor is varied while the final modulated output waveform is observed on an oscilloscope. When a perfect trapezoid is obtained at maximum modulated rf output, the value of R_e may be permanently set, using a fixed 10 per cent tolerance resistor.

Although transistor parameter spreads vary considerably, this system is extremely tolerant, hence economically feasable for use on the production line for applications with transistorized radio-telephone equipments.

In practice, we found it necessary, due to the very low output impedance of transistors operated at low values of E_{cc}, to link-couple the transistor to the tank coil. We thus were able to preserve a good operating Q of between 30 to 45. A brass-slug tuned inductor in series with the output line is used to set fine loading of the final tank. A dust core proved undesirable due to hysterisis heating and losses.

Although servicable operation of the transmitter is entirely possible without the final being neutralized, an extra 2-3-db gain may be obtained by use of a neutralizing network. Such a network also tends to improve over-all performance.

Neutralization must be adjusted while the final is modulated 100 per cent by a 1-Kc sine wave and output is monitored by a cathode-ray oscilloscope connected directly across the output terminals with the dummy load. A perfectly clean trapezoid is indication of correct operation of the final.

Design of a transmitter final, using this modulation system, closely follows vacuum-tube practice except for impedance differances. The modulation transformer impedances are determined in the usual vacuum-tube fashion.

It is possible to use this same system with grounded-base finals if the designer allows for varying loading on the driver due to the modulation's effect on the final transistors' emitter.

Leonard E. Geisler, senior applications engineer, Apollo Industries, Ltd., Meguro, Tokyo.

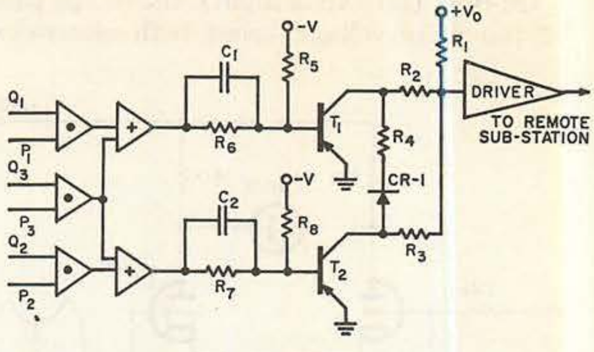


Fig. 1. AND/OR gate multiplexer samples three logical inputs, then codes and transmits them in terms of voltage amplitude.

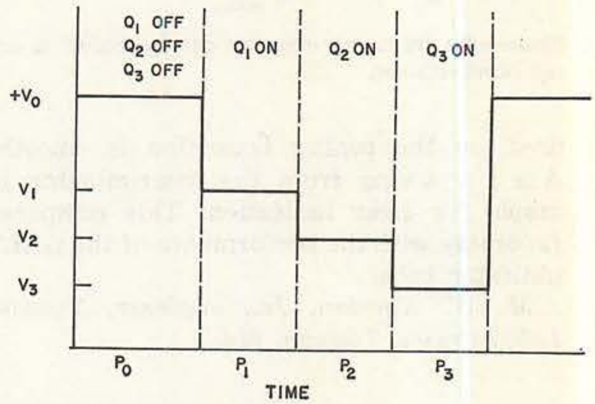


Fig. 2. Output voltage vs logical input for sampling times *P1*, *P2* and *P3*.

AND/OR Gate Multiplexer Uses Voltage-Amplitude Coding

The circuit in Fig. 1 codes the intelligence from three logical signals for transmission through a long, single conductor to a remote station, where the signals are decoded and used as required. The coding is done in terms of voltage amplitude.

Fig. 2 shows the output voltage vs control-signal timing.

If *Q1* is true at time *P1*, transistor *T1* saturates, and the output voltage V_1 is calculated by:

$$V_1 = V_0 \left[\frac{R_2(R_3+R_4)}{R_1R_2+R_1R_3+R_1R_4+R_2R_3+R_2R_4} \right] \quad (1)$$

Similarly, if *Q2* is true at time *P2*, transistor *T2* saturates and V_2 is:

$$V_2 = V_0 \left[\frac{R_3}{R_1+R_3} \right] \quad (2)$$

If *Q3* is true at time *P3*, both *T1* and *T2* saturate and V_3 is:

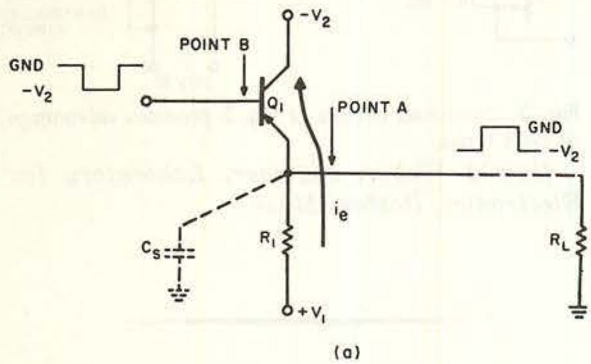
$$V_3 = V_0 \left[\frac{R_2R_3}{R_1(R_2+R_3)\ +\ R_2R_3} \right] \quad (3)$$

Germanium diode *CR-1* and resistor *R4* conduct only when *T1* is saturated and *T2* is cut off. Their purpose is to reduce the voltage of the first step at *P1*, so that changes between voltage steps are equal.

Jack McGruder, electrical engineer, Hughes Aircraft Co., Fullerton, Calif.

Reducing Power Dissipation In Emitter-Follower Circuits

The emitter-follower circuit shown in Fig. 1a is used to send digital information over a transmission line into a load represented by R_L. In this circuit it is desirable to have i_e as large as possible (and, therefore, have R_1 as small as possible) in order to (1) reduce the rise time—this time



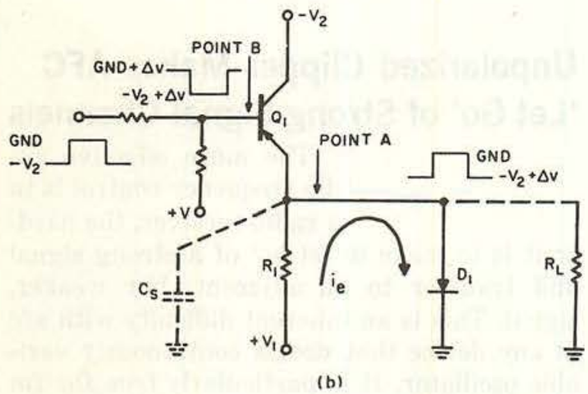


Fig. 1. Basic emitter follower (a) is modified by adding diode (b) to reduce current through transistor.

is a function of R_1 and the stray capacitance C_s, (2) provide for impedance matching between the load and source, and (3) reduce any noise which could feed back from the line through Q_1. When the input is at the $-V_2$ potential, the power dissipation in Q_1 is small because the voltage drop across Q_1 is negligible; but when the input is at ground potential the power dissipation equals i_eV_2. Thus the maximum power dissipation rating of Q_1 can be the limiting factor for i_e.

To increase i_e and still not exceed the maximum power-dissipation rating of Q_1 the circuit shown in Fig. 1b is recommended. Here a biasing network inserted at point B, and a clamping diode inserted at point *A* is used to modify the original circuit. When the input signal goes to ground, point *B* is biased slightly positive, the emitter-to-base junction of Q_1 is then back-biased because point *A* is clamped to ground through D_1, and Q_1 cannot conduct. Thus the previous power-dissipation problem is eliminated because the current through Q_1 is zero, and a transistor having a much smaller power dissipation rating can be used. The current, i_e, which previously went through Q_1 to the collector now goes through D_1 to ground.

Norton Markin, engineer, Sherman Oaks, Calif.

Current Overload Protectors Use Simplified Designs

Many an engineer's face has turned as red as the anode of the power tube he was working with, when he glanced at the plate-current meter and observed that it was indicating much greater current than normal.

Although most modern tubes will not be severely damaged by a momentary overload, the problem is a constant threat to the engineer of high power equipment using expensive tubes and components.

Many commercial communications and broadcasting transmitters and other industrial equipment include automatic overload protection. These overload circuits are frequently quite complex.

In the simple protective device shown in Fig. 1, a warning light is turned on if excess

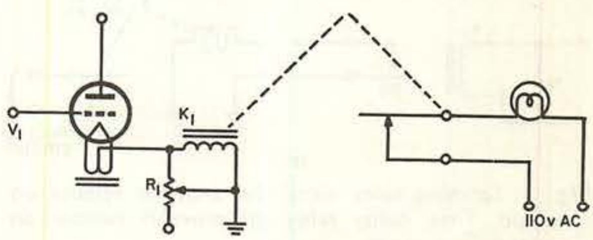


Fig. 1. Basic circuit provides indication of tube overload.

current is drawn. The circuit requires only the addition of a paralleled relay and potentiometer (K_1 and R_1) in series with the filament or cathode return of the power tube. The potentiometer's value must be such that when paralleled with K_1, the voltage drop of the combination is equal to the pull-in voltage of the relay at the maximum current of the circuit. This can be computed by:

$$R_1 = \frac{E_{K1}}{I_{max} - \left(\frac{E_{K1}}{R_{K1}}\right)}$$

where: E_{K1} = operating voltage of relay K_1.
R_{K1} = dc resistance of K_1
I_{max} = maximum circuit current in amps.

For example, if a 24-v, 200-ohm relay is used, R_1 is 300 ohms to limit the maximum current to 0.2 amp. Since the pull-in voltage of a relay is lower than the operating voltage, R_1 is made adjustable and can be set to adjust the sensitivity of the relay. If the pull-in voltage of the relay is known, the pull-in voltage may be substituted for E_{K1} in the formula. The resistance of a relay and, therefore, its sensitivity change significantly with heat. Pull-in current should be set after the relay has reached operating temperature. Wattage rating of the resistor may be computed from:

$$P = E_{K1}\left(I_{max} - \frac{E_{K1}}{R_{K1}}\right).$$

More sophisticated control circuits can be derived from this basic circuit configuration. Fig. 2 demonstrates two possibilities. In Fig.

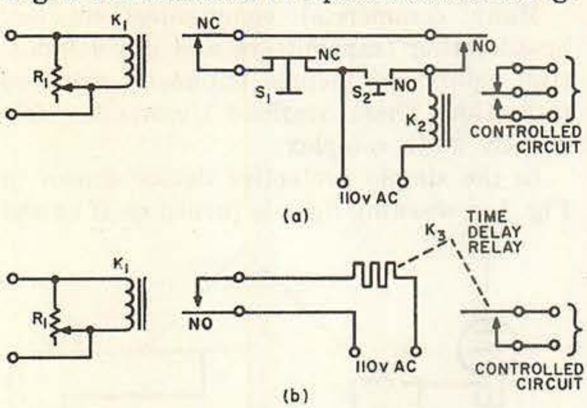


Fig. 2. Latching relay circuit (a) provides release on overload. Time delay relay (b) prevents release on momentary overloads.

2a, when an overload occurs, latching relay K_2 opens due to the breaking of the latch contact circuit by the normally closed contacts of K_1. In the circuit shown, pushbotton S_2 latches and S_1 releases the circuit.

Fig. 2b demonstrates a method for overcoming the releasing of K_2 on transients or momentary overloads, as would be the case with Fig. 2a. A short-cycle thermal relay is employed. A delay of 3 to 10 sec normally would be adequate. The delay should equal the longest anticipated normal momentary overload, such as is encountered when returning a transmitter tank circuit. A further sophistication may be achieved by consolidating the circuits of Fig. 2 into the circuit shown in Fig. 3.

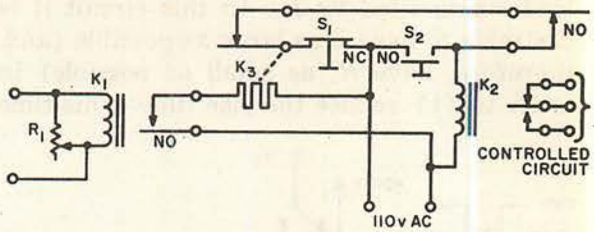


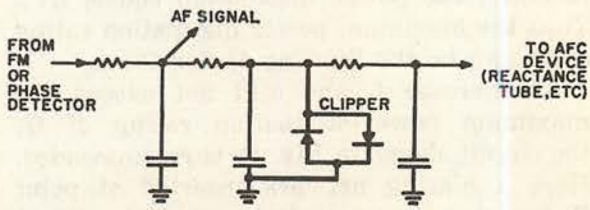
Fig. 3. Combined circuits of Fig. 2 provides advantage of both types.

Harold Weber, engineer, Laboratory for Electronics, Boston, Mass.

Unpolarized Clipper Makes AFC 'Let Go' of Strong Signal Channels

The more effective audio frequency control is in a radio receiver, the harder it is to make it 'let go' of a strong signal and transfer to an adjacent, but weaker, signal. This is an inherent difficulty with afc in any device that uses a continuously variable oscillator. It is particularly true for fm receivers that have broadband IF and derive the afc bias from the fm detector. With full afc action it may be impossible to tune in weaker channels without first disabling the afc.

Of several possible methods for disabling afc, adding an unpolarized dc clipper, as shown, proved to be the simplest. It can be added to any afc without disturbing the original circuit and without adding tubes or tuned circuits.



Clipper added to circuit prevents development of large afc bias voltages, which can swamp out weak signals.

The clipper can consist of two very low voltage Zener diodes or, each "diode" can consist of 7 to 9 small selenium plates in series. (The latter arrangement uses the forward knee of the rectifiers.)

At or near center tuning the full gain of the afc circuit is available for holding on to weak signals. When tuning away from strong signals, the clippers prevent large afc bias from being developed. They permit the signal to be released so that weak adjacent channels can be received.

In the Craftsmen C500 receiver in which the clipper was installed, it was found that if it limited maximum afc bias to 1.5-2.5 v, the results were excellent. (Without the clipper, strong signals produced as much as 20-25 v maximum bias, completely swamping out nearby weaker channels.)

Louis W. Reinken, Reinken Rectifier Engineering, Plainfield, N. J.

Transistor Squelch Circuit Uses Minimum of Components

A simple and inexpensive squelch circuit is designed for use with transistorized communications receivers is shown in the illustration. With no signal coming into the receiver, an agc-controlled IF amplifier, Q_1, is operating at its maximum collector current and gain.

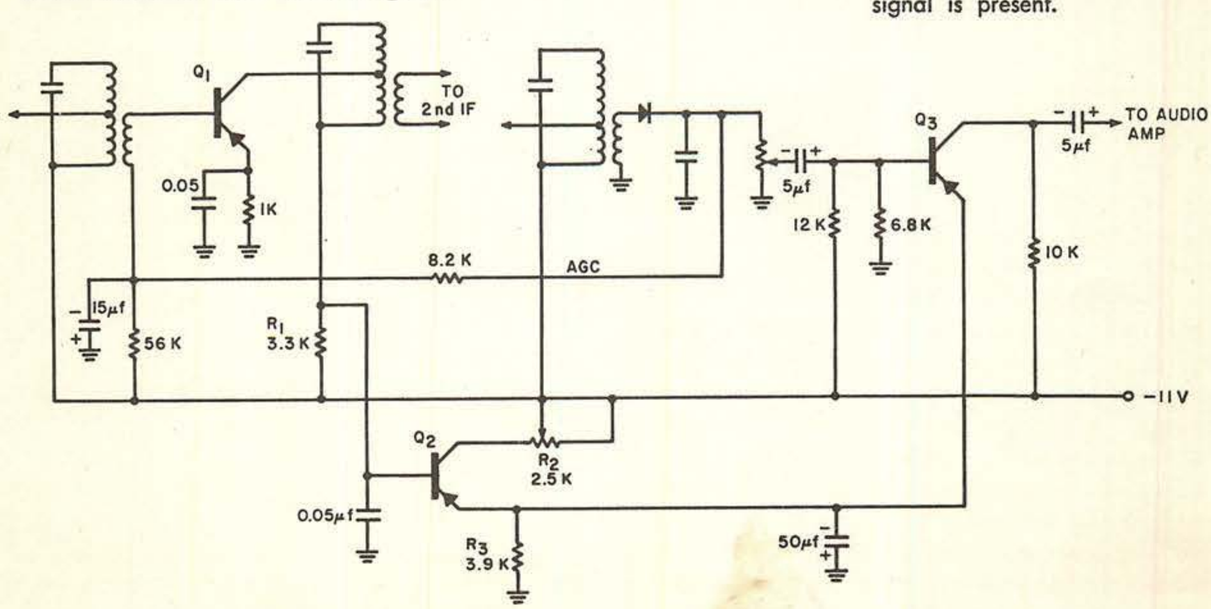
The voltage drop across R_1 is also maximum under this condition and is used to bias Q_2 on. Current through Q_2 is determined by the value set on the variable resistor R_2. If R_2 is set so that the current flow in Q_2 causes the voltage drop across R_3 (which is also the emitter resistor of Q_3) to be greater than the voltage from base-to-ground minus V_{be} of Q_3, then Q_3 is turned off. Under these conditions, there is no output from Q_3 and the receiver is quiet.

Now, assume that a signal is being picked up by the receiver. Under this condition, when the agc is acting, the current through Q_1 is reduced by an amount depending on the signal strength. When this occurs, the voltage drop across R_1 is reduced, thus reducing the collector current through Q_2. This produces a lower voltage drop across R_3. When this voltage becomes less than the voltage from base-to-ground minus V_{be}, Q_3 begins to turn on; this results in an output to the audio amplifier.

This circuit has been incorporated into citizens-band receivers with good results. It was found to be capable of squelching out up to a 300-μv signal into the receiver and still maintain good control down to something less than 1 μv.

It should be noted that between full on and full off conditions of Q_3, small amounts of collector current may result in some audio distortion depending on the volume-control setting. This distortion should not be ob-

Squelch-control transistor, Q_2, cuts off audio when no input signal is present.



jectional and under ordinary operating conditions (where the signal received is many times greater than that required to just trigger the squelch), this distortion is not present.

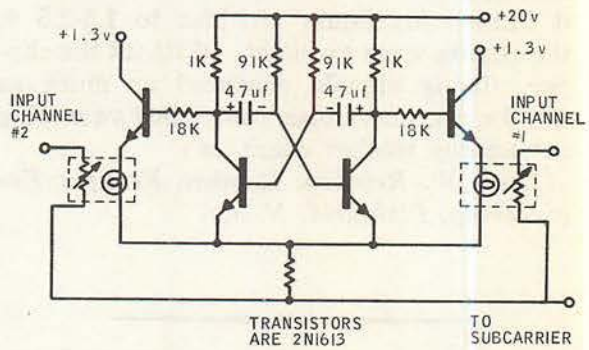
When the squelch pot R_2 is set to just squelch out any signal being received, approximately 3-db increase in signal level will completely override the squelch action. The amount of increase is dependent, of course, on the level of the original signal, the gain and sensitivity of the receiver and the amount of agc of the receiver.

The IF transistor was set to idle approximately 0.75 ma under no-signal conditions. The squelch transistor may be any small-signal npn unit, such as the 2N1304; the audio transistor may be any small-signal pnp alloy, such as the 2N1274.

Richard E. Morgan, Transistor Products Div., Texas Instruments Inc., Dallas, Tex.

Photoconductor Commutator Has Simplicity and Isolation

We needed an electronic, two-channel, low-rate commutator to be used with an fm-fm telemetry system. Such a commutator was devised using the Raytheon CK1114 Raysistor. This is a photoconductive cell plus a grain-of-wheat lamp in a TO-5 case. By switching the lamp full on and off, the resistance of the photoconductor may be switched from 0.5 ohm to 300 megohms. Connecting two CK1114s to the input of a subcarrier oscillator and alternately switching them, we had a two-channel commutator with 300 megohms isolation between input channels to the telemetry system.



Electrical isolation in telemetry circuit is obtained by using light-controlled photoconductor.

The circuit is an astable multivibrator driving two transistor switches, which switch the lamps in the Raysistors alternately. Since the photoconductors are electrically independent of the lamps, the signal channels are electrically isolated from the switching circuit. Dc input signals may be of either polarity. The passband is flat from dc to 1 mc.

C. G. Blanc, instrumentation engineer, R. C. Crawford, electrical technician, U.S.N.O.Y.S., China Lake, Calif.

Flip-Flop Relay Driver Eliminates Emitter Follower

The conventional method of driving a solenoid or relay from a control flip-flop is to drive the transistor switch with an emitter follower. The follower, in turn, is

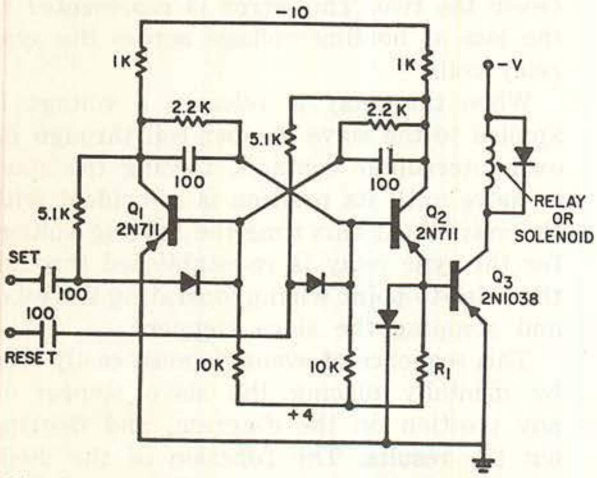


Fig. 1. Emitter follower is eliminated by putting transistor switch Q_3 in the emitter circuit of driver.

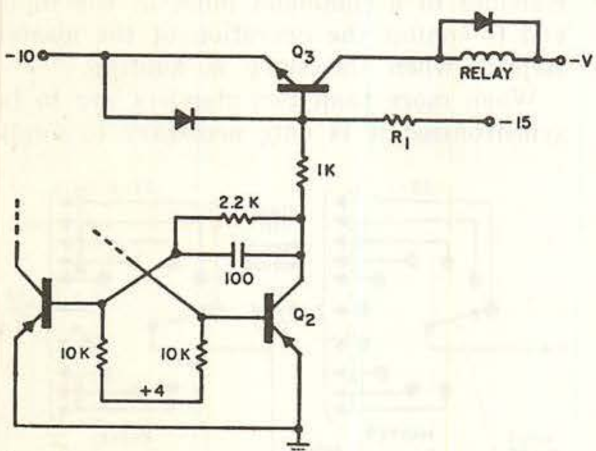


Fig. 2. Alternate switch driven method places switching transistor in collector circuit of flip-flop transistor.

driven by the flip-flop.

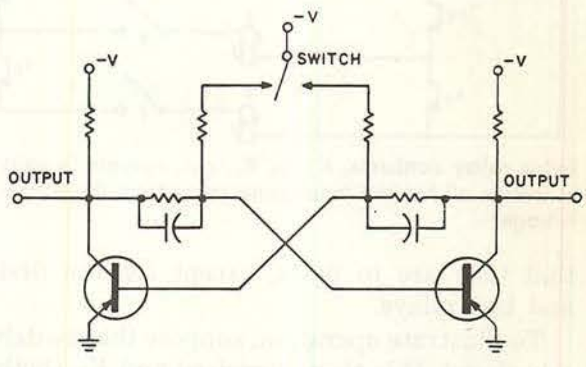
By putting the switch directly in the emitter circuit of one of the flip-flop transistors, the emitter follower can be eliminated. In Fig. 1, Q_2 draws virtually all of its emitter current through the base of Q_3. Resistor R_1 supplies I_{co} to the base of Q_3, and diode CR_1 clamps the emitter of Q_2 when Q_1 is conducting.

An alternative method is shown in Fig. 2, using an npn transistor in series with the collector of Q_2. This, however, is slightly less efficient than the first circuit, because R_1 must supply I_{co} and base current for Q_1 when Q_2 is cutoff.

Phillip Joujon-Roche, engineer, Aeronautics, Newport Beach, Calif.

Multi-Buffered Switch Eliminates Contact Bounce

In circuits using toggle or push-button switches to generate digital logic levels or trigger signals, contact bounce may cause objectional noise voltages. However, this noise can be eliminated by letting the switch



Problem of contact bounce in switch-generated trigger signals is eliminated by using the switch to flip a bistable multi, then using the multi's outputs.

trigger an ordinary resistance-coupled bistable multivibrator as shown in the figure. The switch is a double-throw type. As it is thrown from one contact to the other, the complementary outputs of the multi provide fast, simultaneous switching, free of contact noise.

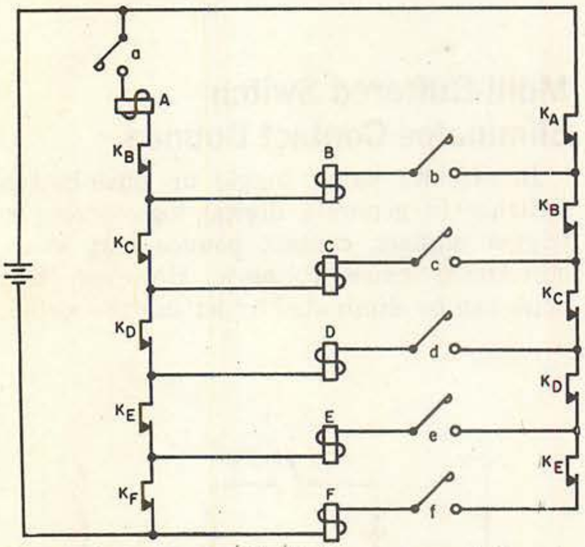
Stewart T. Coffin, Head of Engineering, Dynamic Controls Co., Cambridge, Mass.

Relay Bridging Avoided Through Isolation Scheme

This circuit will prevent bridging or erroneous closing of relay contacts in a multiplexer, even when the drive switches have considerable chatter and could accidentally make contact. The circuit applies to any number of relays. Only two extra contacts per relay are required.

When any switch energizes a specific relay, no other coil can be energized until the contacts of the first relay change state.

In Fig. 1, the relay coils are designated *A, B, C*, etc., and the corresponding switches that control them are labelled *a, b, c*, etc. The additional, normally closed contacts that provide circuit isolation are K_a, K_b, K_c. . .; note



Extra relay contacts, K_a, K_b, K_c, etc., operate in pairs to isolate all but the functioning relay from the supply voltage.

that they are in pairs, except for the first and last relays.

To illustrate operation, suppose that switch *d* is closed. *D* is then energized and K_D (both poles) opens. The subsequent closing of any combination of drive switches *a, b, c*. . .etc. cannot affect any other relay. Should the relays be of the make-before-break type, the shorting time is so much shorter than the closing delay that bridging still cannot occur.

The scheme may be applied to similar situations in solid-state logic gating.

Patrick F. Howden, project engineer, Thompson-Ramo-Wooldridge Computers Co., Canoga Park, Calif.

Stepping Switch Provides Fool-Proof Synchronization

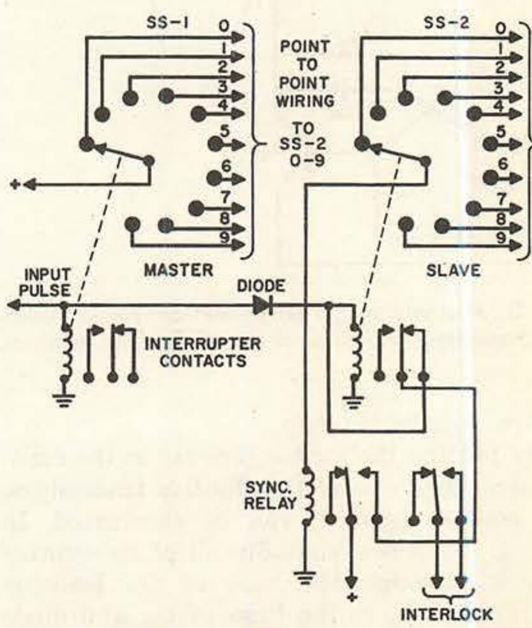
In the circuit configuration shown here loss of synchronization is virtually impossible. There is no way for the master and slave steppers to get out of step except by component failure.

The action of the sync relay is to detect, through the continuity of the point-to-point wiring, from the master to the slave steppers, an error in relative position between the two. This error is represented as the loss of holding voltage across the sync relay coil.

When the relay is released a voltage is applied to the slave stepper coil through its own interrupter contacts, forcing the slave to move until its position is coincident with the master. At this time the holding voltage for the sync relay is re-established through the point-to-point wiring, operating the relay and stopping the slave stepper.

This sequence of events is most easily seen by mentally placing the slave stepper in any position on the diagram, and figuring out the results. The function of the diode between the two stepper coils is to permit simultaneous operation of the steppers in response to a command pulse at the input and to inhibit the operation of the master stepper when the slave is hunting.

When more than two steppers are to be synchronized it is only necessary to dupli-



Loss of synchronization of stepping switches energizes slave stepper until proper position is found.

cate relay point-to-point wiring and supply a sync relay for each additional stepper. The contacts on the sync relay labeled "interlock" may be utilized to disable other circuits when the slave stepper is hunting. The only requirement for the sync relay is that it must stay closed between steps and must release fast enough to stop the slave units in the correct position.

Frank L. Egenstafer, project engineer, Jerrold Electronic, Hatboro, Pa.

'Constant-Current' Supply Interlocks Relay Network

Interlock circuits become complex when a large number of relays are to be remote-controlled by push-buttons, or where latching of a new relay must unlatch any previously energized relay.

The circuit of Fig. 1 permits these functions with only a single form A contact on each relay for the latching function. Additionally, the remote push-buttons require only a single wire plus common for each relay to be controlled.

In Fig. 1, a quasi-constant current source is obtained by starting with a battery voltage of at least four times the desired pull-in voltage for the relays, and inserting a resistor in series.

The low side of each relay coil goes to ground through a Zener diode rated at the pull-in voltage. This removes voltage from any previously latched coils when a new one is engaged. If reed type, or other fast relays are used, the capacitors shown across the push buttons are not required. With relatively slow drop-out on the relays, a previously engaged coil may not drop out if the button is tapped only momentarily. A firm touch on the button, shunt capacitors, or faster relays will cure the problem.

The power waste in the large series resist usually is not serious since only one relay

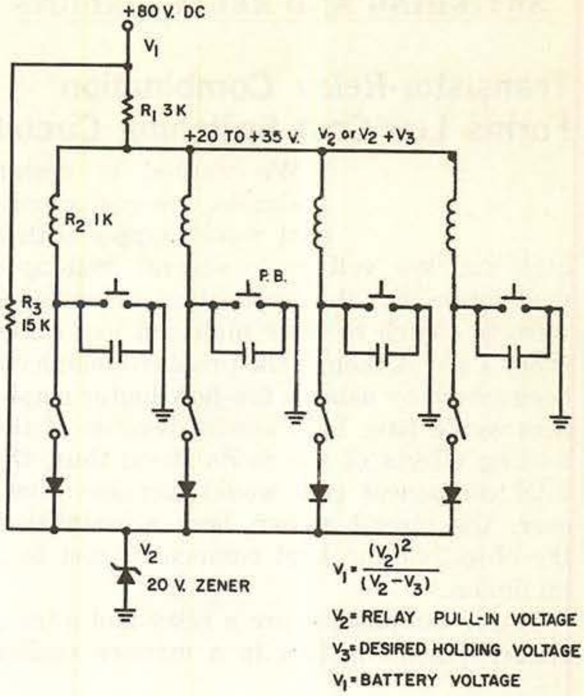


Fig. 1. Interlocks relay control requires only a single conductor for each relay.

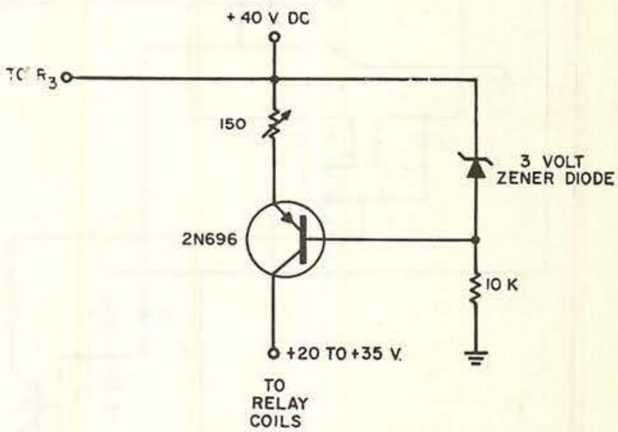


Fig. 2. Transistor regulator reduces power drain on constant-current source.

is energized at a time and the power involved is rather small.

Power loss can be reduced by using a more efficient constant-current source. This could be a fast barretter, a barretter and choke coil in series, or a transistor circuit, as in Fig. 2. The series diodes in the low side of each relay coil prevents the grounding of the Zener diode.

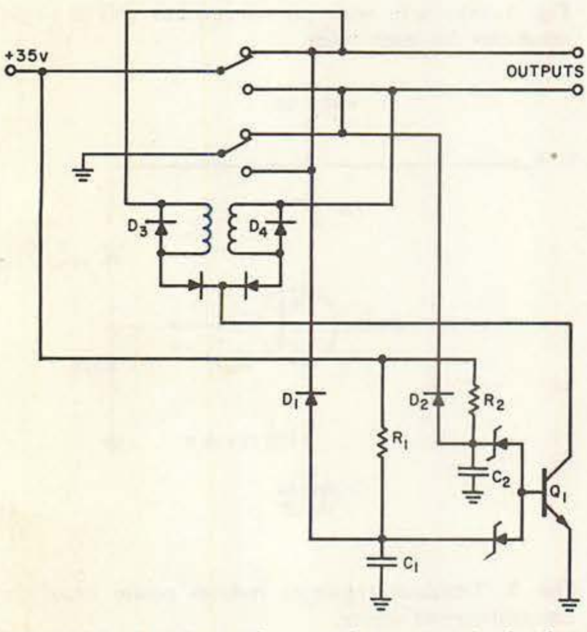
Daniel Cronin, executive vice president Bell Sound Studios, New York, N. Y.

Transistor-Relay Combination Forms Low-Cost Switching Circuit

We wanted to design a simple, low-cost circuit that would supply both a high and low voltage to several dual-input modulators. For this application, the voltages were to switch to their high and low states every 4 sec. Although the problem could have been solved by using a flip-flop, buffer amplifiers would have been needed because of the loading effects of the modulators; thus, the total component cost would increase. However, the circuit shown here accomplished the objective and kept component cost to a minimum.

Major components are a relay and a transistor, which function in a manner similar to a multivibrator due to the R-C networks at the base of the transistor. The junction of R_1C_1 (and R_2C_2) is alternately connected to ground through the relay contacts. If the junction of R_1C_1 is disconnected from ground, C_1 will begin to change to 35 v through resistor R_1 when power is applied to the circuit. When the charge on C_1 is +15 v dc, the Zener diode breaks down and transistor Q_1 turns on. When Q_1 is on, the relay is energized, the R_1C_1 junction becomes grounded, C_1 discharges, and transistor Q_1 turns off. Capacitor C_2 now can begin to charge and the cycle is repeated.

Michael Cianciola, V. Lemley, engineers, General Precision Inc., San Marcos, Calif.



Alternate charging of C_1 and C_2 provide flip-flop action.

Voltage-Controlled Relay Selector System

There are many applications requiring a relay system to work from a voltage-controlled device. Fig. 1 shows a circuit capable of closing three relays in 1, 2, 3 order as the input voltage is raised beyond 10, 12, and 15 v respectively. (The voltage levels are arbitrary so long as sufficient collector voltage is available to actuate the relays.)

The transistor base resistors (R_B) are, in effect, limiting resistors that absorb any input voltage in excess of that needed to turn on the relays. Their values determine the turn-on sensitivity of the relay circuit. The following formulas can be used to calculate the sensitivity ΔV, that is, the input voltage in excess of the Zener voltage required for relay turn-on:

$$V_{IN} = V_Z + \Delta V$$
$$V_{IN} \approx V_Z + R_B$$
$$R_B \approx \Delta V \mathrm{B}/I_C$$

where

V_{IN} = the minimum input voltage required for relay closing,

V_Z = breakdown voltage of Zener diode,

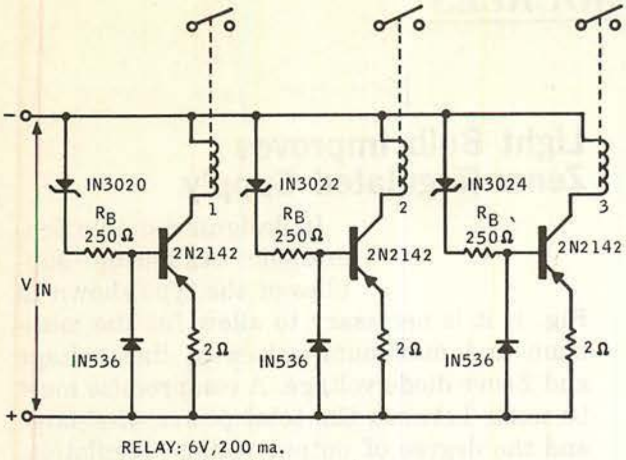
I_C = collector current required for relay closing,

B = dc current gain of the transistor, and

R_B = base resistance.

The 2-ohm resistor and the silicon rectifier in the transistor emitter and base circuits form a clamp that limits the current into the relays when the input voltage rises to the firing levels of succeeding relays.

The characteristics of the silicon rectifier



As V_{in} is increased beyond 10, 12 and 15 v, relays 1, 2 and 3 are actuated in sequences.

(1N536) are such that the unit does not conduct until the voltage reaches a level of about 0.7 to 0.8 v. Typically, the base-emitter voltage of the transistor in the forward-biased condition is about 0.3 v. To cause the rectifier to conduct, the voltage drop across the 2-ohm resistor must be at least 0.4 v, corresponding to an emitter current of about 200 ma.

As the input voltage is raised enough to turn on the last relay, the current through the earlier relays normally would increase, resulting in excessive relay current through these stages. This increase, however, is held to a minimum due to conduction of the rectifier which clamps the emitter current to a predetermined level.

As the number of relays in such a system increases, the power dissipation of the transistors, which are controlled by the lower voltage Zeners, increases because of the higher collector-emitter voltages, and heat sinking must be provided.

The circuit shown has the advantage of rapid turn-off when the input voltage falls below the Zener control voltage since all base drive is then eliminated. This provides positive turn-off as the input voltage drops below the initial turn-on level.

J. D. McCall, engineer, Motorola Semiconductor Products, Inc., Phoenix, Ariz.

Currector and Zener Diode Reduce Ripple In Power Supply

The constant current characteristic of a currector—a relatively new solid-state device—can be combined with the

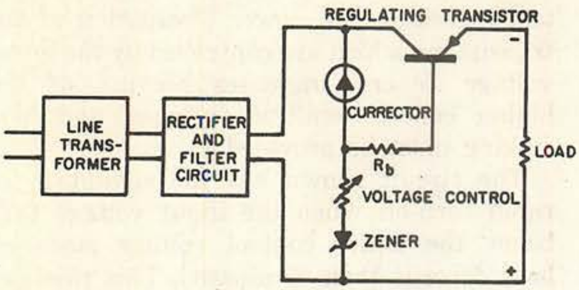


Fig. 1. Series connected currector-Zener reduces ripple in power supply.

constant-voltage characteristic of a Zener diode to give a simple power supply with very low ripple.

The ripple voltage appearing at the input to the base resistor is attenuated by the factor $\dfrac{R_z + R}{R_z + R + R_c}$

where:

R_z = dynamic resistance of the Zener, which is small.

R = voltage control resistor, which may be kept small.

R_c = dynamic resistance of the currector, which is very large.

Resistor R_b protects the transistor from excessive base current in case of a short cir-

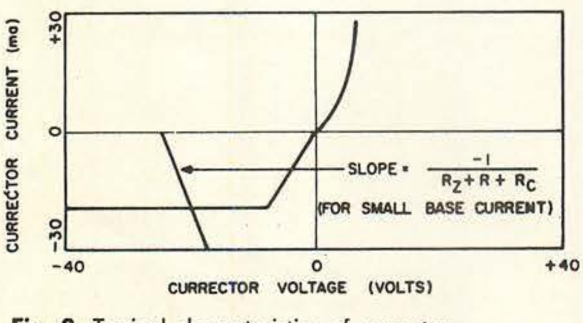


Fig. 2. Typical characteristics of currector.

cuit. This arrangement provides a simple, voltage controlable power supply with good regulation and low ripple.

Edward P. Mitchell, member of technical staff, Hughes Aircraft Co., Fullerton, Calif.

Light Bulb Improves Zener-Regulated Supply

In designing simple Zener-diode regulating supplies of the type shown in Fig. 1, it is necessary to allow for the minimum and maximum values of line voltage and Zener diode voltage. A compromise must be made between the total power dissipated and the degree of output voltage regulation.

The voltage regulation of the power supply shown in Fig. 1 can be improved by increasing the secondary voltage of the transformer. But this results in higher power

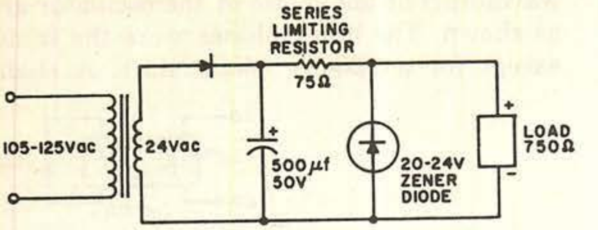
	75-ohm resistor	No. 1447 lamp
Minimum Zener Current	19 ma	20 ma
Maximum Zener Current	138 ma	57 ma
Minimum Zener Power	0.38 w	0.40 w
Maximum Zener Power	3.30 w	1.37 w
Maximum Lamp/Resistor Power	2.35 w	0.97 w
Maximum Total Power	5.65 w	2.34 w

dissipation in the series limiting resistor and requires higher voltage ratings for the capacitor and rectifier.

A significant improvement in over-all circuit performance can be achieved with no increase in cost by substituting a properly rated incandescent lamp for the series limiting resistor. The increase in the effective resistance of the lamp with applied voltage stabilizes the current through the Zener diode. For the case shown, the type 1447 lamp proved an ideal choice since it has a resistance of about 75 ohms under the minimum voltage conditions (minimum line voltage and maximum Zener voltage) and is within its nominal 18-v rating under maximum voltage conditions (maximum line voltage and minimum Zener voltage).

A comparison of the two cases is shown in the table. It is seen that the minimum Zener current is equal in both cases, as required. However, the total range of Zener current is three times larger with the resistor than with the lamp. The voltage regulation will, accordingly, be about three times better with the lamp than with the resistor.

With the lamp the total Zener power dissipation is low enough to permit a 3.5-w unit to be used conservatively. With the resistor, a 10-w unit probably would be required. Since the average operating lamp voltage is approximately 2/3 its rated value a longer than usual lamp life can be expected.



Circuit performance of this simple Zener-diode regulated power supply can be improved by substituting incandescent lamp for the series limiting resistor.

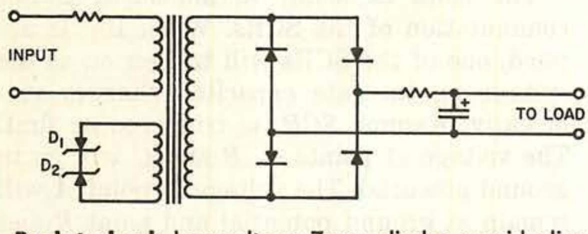
In addition to its function as a series regulating element the lamp also may serve as a pilot lamp or as a trouble-shooting aid.

T. P. Sylvan, application engineer, General Electric Co., Syracuse, N. Y.

Regulated Power Supply Uses Low-Cost Diodes

A simple circuit was required to provide a regulated, temperature-stable, high-voltage dc supply for a fixed load. Methods using vacuum tubes, voltage-regulator tubes or high-voltage Zener-diode strings are elaborate and expensive.

The circuit shown in the figure uses inexpensive, low-voltage Zener diodes with about 5 to 7-v breakdown voltages for 10-w dynamic impedance and low temperature coefficient. Zener diodes D_1, D_2 are placed back to back to operate on both halves of the input voltage.



Back-to-back low voltage Zener diodes provide line regulation.

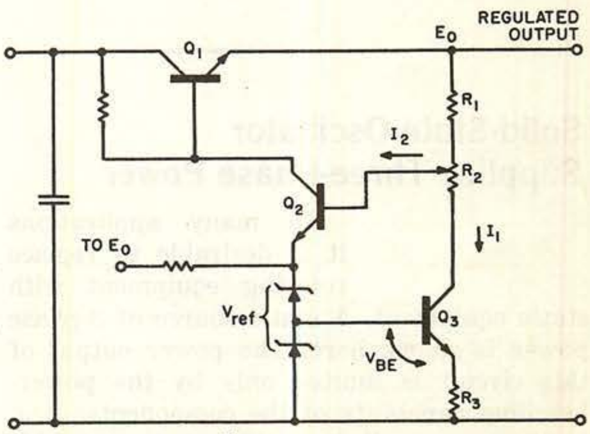
The Zener voltages are multipled by the turns ratio between the primary and low voltage winding. This effectively places a high voltage, double-ending Zener diode in parallel with the primary winding, thus regulating the input voltage to the transformer. The output on the secondary will be a clipped sine wave, which will ease filtering problems. If ultra-low temperature variation is desired, the negative temperature coefficient of the Zener diode in the forward direction can be cancelled out by picking a Zener diode with a positive temperature coefficient in the reverse direction. To insure a minimum of tilt in the output, a transformer with good low-frequency response should be used.

Joseph La Fiandra, project engineer, EDO Corp., Yonkers, N. Y.

Modified Regulator Is Made More Sensitive to Output Changes

In a conventional series regulator circuit, Fig. 1, a change in output voltage, ΔE_o, is seen at the base of Q_2 as $\Delta E_o/n$, where n is the voltage-division ratio determined by $R_1 - R_2$ and R_3.

If n were to approach unity, the regulator would be much more responsive to ΔE_o.



$$I_1 \approx \frac{\frac{V_{ref}}{2} - V_{BE}}{R_3}$$

Fig. 1. In conventional series regulator, charge in output voltage, ΔE_o, is seen at base of Q_2 as $\Delta E_o/n_o$, where n is voltage division ratio of $R_1 - R_2$ and R_3.

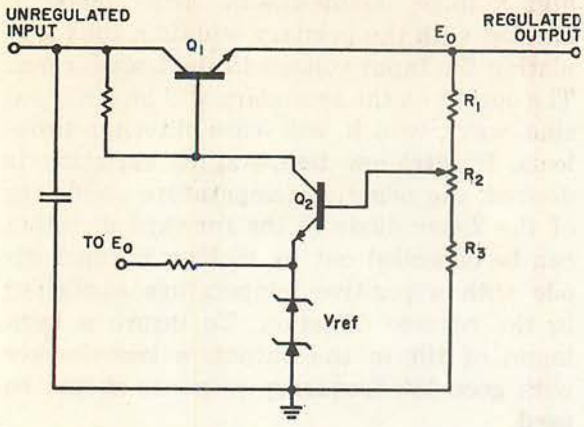


Fig. 2. Modified regulator allows *n* to approach unity so that regulator is more sensitive to changes in output voltage.

This can be done with the circuit shown in Fig. 2. The method is applicable to both vacuum tube and transistor series regulators.

In this circuit Q_3 is a constant current source so that:

$$I_1 >> I_2 \text{ and } I_1 \cong \frac{V_{ref}}{2R_3}$$

Since I_1 is independent of E_o, a change in E_o or ΔE_o, is seen at the base of Q_2. Thus, this circuit avoids the voltage division that is inherent in the series regulator of Fig. 1. This can also be accomplished by replacing R^1 of Fig. 1 by a Zener diode. However, this would lead to temperature drift problems that are several orders of magnitude greater than the temperature drift associated with the circuit of Fig. 2.

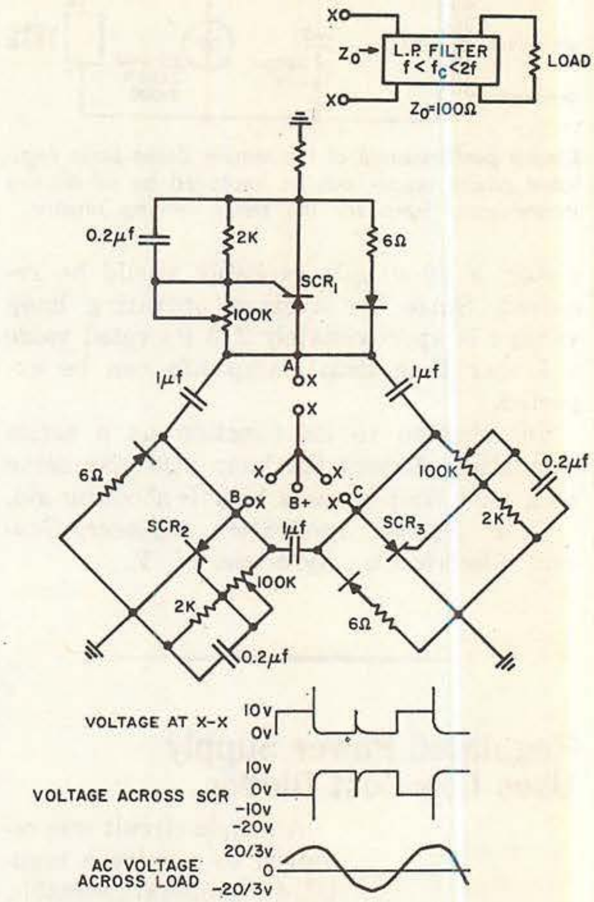
Marvin Shapiro, electrical engineer, Vector Manufacturing Co., Southampton, Pa.

Solid-State Oscillator Supplies Three-Phase Power

In many applications it is desirable to replace rotating equipment with static equipment. A static source of 3-phase power is shown here. The power output of this circuit is limited only by the power-handling capability of the components.

Silicon-controlled rectifiers are available that can handle currents as high as 70 amp at 300 piv. Hence, it would appear that the upper limit on power-handling capacity of this circuit is 21 kw.

A one-watt oscillator was designed to test the technique. The filter and load at *X-X* were replaced with a 100-ohm resistor. The waveforms of one phase of the oscillator are as shown. The other phases were the same, except for a 120-deg phase shift. A slight



FUNDAMENTAL: $\frac{20f}{\pi} \sin \frac{\pi}{3f} \cos 2\pi f t \cong \frac{20}{3} \cos 2\pi f t, f >> 6\text{CPS}$

Basic configuration of 3-phase power source with waveforms for one phase.

potentiometer adjustment was required to balance the circuit.

The delta capacitor arrangement causes commutation of the SCRs. When B+ is applied, one of the SCRs will trigger on as the voltage on its gate capacitor charges sufficiently. Assume SCR_1 is triggered on first. The voltage at points *A*, *B* and *C* will go to ground potential. The voltage at point *A* will remain at ground potential and point *B* and *C* will increase to B+ potential. At this time one of the remaining SCRs—say SCR_2—will trigger on. If SCR_2 triggers on, point *B* will go to ground potential. This will cause a negative potential to appear across SCR_1, turning if off. Just before SCR_2 triggers on, *B* and *C* are at the same potential. When point *B* goes to ground, the charge on the capacitor between A and C will redistribute between this capacitor and the capacitor between points B and C. Hence, just after SCR_2 triggers on, point C will be at the highest potential; therefore, SCR_3 will be the next to trigger on. After SCR_3 triggers on, which triggers SCR_2 off, SCR_1 will trigger on again. In this way each SCR will have a duty cycle of 1/3 and three-phase power will be generated. The filter before the load is a low-pass filter with a characteristic impedance of 100 ohms which passes only the fundamental and dc components of the voltage at X-X.

If the pots are adjusted accurately, the voltages across the three loads will be sinusoidal and 120 deg out of phase with respect to each other.

William B. McCartney, Jr., engineer, Westinghouse Electric Corp., Baltimore, Md.

"Junk-Box" Current Limiter Protects Breadboard Equipment

One of the most discouraging aspects of electronics laboratory work is an accidental short circuit, thermal runaway or other mishap that causes damage or destruction of valuable components in a breadboard or other developmental circuit. Therefore, a current-limiting device that can be inserted between the output of any dc power source and a circuit under test often is desirable.

The circuit shown here is a simple current limiter, which can be used with any existing dc power source. Since it uses only four components, it can be assembled quickly in temporary form from junk-box parts.

Transistor T_1 is a germanium pnp unit and D_1 is a silicon diode. Approximate design equations are as follows.

$$I_{cf} = \frac{V_{D1} - V_{be}}{R_E}$$

$$R_B = \frac{V_{s(min)} H_{FE(min)}}{I_{cf}}$$

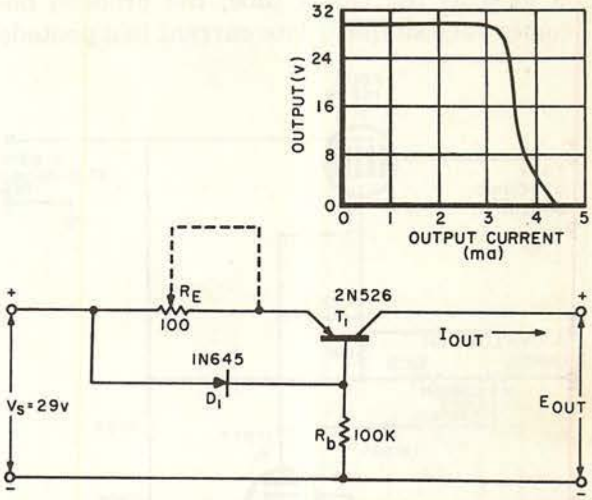
$$I_{C(max)} \geqslant I_{cf}$$

$$P_{C(max)} \geqslant V_{s(max)} I_{cf}$$

$$V_{CEO} \geqslant V_{s(max)}$$

where:

I_{cf} = output cutoff current.
V_{D1} = forward voltage drop of D_1.
V_{be} = forward base-emitter drop of T_1.
$V_{s(min)}$ = minimum supply voltage.
$Vs_{(max)}$ = maximum supply voltage.



Limiter circuit and output characteristics using components and values as shown.

If a potentiometer is used for R_E, a continuously variable cutoff current will be available. By connecting an ammeter across the output and applying an input voltage, R_E can be calibrated directly in μa, ma, or amperes.

The lower limit on cutoff current is the I_{CBO} of the transistor used. The output impedance of the circuit before cutoff is equal to $R_E + R_{sat}$ where R_{sat} is the saturation impedance of T_1.

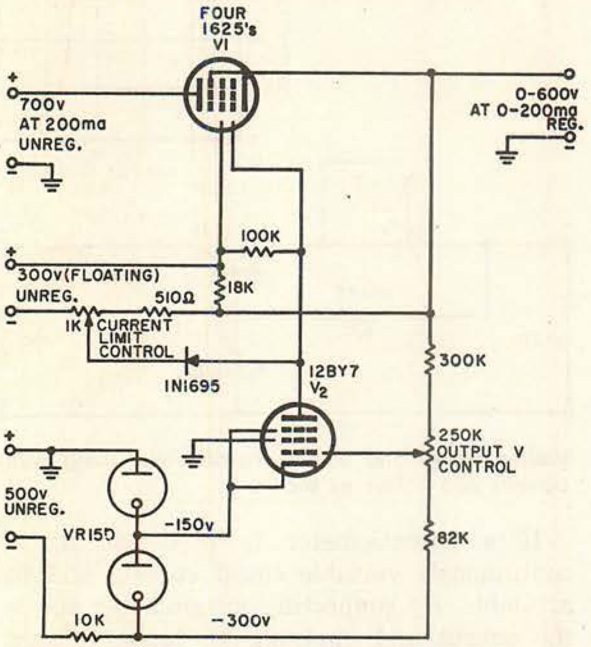
This circuit also makes a handy current source for one-shot timing applications and other uses.

Donald A. Boelter, associate engineer, Martin-Marietta Corp., Baltimore 20, Md.

Pentode Replaces Triode For Current-Limiting in Tube Supply

Current-limiting is a rather common feature of transistorized laboratory power supplies. But it is rather difficult to achieve in a series-regulated vacuum tube supply. This is because it is difficult to sense load current, necessary to bring about current limiting, in this type of supply.

But if a pentode instead of the usual triode is used as the series tube, the problem becomes very simple. Plate current in a pentode



Current-limiting is introduced in vacuum tube supply by using a pentode instead of the usual triode series tube. The control grid of the pentode is clamped with a diode so that its bias is held above a minimum value.

depends primarily upon control grid voltage. Even if the plate voltage should increase greatly, the plate current rises very little. If the control grid is clamped with a diode so grid bias cannot be reduced beyond a set value, the tube automatically will go into current limiting.

Such a supply was built using transmitting pentodes for the series tubes to permit high plate dissipation. Low resistances were used in series with the control grid and the screen grid leads of each tube for parasitic suppression. They also were placed in the cathode of each tube for current equalization.

In the circuit shown in the figure, as load current increases, the output voltage drops slightly. V_2 senses this drop and counteracts it by raising the voltage applied to the grid of V_1. This is the normal operation of a voltage regulator.

However, when load current reaches the current limit value, the grid of V_1 is stopped from going higher by the conduction of the silicon diode. This clamps the grid-to-cathode voltage of V_1 to a preset bias, and current limiting begins.

Herbert Zimmerman, development engineer, Wilcox Electric Co., Kansas City, 27, Mo.

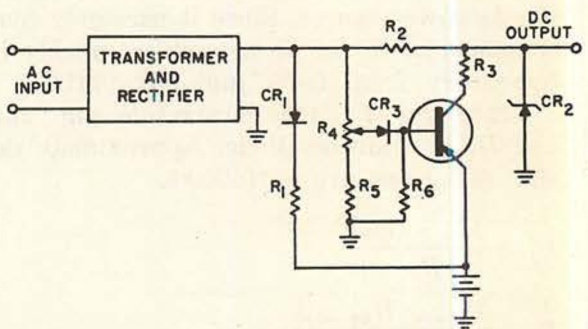
Standby Batteries Protect Supply Against AC Power Loss

We had to design a power supply that could "ride out" a loss of ac power. The simplified circuit shown in the figure served us perfectly.

Zener diode CR_2 and resistor R_2 form a simple shunt regulator. With line voltage applied, the rectifier supplies regulator current and holds transistor Q_1 cut-off through the base-biasing network of R_4, R_5, R_6, and CR_3. Decreasing line voltage drops the base potential and the transistor begins to conduct. Thus, the batteries, through R_3, furnish the regulator current which would normally be lost on reduced input voltage.

When a total loss of ac voltage occurs, the transistor saturates and the batteries carry the entire load. CR_1 and R_1 form a charging network that keeps the batteries charged as long as line voltage is available.

In our system the dc voltage output varies less than 0.2 per cent over the range of 0-130 v rms input.



When a total loss of ac input voltage occurs, the transistor saturates and the standby batteries carry the entire load.

G. Douglas McKinley, development engineer, Forney Engineering Co., Dallas, Tex.

Transient Clipper For DC Converters

In a dc converter, as shown, each transistor is ideally subjected to twice the supply voltage V_{CC} when in the OFF condition. For example, if Q_1 is conducting, the induced voltage across N_2 will be equal to V_{CC} and will add to the supply voltage so that $V_{CE2} = -2V_{CC}$.

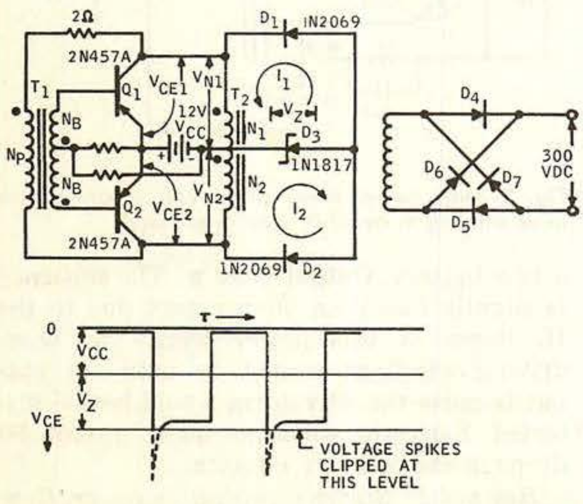
However, due to leakage inductance in the power transformer, this situation is not always realized. When the transistor switches off, the sudden change in current through any leakage inductance in the collector circuit will result in a high voltage spike appearing from collector to emitter on the OFF transistor. This spike may be several times the supply voltage and may exceed the voltage breakdown rating of the transistors.

The method suggested here suppresses these transients. Assume Q_1 is conducting and Q_2 is off. Voltage V_{CE2} will be equal to the supply voltage plus any voltage that is induced in N_2. Zener diode D_3 is chosen to have a breakdown slightly above the supply voltage. When V_{N2} tends to exceed V_Z, the Zener diode breaks down causing a current I_2 to flow as shown. This results in V_{N2} being clamped at approximately V_Z, thereby maintaining V_{CE} at a level within the transistor rating. Transistor Q_1 is protected in a like manner against voltage spikes in N_1 through current path I_1. Diodes D_1 and D_2 prevent a reverse flow of currents I_1 and I_2, which would otherwise short the windings N_1 and N_2, and the applied voltage V_{CC} alternately appears across them.

This system offers an advantage of lower cost over systems where two Zener diodes are used. Another advantage is better control over clipping level. Since Zener tolerance is based on a percentage of voltage rating, closer clipping can be achieved here as tolerance is a percentage of V_{CC} and not $2V_{CC}$.

A small capacitor may be placed across D_3 to help clip the leading edge of very fast rising voltage transients. It is generally not required that diodes D_1 and D_2 have fast recovery times since the transient voltage in this type of circuit lasts for only a small percentage of the period of oscillation. Typical circuit components are shown in Fig. 2a.

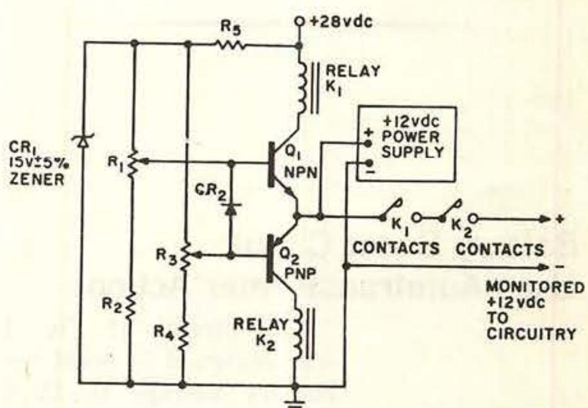
Paul Vergez, Transistor Products Div., Texas Instruments Inc., Dallas, Tex.



Single Zener clipping diode removes waveform spikes.

Voltage, Current Limiter Protects Circuitry From Shorts and Overs

Laboratory breadboards can be protected from shorts and over-voltages by connecting the circuit shown in the figure across the output of the power source. The circuit is quite useful when transistorized circuitry, easily damaged by high voltages or currents, is being tested. Because it is designed to have adjustable limit settings,



Circuit connected across power supply outputs removes power to test circuitry if shorts or over-voltages occur.

the circuit can be used for different applications.

In normal operation, transistors Q_1 and Q_2 are conducting and relays K_1 and K_2 are energized. The relay contacts are closed and the load will receive +12 v dc. Tolerances around +12 v are set by potentiometers R_1 and R_3. R_1 is set for the high voltage limit and R_3 for the low voltage (short circuit) limit.

If the power supply should suddenly produce an over-voltage, transistor Q_1 will be turned off, de-energizing relay K_1 and opening the K_1 contacts. If the load should suddenly short, transistor Q_2 will be turned off, de-energizing relay K_2 and opening the K_2 contacts.

Resistors R_2 and R_4 prevent overdriving Q_1 and Q_2; CR_2 prevents reverse-biasing the transistors more than about 1 v. Zener diode CR_1 provides a stable reference voltage.

For fast operation, relays K_1 and K_2 should be mercury-wetted contact relays similar to the C. P. Clare HGS series. These relays permit 2 msec drop-out times that are compatible with transistor thermal time-constants. With the Clare relays and military components, the MTBF of the circuit is about 12,000 hours.

Cecil R. Frost, senior project engineer, Canoga Electronics Corp., Van Nuys, Calif.

Battery Boost Circuit Uses Autotransformer Action

The circuit of Fig. 1 was designed to boost 6-v battery voltage to 12 v at about 5-w output. Since no secondary winding is used and the collectors are grounded, the circuit is very simple and can be constructed in a very small space. An efficiency of 90 per cent is obtained because the dc-to-dc converter is essentially in series with the battery. Battery voltage appears at the output even when the converter is disabled.

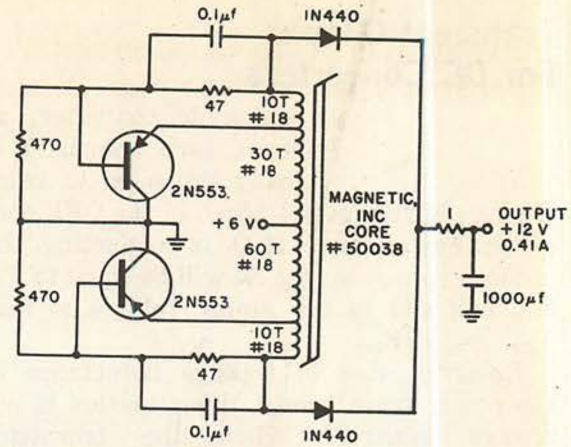


Fig. 1. High efficiency is obtained in battery boost circuit by permitting batttery to supply half of the voltage directly.

The circuit has been used in a 6-v automobile to power a 12-v radio. The circuit of Fig. 2 provides higher power to operate a 24-v image-orthicon television camera on a 12-v battery. Output is 90 w. The efficiency is slightly less than 90 per cent due to the IR losses at this power level. The base-driving windings cannot be used for output because the waveform would be too distorted. Separate windings make up for IR drops in the primary circuits.

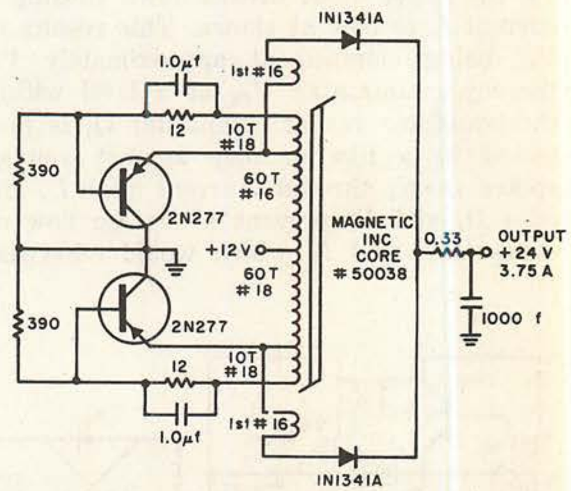


Fig. 2. High power boost circuit uses separate base drive winding to avoid IR drop in primary.

Howard F. Stearns, project engineer, General Electric Co., Syracuse, N. Y.

Inexpensive DC-to-DC Converter Operates Without Rectifier Diodes

Most dc-to-dc converters use two transistors, a transformer and rectifier diodes. The converter circuit shown in Fig. 1 uses a single transistor, which serves as both oscillator and rectifier.

The load and its shunting capacitor bias the transistor off. The transistor therefore is turned on for only a small angle in each cycle. A conversion efficiency of 70 per cent has been attained for a 5-to-1 (3 v-to-15-v) step-up at an operating frequency of 100 kc.

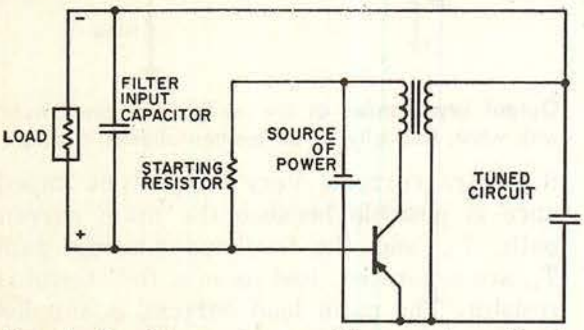


Fig. 1. Transistor acts as oscillator and rectifier in dc-to-dc converter.

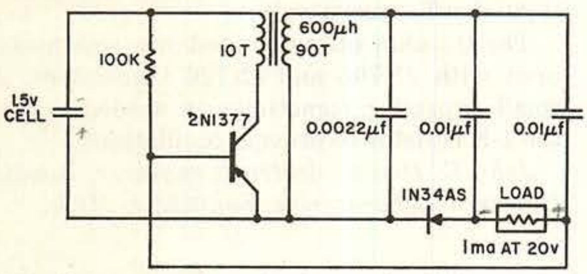


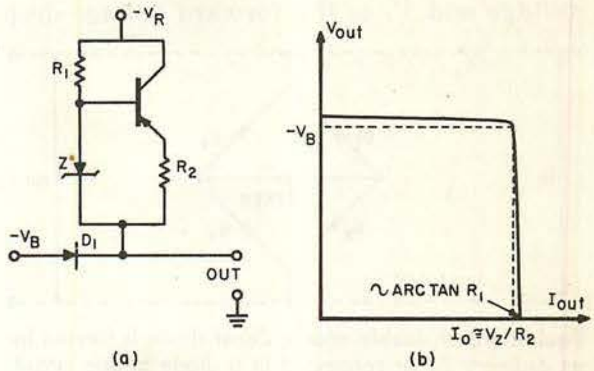
Fig. 2. Base-to-emitter diode provides full-wave doubling for higher step-up ratios.

For higher step-up ratios, or where the load voltage would exceed the base-to-emitter avalanche potential in the circuit of Fig. 1, the modified circuit of Fig. 2 is useful. Here, the base-emitter diode serves in a full-wave doubler. This circuit has been used to supply 20 v at 1 ma when driven by a penlight cell.

W. R. Kundert, development engineer, General Radio Co., Concord, Mass.

Simple Current-Limited Voltage Source

During the experimental stages of a circuit design it sometimes is convenient to use individual current-limited voltage sources for some critical stages—for instance a transistor stage containing only a transformer in its collector circuit.



Current-limited voltage source (a) provides output characteristic at (b).

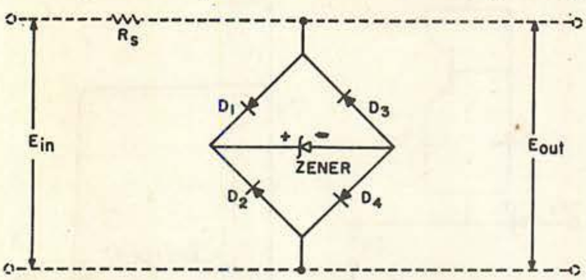
The circuit consists of a constant-current generator (transistor, Zener diode and resistors R_1 and R_2) and a catching diode biased by the voltage $-V_B$.

If the current required by the load is less than that supplied by the constant-current generator diode D conducts. Hence the output voltage is slightly less than $-V_B$ and the output impedance is equal to the differential resistance of the diode D. However, if the current required by the load equals or exceeds the current supplied by the constant-current generator, the diode D is cut off and the output impedance of this device increases rapidly to that of the constant-current generator.

Zvi Netter, research engineer, Ministry of Defense, Hakirya, Tel Aviv, Israel.

Zener, Diode Bridge Form Double-Ended Clipper

A symmetrical double-anode Zener diode equivalent can be constructed very simply by using an ordinary Zener and a diode bridge circuit. The output voltage is equal to $\pm (V_Z + 2V_f)$ where V_Z is the Zener voltage and V_f is the forward voltage drop



Equivalent of double-anode, Zener diode is formed by an ordinary Zener connected in a diode bridge circuit.

of the diodes. The output waveform is almost perfectly symmetrical since both positive- and negative-going voltages are clipped by the same Zener. If extreme symmetry is required, D_1 and D_2 and D_3 and D_4 can be matched diode pairs.

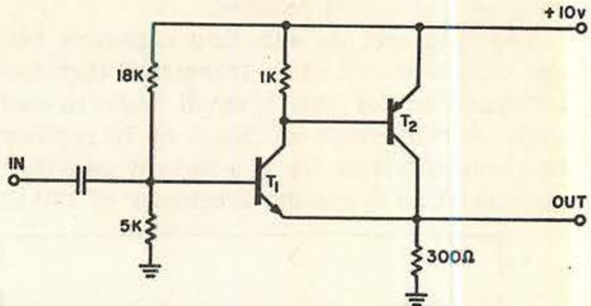
Donald A. Boelter, associate engineer, Martin-Marietta Corp., Baltimore, Md.

Modified Emitter Follower Has Very Low Output Impedance

Low output impedances often are obtained from a transistor amplifier by using an emitter-follower output stage. For very low impedances, two emitter followers in cascode may be used. But the output impedance of any emitter follower is limited by the properties of the transistors. For example, with the much-used 2N706, the output impedance will not go below about 3 ohms with double or even triple emitter followers.

With the circuit shown, however, we were able to overcome this limitation and operate with a measured output impedance of less than 0.1 ohm.

Transistor T_1 operates as an ordinary emitter follower; T_2 "helps" T_1 by supply-



Output impedances as low as 0.1 ohm are possible with what, basically, is an emitter-follower circuit.

ing extra current. Very low output impedance is possible because the main current path, T_2, and the feedback sensing path, T_1, are separated, just as in a four-terminal resistor. The main load current is supplied by the collector of T_2. T_1 senses the difference between input and output voltage and regulates T_2 accordingly.

The 0.1-ohm output impedance was measured with 2N706 and 2N726 transistors. A small capacitor sometimes is needed across the 1-K resistor to prevent oscillations.

John K. Dixon, electrical engineer, Bendix Research Laboratories, Southfield, Mich.

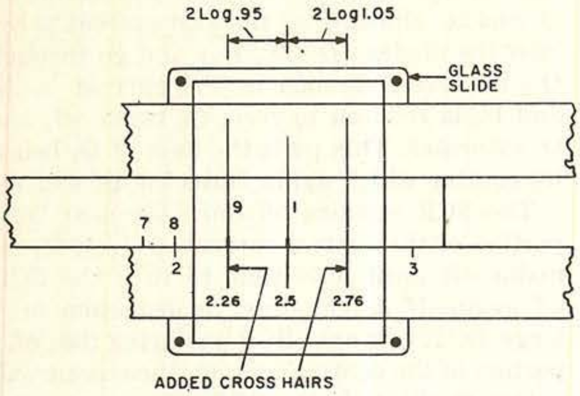
Component Tolerance Calculator Can Be Built Into Your Slide Rule

When calculating the worst-case conditions of a circuit it is often necessary to multiply or divide parameters with their tolerances and to obtain the extreme values of the result. But, by adding two cross hairs to the glass slider of the slide rule, the extremes may be read at the same time as the nominal value.

The extra lines are placed according to the principle that moving along a log scale a fixed distance is the same as multiplying or dividing by a constant factor.

For 5 per cent components one cross hair is located at twice the distance from the index to 1.05 and the other is located at twice the distance from the index to 0.95.

For 10 per cent components the same slider may be used on the *A* and *B* scales.



Added cross hairs drawn on glass slide are placed to indicate extreme values when divided numbers have ± tolerances. Illustration shows settings for 2 ±5% / 8 ±5%.

A slider scale calibrated at log 1.02, 1.04, 1.06, etc., would be more useful. The answer tolerance would be read at the line indicating the sum of the component tolerances.

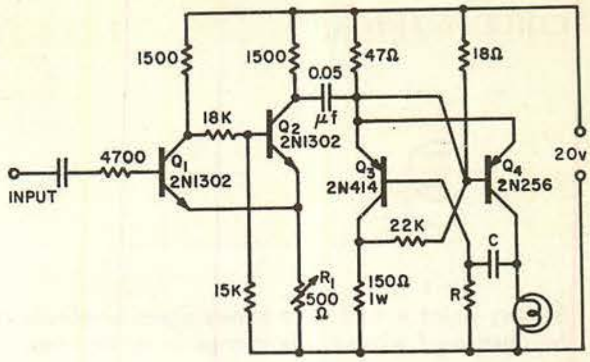
To use this device the calculation is performed with the nominal values. Then, with the slider hair line on the answer, the extreme values appear under the added cross hairs.

Bruce Ross, associate research engineer, The Boeing Co., Seattle, Wash.

Schmitt Circuit Monitor Triggers on Overloads

A simple overload alarm was needed to give a visual signal when the output of an amplifier exceeded a certain limit. The circuit shown was designed to work in the range of from 1 to 6 v rms.

Transistors Q_1 and Q_2 form a conventional Schmitt trigger with Q_2 conducting and Q_1 cut off. The emitter of Q_1 is at a potential determined by the current flowing in the common emitter resistor R_1. When



Overload alarm with 10-mv sensitivity uses Schmitt trigger circuit driving multivibrator.

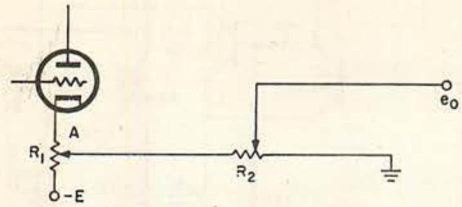
the input voltage does not exceed this level, the alarm presents to the source a very high impedance (that of the reverse-biased base-emitter diode of Q_1). When the peak of the ac signal exceeds the emitter bias of Q_1, the circuit is triggered, providing a positive pulse at the collector of Q_2. This pulse is coupled to the one-shot multivibrator formed by Q_3 and Q_4. The alarm lamp is the collector load of Q_4 which normally is cut off. If the period of the one shot (determined by *R* and *C*) is chosen close to the period between successive peaks of the input waveform, the lamp will remain lit as long as the limit is exceeded and stay out at all other times. These sensitivity is about 10 mv.

Bill Gutman, project engineer, Kearfott Div., General Precision, Inc., Clifton, N. J.

Amplitude Control in DC-Coupled Circuit Holds DC Level Constant

Usually if a potentiometer is used to pass on an amplitude variation in a dc-coupled amplifier, the average dc level will also change. To prevent this the arrangement shown in the figure has proved to be quite effective.

Point *A* is varied until it is at 0 v dc. Then any variation in R_2 will only affect the amplitude of the ac signal and not the dc level. R_2 can be much larger than R_1 so that the full amplitude at the cathode can



Setting point *A* to 0 v dc allows signal amplitude to be attenuated without any change in its dc level.

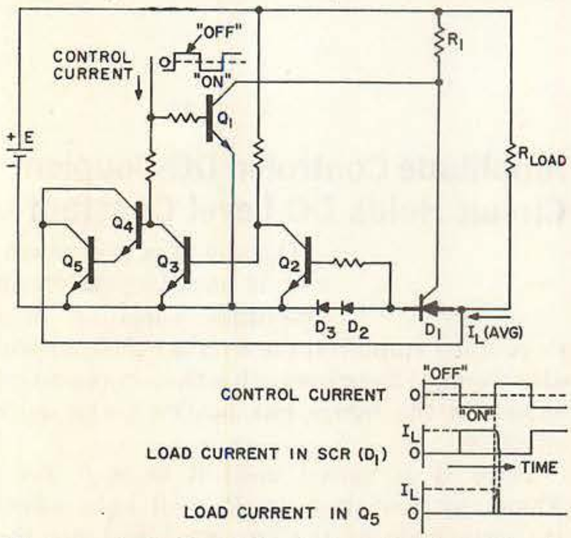
be utilized. However, the designer must choose between the increased output impedance and the larger useful amplitude.

Irving Bayer, senior member, technical staff, Radio Corp. of America, New York, N. Y.

Shorting Transistor Reduces SCR Turn-Off Time

Silicon-controlled rectifiers have not been used extensively in dc circuits because of the difficulty in turning them off. But an SCR can be turned off efficiently by shorting it with a transistor for the brief period required to turn it off.

In the circuit shown in the figure, control current is fed into the bases of transistors Q_1 and Q_4. When the "on" portion of the control current flows, Q_1 is off and allows gate current through R_1 to the SCR, D_1. Transistors Q_4 and Q_5 are also off, so that the entire load current passes through D_1, D_2, and D_3. The voltage drop across D_2 and D_3 is used to saturate Q_2. This blocks any base current to Q_3.



Silicon-controlled rectifier is turned off rapidly by shorting it with a transistor, Q_5, during the brief turn-off transition.

When the "off" portion of the control current flows, it saturates Q_1, which cuts off gate current to the SCR. The control current also saturates the Darlington connection of Q_4 and Q_5, allowing all the load current to bypass the diodes (D_1, D_2, D_3) and go through Q_5. However, as soon as the current in D_2 and D_3 is reduced to zero, Q_2 turns off, and Q_3 saturates. This pulls the base of Q_4 below its emitter which again turns off Q_4 and Q_5.

The SCR remains off until the next "on" portion of the control current arrives. Q_5 remains off until it is time to turn the SCR off again. If a temporary malfunction or a large dv/dt turns the SCR on during the "off" portion of the control current, the circuit will automatically turn the SCR off.

A typical turn-on time for a high current SCR is 0.5 μsec, while a typical turn-off time is 5.0 μsec. Transistor Q_5 handles only a small fraction of the total load current while the SCR does most of the work. Thus, high switching rates can be realized without dissipating a great deal of power in Q_5. Since the SCR is either off or on, it also dissipates very little power.

It might be noted that insulation hardware for Q_5 and D_1 is not necessary if the circuit is grounded to the chassis at the anode of D_1.

D. K. Phillips, Member of the Technical Staff, Hughes Aircraft Co., Culver City, Calif.

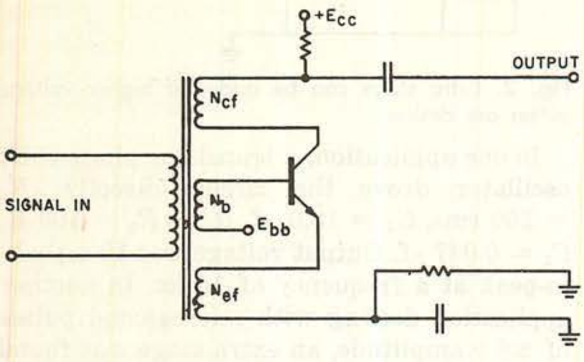
Positive, Negative Feedback Combine to Reduce Noise

Negative feedback does not reduce collector noise because both signal and noise are decreased proportionately. Positive emitter feedback may be helpful but results in signal instability.

The circuit in Fig. 1 combines positive emitter feedback with negative collector feedback in equal proportions. Consequently, the signal amplification is unchanged, but the collector noise is reduced by

$$\beta\left(\frac{N_{cf}}{N_b}\right)$$

where β = forward current transfer ratio, N_{cf} = number of turns on collector feedback



Positive emitter feedback and negative collector feedback combined reduces noise figure while retaining good signal amplification.

winding, and N_b = number of turns on the base winding. N_{ef} in Fig. 1 represents the number of turns on the emitter feedback winding.

In this circuit, there is no need to isolate the dc component of the feedback currents because their effects in the transformer cancel.

Kermit Norris, technician, General Dynamics, Pomona, Calif.

VR Tube, Removed From Load, Is Fired By Source

Occasionally, a voltage regulator tube will not be able to fire because there is insufficient potential across its terminals. Such a situation is present, for example, in the circuit of Fig. 1.

A solution to the problem is shown in Fig. 2. A silicon diode D_1 disconnects R_1

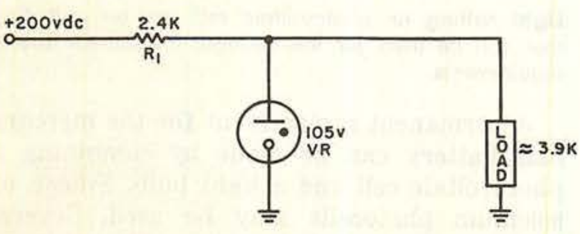


Fig. 1. VR tube will not fire because circuit does not put sufficient potential across its terminals.

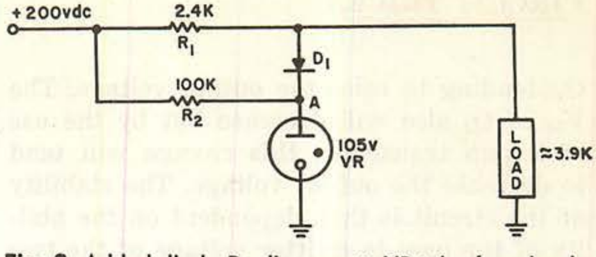


Fig. 2. Added diode D_1 disconnects VR tube from load, fires it directly from source.

and the load from point A until the VR is fired by the potential applied through R_2. The circuit then operates in the normal manner except for about a 0.5-v increase in the regulated voltage because of the drop across D_1.

Matt Cousins, electrical engineer, Airtronics Inc., Washington, D. C.

Temperature-Stabilizing Emitter Followers

There are many applications in instrumentation systems for emitter followers. The conventional emitter follower in Fig. 1a has two problem areas when used in dc-circuit applications. These are: an inherent drift with temperature produced by variations in the base-to-emitter voltage (V_{be}), and a dc offset equal to the base-to-emitter voltage. The temperature stabilized emitter follower of Fig. 1b provides a simple but effective method of reducing both effects.

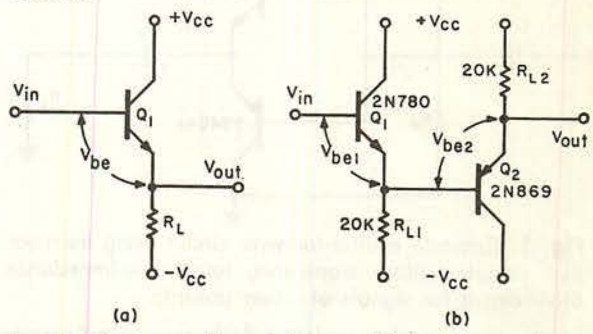


Fig. 1. Basic emitter follower (a) is temperature-stabilized by addition of opposite-polarity stage (b).

The principle of operation of the temperature-stabilized emitter follower is based on the cancellation of the base-to-emitter voltage (V_{be}) of the first stage by an equal, but opposite polarity, base-to-emitter voltage in the second stage. An increase in temperature will produce a decrease in V_{be} of

Q_1, tending to raise the output voltage. The V_{be} of Q_2 also will decrease but by the use of a pnp transistor, this change will tend to decrease the output voltage. The stability of the circuit is then dependent on the ability of the base-to-emitter voltage of the two transistors to track over an extended temperature range.

A series of tests has been conducted using 2N780 transistors as Q_1 and 2N869 transistors as Q_2. The maximum drift obtained over the temperature range of 20 F to + 200 F was 30 mv, with the average drift about 25 mv. The transistors used in this test were randomly selected and no attempt was made to match transistor characteristic.

M. H. Schmidt, instrumentation engineer, McDonnell Aircraft Corp., Florissant, Mo.

Dual-Polarity Signal Drive Uses Cascode Emitter Follower

To provide a low-output impedance drive circuit for signals of either polarity, we connected complementary transistors, as shown in Fig. 1. This circuit has a low current drain and can use inexpensive transistors. The pnp and npn transistors

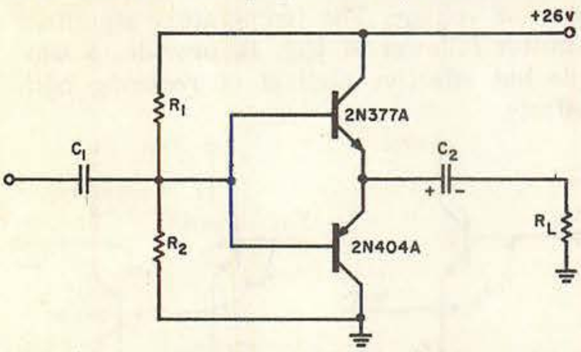


Fig. 1. Cascode emitter-follower circuit using inexpensive complementary transistors, forms low-impedance drive circuit for signals of either polarity.

form a cascode emitter-follower configuration in which they cut each other off, drawing a negligible amount of quiescent current.

A signal of either polarity, swinging around the bias determined by voltage dividers R_1 and R_2, causes the respective transistor to conduct. This shuts off the other transistor even more, and all the current flows to the load R_L through C_2.

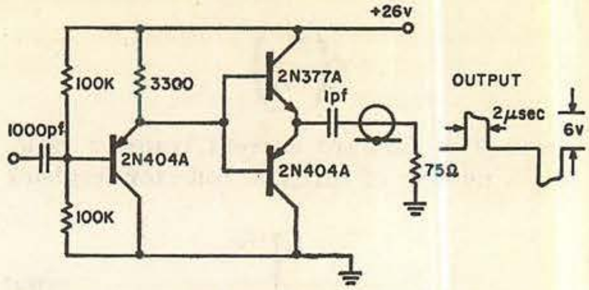


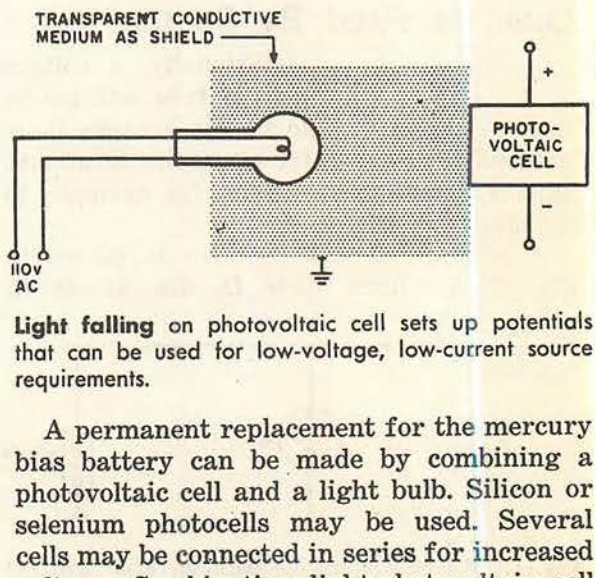
Fig. 2. Extra stage can be added if higher voltage pulses are desired.

In one application, a transistor phase-shift oscillator drove the circuit directly: $R_L = 200$ rms, $C_2 = 10.0\ \mu f$, $R_1 = R_2 = 100$ K, $C_1 = 0.047\ \mu f$. Output voltage was 10 v peak-to-peak at a frequency of 10 kc. In another application dealing with microsecond pulses of ± 6 v amplitude, an extra stage was found desirable, Fig. 2.

Gunnar Richwell, staff engineer, Reflectone Electronics Inc., Stamford, Conn.

Lighted Photocell Replaces Mercury Bias Battery

Requirements in electronics equipment for low-voltage, low-current biasing sources are met, when convenient, with mercury batteries. Their great disadvantage is that they must be periodically replaced. The alternative of a transformer-rectifier-Zener diode combination is relatively bulky and expensive.



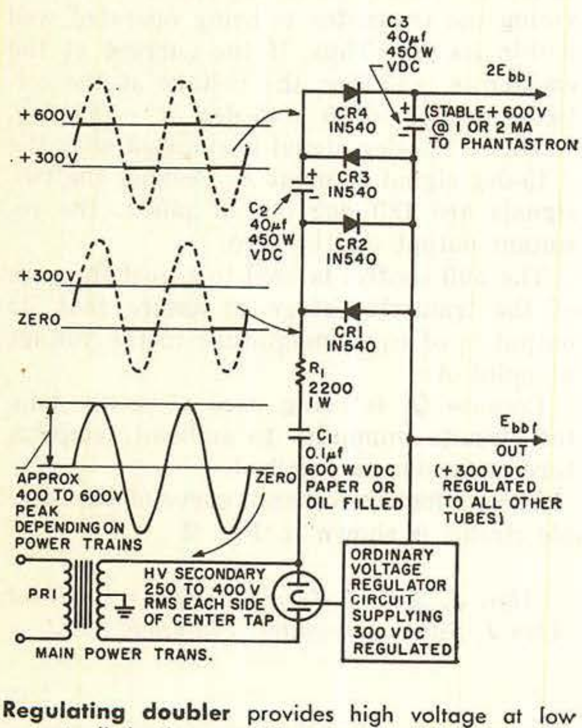
Light falling on photovoltaic cell sets up potentials that can be used for low-voltage, low-current source requirements.

A permanent replacement for the mercury bias battery can be made by combining a photovoltaic cell and a light bulb. Silicon or selenium photocells may be used. Several cells may be connected in series for increased voltage. Combination light-photovoltaic cell units can be made for this purpose, analogous to the light-photoresistor combinations that are now commercially available.

Dr. F. W. Cope, research scientist, Aviation Medical Acceleration Laboratory, Naval Air Development Center, Johnsville, Pa.

Regulator-Doubler Eliminates Need for Second Power Supply

Design of a vacuum-tube circuit included one phantastron-type stage to generate a timing sawtooth having a peak-to-peak amplitude of 500 v. This stage required stable B+ voltage of +600 v at a current of approximately 1 ma. The remainder of the tubes in this chassis required only +300 v, which was being supplied by a regulated power supply circuit of ordinary design. The accompanying schematic shows a "regulated voltage-doubling circuit" used to supply the required 1 or 2 ma at +600 v to



Regulating doubler provides high voltage at low current eliminating need for second power supply.

the phantastron only. This eliminated the need for an additional regulated power supply in the design.

Diodes CR_1, CR_2, CR_3, CR_4 and capacitors C_2 and C_3 need withstand only 300 v. C_1 must withstand the peak value of the ac voltage across half of the secondary winding. R_1 was included to limit the surge current during the turn-on transient. A much lower resistance value—perhaps 47 ohms—would probably be satisfactory for most applications.

A bleeder resistor (now shown) of about 1 megohm may be desirable for safety reasons in some applications. If used, it may be connected in parallel with C_3.

This circuit is quite efficient because it does not employ a power-dissipating principle to accomplish its voltage-regulating function. It does not, therefore, contribute to the chassis heat-dissipation problem.

James W. Carroll, staff member, Sandia Corp., Albuquerque, N. M.

Low Impedance Line Driver Uses Standard Coils

Frequently, a terminated cable must be driven from a wideband if source. The optimum matched condition for this calls for an equal-Q, transitionally-coupled stage.

To achieve equal-Q loading under conditions of low load impedance, it is usually

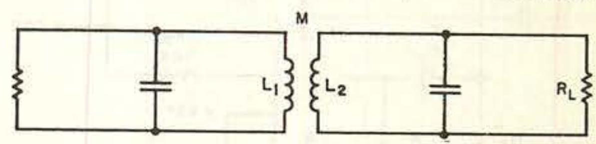


Fig. 1. If shunt capacity is required in a transformer secondary to achieve equal-Q loading with a low load impedance, negative values of M may be required in the Tee-transform of the circuit.

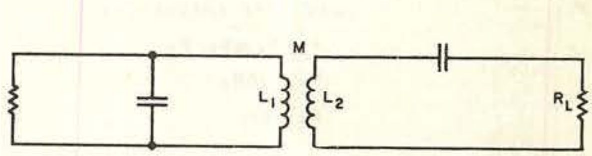


Fig. 2. With a series-tuned circuit in the secondary, L_2 will be much greater than M.

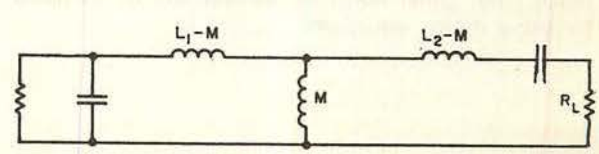


Fig. 3. The result of series tuning is that standard, not negative, coils are present in the Tee-transform circuit.

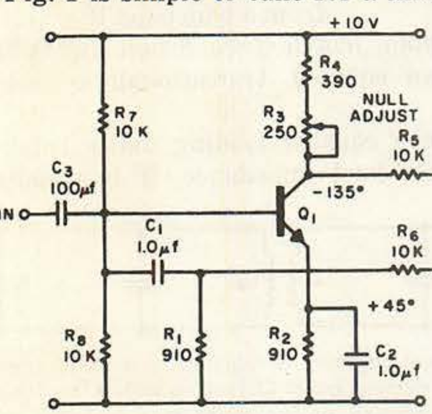
necessary to load the transformer secondary with shunt capacity as in Fig. 1. This reduces the value of the secondary inductance L_2. For values of L_2 less than M, a negative value of M is required in the Tee-transform of the circuit.

However, if a series-tuned circuit of the same Q is used in the secondary, Fig. 2, values of L_2 much larger than M are obtained. This allows standard coils to be used in the Tee-transform of the transformer, as in Fig. 3.

Martin E. Doyle, design group supervisor, Raytheon Co., Airborne Operation, Sudbury, Mass.

Novel Notch Filter Is Easy to Tune for Null

Unlike a bridged-T or parallel-T circuit, the 185-cps notch filter shown in Fig. 1 is simple to tune for a minimum null,



DESIGN EQUATIONS

$R_1 = \frac{1}{\omega C_1}$

$R_2 = \frac{1}{\omega C_2}$

$R_4 \approx \frac{R_2}{2}$

$R_3 \approx \frac{R_2}{4}$

(ω = CUTOFF FREQUENCY)

$R_5 = R_6 \approx R_7 = R_8$

$R_5 \approx 10 R_2$

$C_3 > 10_{C1}$

Fig. 1. Circuit gives component values for 185-cps notch filter. Other notch frequencies can be obtained by using design equations.

and does not require close tolerance components.

The X_c of C_1 is equal to R_1 at the rejection frequency, providing a 45-deg voltage lead at point A.

The X_c of C_2 equals R_2 at the rejection frequency, providing a 45-deg current lead at the emitter of Q_1. The voltage at the collector of Q_1 will always be 180 deg out

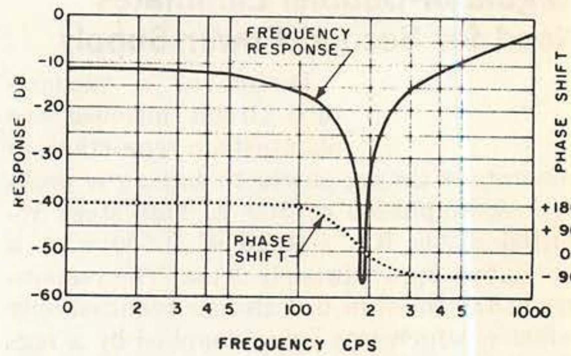


Fig. 2. Response and phase shift for 185-cps notch filter.

of phase with the current at the emitter, providing the transistor is being operated well within its f_{hfe}. Thus, if the current at the emitter is +45 deg, the voltage at the collector will be −180 +45 deg or −135 deg.

This −135-deg signal is summed with the +45-deg signal at point A. Because the two signals are 180 deg out of phase, the resultant output will be zero.

The null control is used to adjust the gain of the transistor stage to assure that its output is of equal magnitude to the voltage at point A.

Because Q_1 is being used at unity gain, the circuits immunity to ambient temperature variations is excellent.

The frequency response curve of the sample circuit is shown in Fig. 2.

Alan J. Adler, circuit design consultant, Alan J. Adler Associates, Plainfield, N. J.

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