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COLLEGE OF ENGINEERING
Department of Electrical Engineering
Space Physics Research Laboratory

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TRANSISTORIZED CIRCUITS FOR USE IN SPACE-RESEARCH INSTRUMENTATION

Prepared for the project by

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ABSTRACT

The development of various accessory circuits for use in space research instrumentation is described. The circuits include dc current detectors, voltage-controlled oscillators, voltage generators, timers and dc-dc convertors.

1. INTRODUCTION

In carrying out Langmuir probe and other experiments under the sponsorship of organizations listed elsewhere in the report, it has become necessary to develop certain electronic circuits to perform the various tasks of voltage generation, current detection, timing, calibration, telemetry, etc. In the past, such functions were provided by vacuum-tube devices; but, due to the limited volume and energy requirements placed on instrumentations of high-altitude sounding rockets, it seemed desirable to perform as many as possible of these functions using solid-state devices, thereby eliminating the need for filament power and reducing the problem of heat dissipation.

The purpose of this report is to record and describe for the possible benefit of others in the field many of the circuits and their present form. In some cases, more than one circuit was developed to carry out a specific function; and in those cases, the advantages and disadvantages of each is discussed.

2. Current Detectors

The need for a 'dc current detector,' which in effect converts a slowly varying dc current (perhaps several orders of magnitude less than a μa) into a voltage suitable for telemetry, occurs frequently in high-altitude and ionosphere research. Therefore a continuing development program is being carried out to provide dependable and sensitive circuits of this type. Two of these are discussed below.

The semiconductor current detector shown in block diagram form in Fig. 2.1, schematic Fig. 2.2, and photograph Fig. 2.3 utilizes a diode modulator, an ac amplifier, and a bridge rectifier demodulator. The modulator is a ring-bridge type using four silicon diodes, two miniature transformers, and a unijunction transistor oscillator (2 kc) as the carrier or reference generator for the modulator. The basic system has wide application and has been described in the literature.[1-5]

Operation of the system may be described as follows. On alternate half cycles of the carrier, D_1 and D_2 and then D_3 and D_4 become forward-biased. This alternately completes a path for dc input current to flow through the upper half, and then the lower half of T_2 primary, resulting in a full-wave chopping of the input current which produces a square wave of the modulator reference frequency at the secondary of T_2, the input of the amplifier. The amplitude of this square wave is proportional to the magnitude of the dc input current.

The amplifier is a high-gain miniaturized transistor amplifier similar to one that is available commercially.* Its output is coupled to the bridge rectifier by means of a step-up transformer T 3, and the rectified output is filtered by the low-

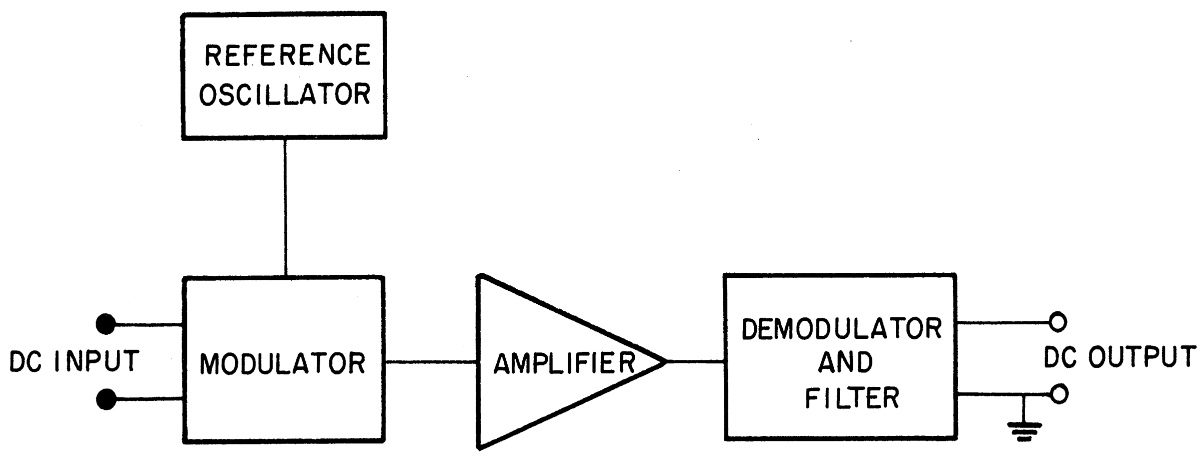
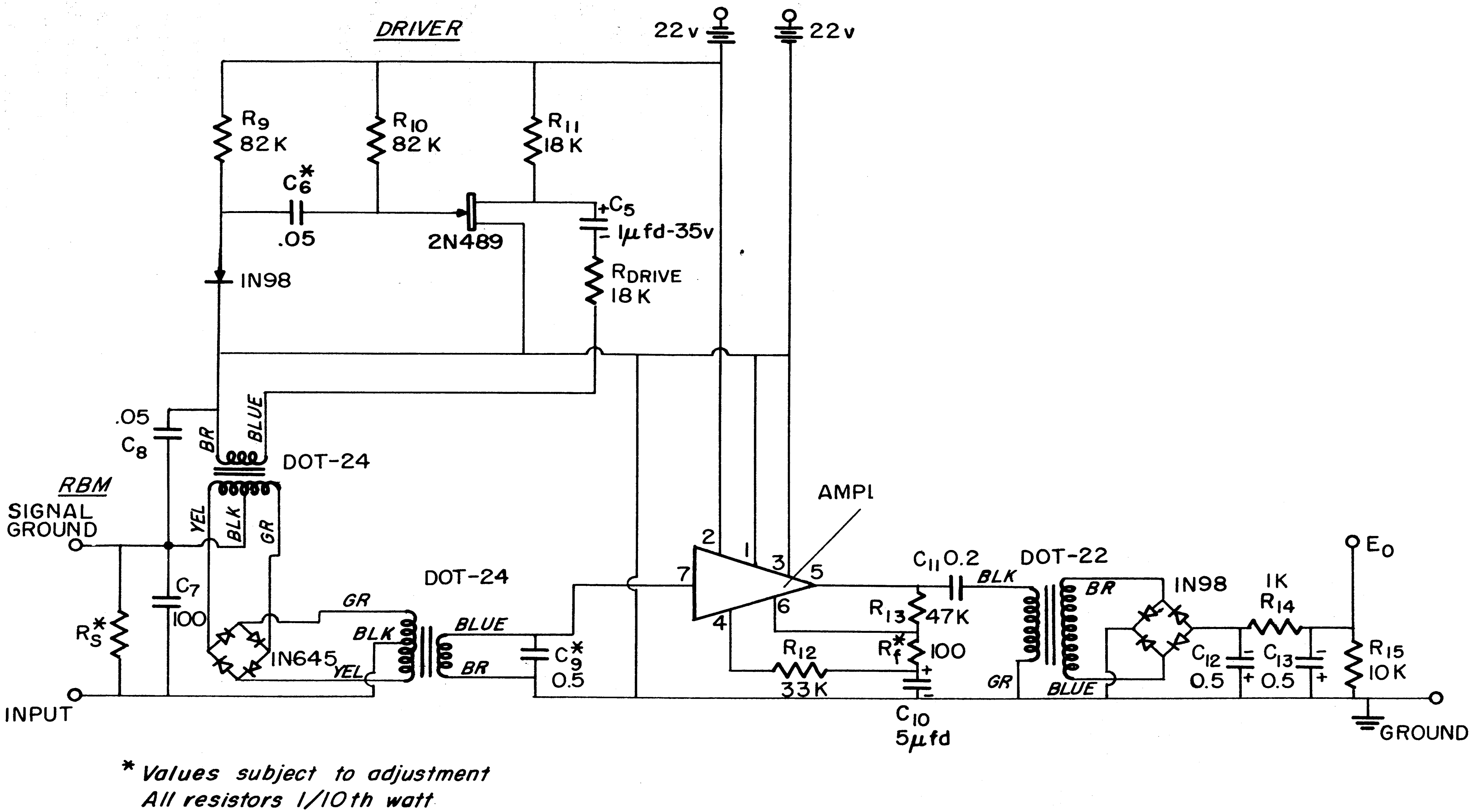


Fig. 2.1. Current-detector block diagram.

*Taber Instrument Company, No. 204G.



PROBE CURRENT DETECTOR

Fig. 2.2. Four-diode current detector.

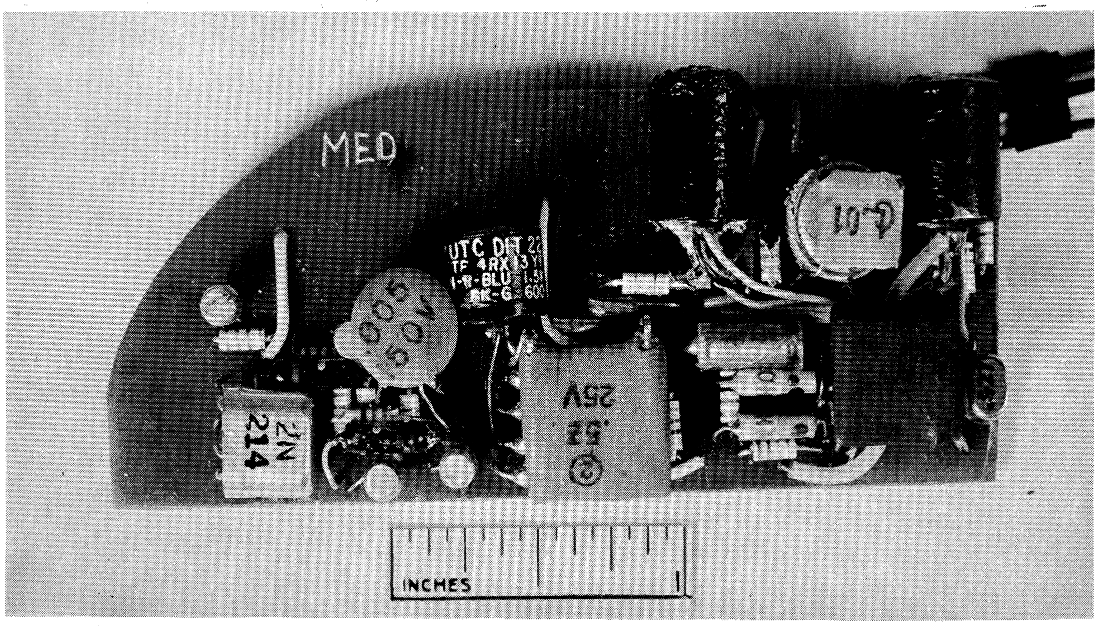


Fig. 2.3. Ring bridge and 2-diode type of detector.

pass network C 12, C 13, and R 14, providing a dc output voltage proportional to the bridge-biasing current, the input.

Some of the important characteristics of the detector are discussed below:

(a) The relation between input current, I_{in}, and output voltage, E_o, is shown in Fig. 2.4. The curve is linear over a range of 0 to 4 volts, the latter being defined as full-scale output voltage. Similarly, full-scale current is that current which produces a full-scale output voltage of 4 volts.

(b) The maximum sensitivity of the current detector is determined primarily by the quality of dynamic balance which can be obtained by selection of silicon diodes in the modulator. The slight imbalance inevitably remaining results in a small modulator output when no input current exists. By definition, the minimum detectable current must produce an output of the same order as this null value.

The maximum amplifier gain that can be employed is limited by the allowable null value at the dc output, which, in reference to bipolar probe use, has arbitrarily been chosen as 2% of full-scale E_o. Input current sensitivity presently attainable with 4-diode modulator (based upon 2% of full-scale null output) is approximately 1×10^{-7} amperes full scale, which corresponds to a minimum detectable current of 2 or 3×10^{-9} amperes. Improved sensitivity can be obtained by certain modulator modifications as described below in paragraph (c).

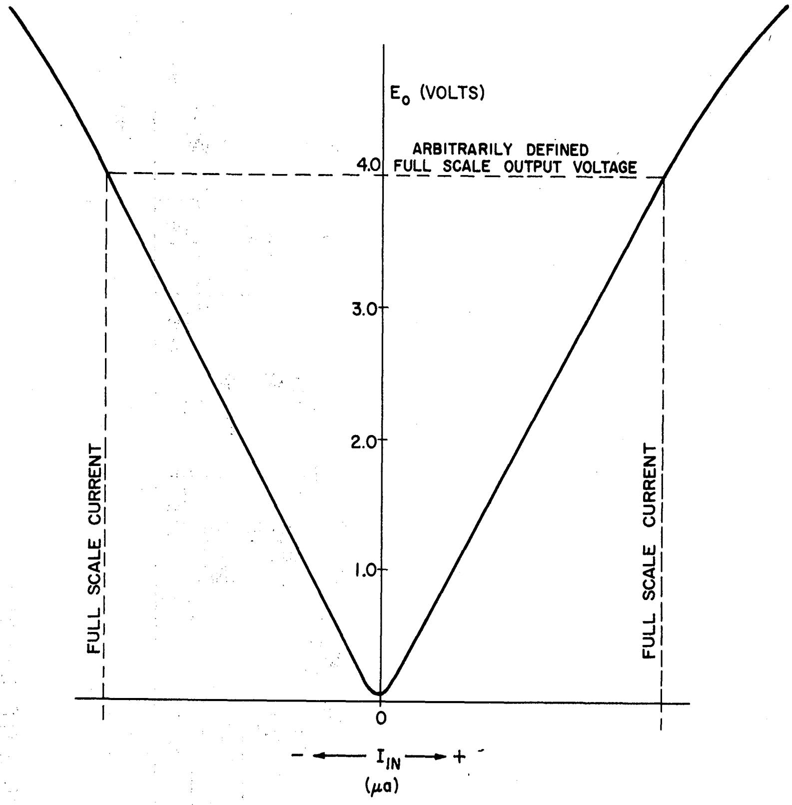
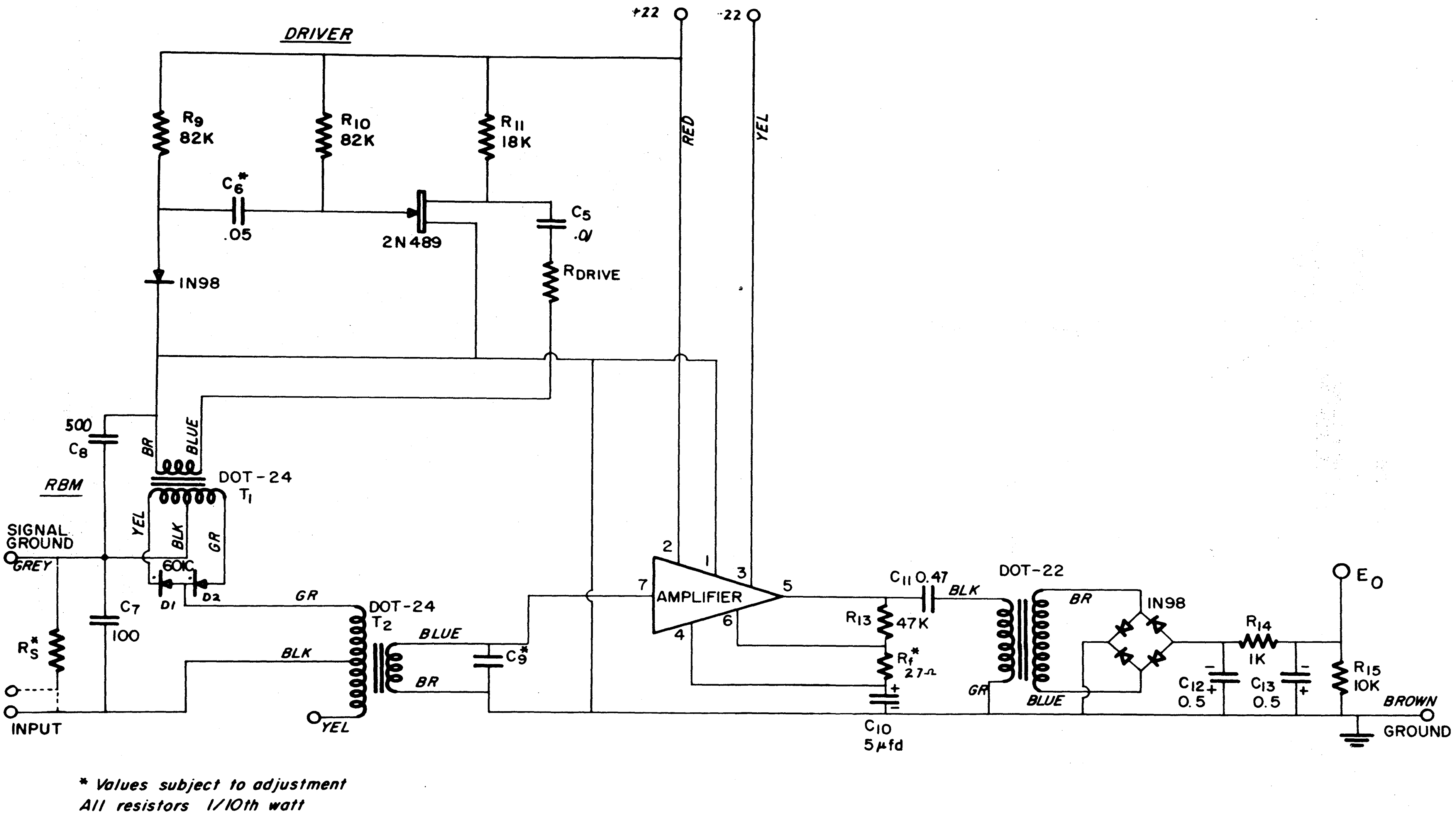


Fig. 2.4. Typical input-output curve of either the 2- or the 4-diode current detector.

(c) The 4-diode current detector system as shown in Fig. 2.2 can be modified slightly to obtain a much lower null level, and consequently greater ultimate sensitivity, provided the amplifier gain is increased accordingly. The modification involves replacement of the 4 diodes in the ring bridge by only 2 diodes, as shown in Fig. 2.5, and a reduction in reference current (increase R_{drive}). A nearly ideal dynamic balance can be obtained when only 2 diodes need to be matched, and the resulting null level can be decreased about two orders of magnitude, making possible correspondingly greater current sensitivity. The modified system permits higher amplification because of the reduction of the signal at null. The use of an additional transistor amplifier inserted at the modulator output has provided a full-scale current sensitivity of 1×10^{-9} amperes and minimum detectable current of 2 or 3×10^{-11} amperes.

(d) The input impedance of this particular ring-bridge modulator is about 50 k ohms and that of the 2-diode modulator is about 1 megohm. In general, both of these current detectors are adjusted initially for <u>maximum</u> sensitivity (as defined in paragraph (b) above), and then a suitable shunt resistor is applied at the input to obtain the <u>desired</u> sensitivity. This procedure results in a lower impedance detector which is naturally desirable in a current-measuring device. As an example, a typical 1-μa



PROBE CURRENT DETECTOR

Fig. 2.5. Two-diode detector circuit.

detector will have an input impedance of approximately 15 k ohms resulting in a 15-mv drop across the detector input or an input power of 1.5×10^{-10} watts.

(e) The detector responds nearly identically to both polarities of input current; the unsymmetry, which is less than 5%, can be minimized by adjustment of the reference current (R_{drive}).

(f) The frequency response of the detector to varying dc inputs is limited primarily by the low-pass RC filter at the rectifier output. The filter design is a compromise between the necessity of maintaining a low carrier ripple component on the dc output and the requirement of fast response to input current changes. The 2000-cps chopping rate is high enough to provide reasonably good reproduction of a 20-cps square wave input but the filter reduces this to about a 10-cps square wave and maintains the ripple level below 10 mv at full-scale output. If greater ripple is not objectionable, higher frequency response is possible. The chopping rate of 2000 cps was chosen because the response of T_1 and T_2 peak near this frequency results in maximum sensitivity of the modulator. Higher modulating frequencies requiring less filtering may be used at reduced sensitivity if faster response is required.

(g) The power supply batteries that are employed are specially assembled packs of RM-400 Mallory Mercury Cells (80 MAH) delivering approximately 2 ma at 22 volts (with zero modulator input current), thus providing an operational life of over 30 hours, amply adequate for typical setup, testing and flight.

(h) The input is isolated from circuit ground by T_1 and T_2 but is maintained at ac ground by C_8. C_7 reduces the 60-cycle and transient pickup at the detector input.

Figure 2.6 shows the etched circuit board used for both the 2- and the 4-diode detectors; Fig. 2.7 shows the completed detector after potting with EP Fome* to hold components rigidly in position.

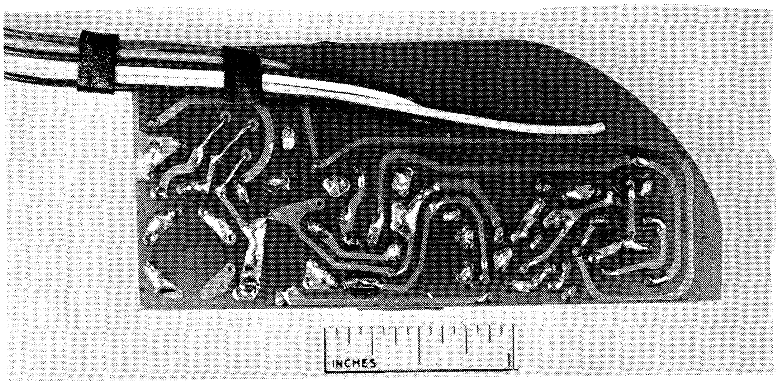


Fig. 2.6. Back of current detector etched circuit board.

*Electronic Plastic Corporation, Brooklyn 7, New York.

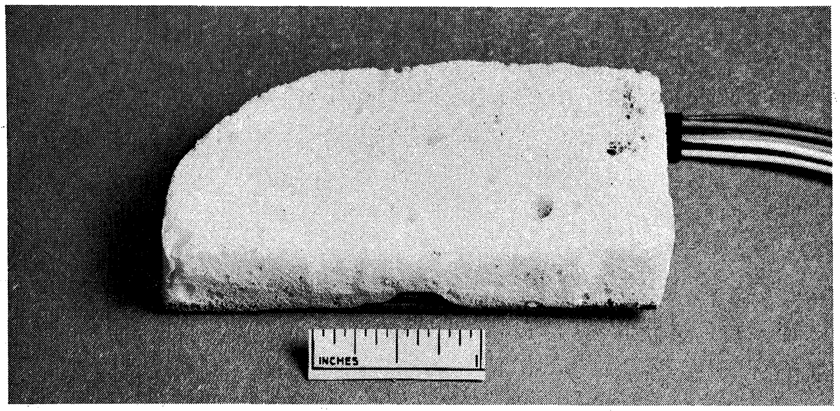


Fig. 2.7. Current detector after potting.

3. THE VOLTAGE-CONTROLLED OSCILLATOR

Several factors indicate the type of oscillator circuit best suited for voltage-controlled oscillator (VCO) use: (a) available space requires an oscillator which would occupy a volume less than approximately 0.5 cu. in.; (b) low current drain (< 5 ma) from the supply battery would make possible the use of batteries of relatively small volume; (c) a high input impedance is desirable to prevent possible loading of the driving source; (d) a linear relationship between frequency and input voltage would simplify data recovery; (e) the frequency must be relatively independent of ambient temperature and supply voltage changes, and (f) the required deviation of the subcarrier frequency is plus and minus 7-1/2% of the center frequency of the particular IRIG channel used.

A transistor multivibrator was chosen as it can be made to conform to these requirements by using silicon transistors and by carrying out some rather simple temperature compensation procedures. Also, the multivibrator is inherently dependable in starting, uses smaller components than an audio LC oscillator, and has a relatively low output impedance so that its frequency is relatively unaffected by the required load.

The multivibrator suffers from the disadvantage that its output is a square-wave which must be filtered in some manner as discussed later in this section.

OPERATION

Figure 3.1 shows the VCO schematic and Fig. 3.2 is a photograph of the etched circuit. The base bias level of silicon transistors Q_1 and Q_2 is set by R_3 and R_4,

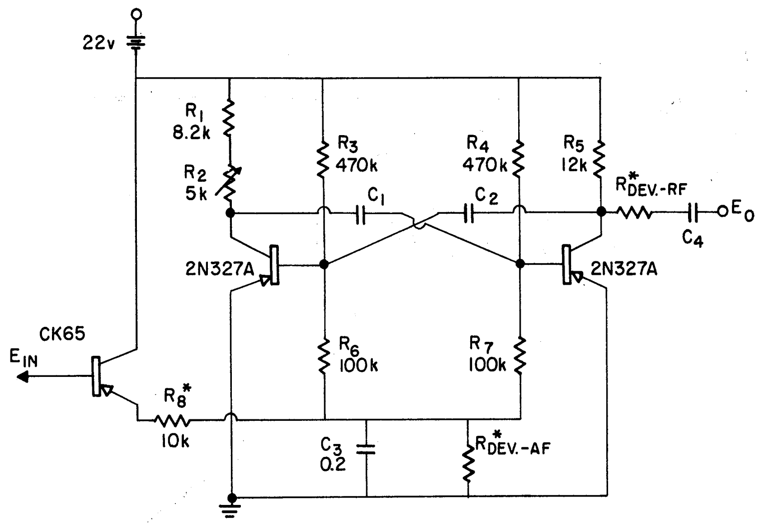


Fig. 3.1. VCO schematic.

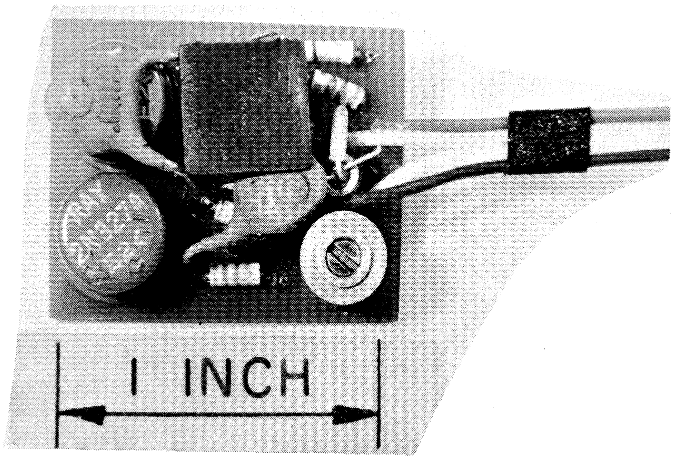


Fig. 3.2. Component view of VCO with unfiltered output.

respectively. By conventional multivibrator action, Q_1 and Q_2 alternately conduct and cut off, producing a square-wave output at each collector which is coupled to the output terminal through C_4 and $R_{dev\text{-}rf}$. Q_3 is used as an emitter-follower providing the VCO with a high impedance input (approximately 400 k ohm). The input signal, E_{in}, is applied at the base of Q_3 and appears at the emitter. The voltage divider R_8 and $R_{dev\text{-}af}$ applies a certain fraction of this as a change in bias of Q_1 and Q_2, causing a change in frequency of the VCO. Figure 3.3 shows a typical curve of oscillator frequency vs. input voltage. Note that the curve is nearly linear within the band and maintains reasonable linearity up to twice full-scale input voltage (approximately 8 volts).

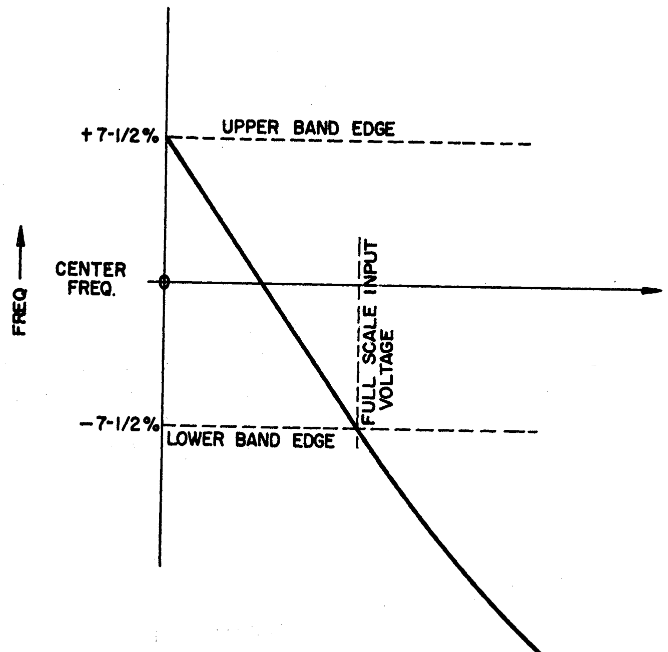


Fig. 3.3. Typical VCO frequency vs. input voltage curve.

Several important characteristics of the circuit are worthy of note.

(a) C_1 and C_2 are selected to satisfy two requirements. With the 'Trimpot' R_2, adjusted for mid-value, C_1 and C_2 are chosen such that the desired frequency is obtained (in this case, the upper-band-edge frequency for $E_{in} = 0$). C_1 and C_2 also must have a slightly negative temperature coefficient which just balances the frequency drift due to the effect of temperature on Q_1 and Q_2. This temperature compensation results in a frequency stability of about 1% over a range of 0° to 50°C.

(b) The frequency dependence upon supply voltage variation is approximately 1/2% per volt in Ebb, thus requiring the use of either a regulated supply for the VCO or a supply battery with a relatively flat discharge curve. The latter was used in early applications. For this purpose a special battery using 16 zinc-mercuric oxide RM-400 cells was assembled by Mallory Battery Corporation. The current drain is 2 ma at 21.6 volts so that the operational life is in excess of 30 hours (80 MAH cells). Alternatively, the supply source can be zener-diode-regulated output of a transistor power supply or some other regulated source.

(c) The IRIG Standards for Fm-Fm telemetry systems stipulates that the upper and lower band edge be ± 7-1/2% above and below the center frequency, respectively. This 15% frequency deviation for an input change of 4 volts is obtained by selection of the voltage divider resistor, R_{dev-af}.

(d) Optimum deviation of the RF carrier is achieved by providing the proper amplitude of VCO output at the transmitter signal input. This amplitude varies depending upon the particular transmitter but is normally about 1 volt peak to peak. Proper selection of R_{dev-rf} will provide the required amplitude.

(e) C_3 bypasses the bias drive point and reduces feedback from the oscillator through Q_3 to the input. C_3 must be optimized to provide sufficient bypassing and yet not impair the response of the VCO to rapid changes in input voltage.

The output waveform of the above VCO is a somewhat rounded squarewave (see Fig. 3.4). RC filtering of the output improves it somewhat but also causes reduction in output amplitude as the oscillator is swept from the low to the high end of the channel.

Two factors must be considered when using a VCO whose output waveform is not sinusoidal or nearly so.

(1) <u>The Telemetry Transmitter</u>.—The transmitter may be phase-modulated, PM-FM, or frequency-modulated, FM-FM. In the former case, the modulation is proportional to the rate of change of input voltage (VCO output); thus a squarewave in-

put results in a differentiated modulation, as depicted in Fig. 3.5. This differentiated waveform may produce full or optimum carrier deviation, but when viewed at the receiver possesses only a small fundamental component, making recording and discriminating functions more difficult. In the FM-FM system this becomes a less serious problem since the receiver records the square wave output which possesses a large fundamental component.

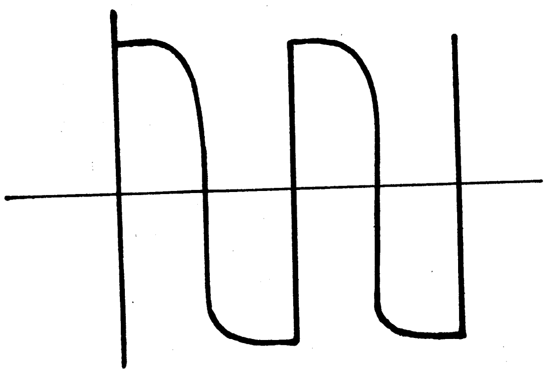


Fig. 3.4. Unfiltered output waveform of VCO.

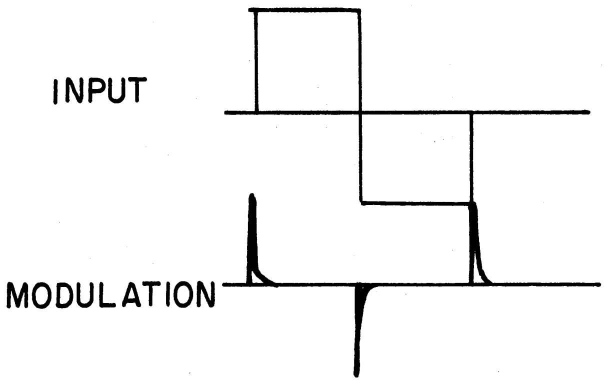


Fig. 3.5. Input to a PM-FM transmitter and the resulting modulation.

(2) Crosstalk Between Channels.—When the VCO output waveform from a given channel possesses harmonics which fall into the band of a higher channel, interference with the higher channel information may occur. As long as only two or three channels are used, the frequencies can be chosen such that crosstalk does not occur.

Thus the only justification for the use of nonsinusoidal VCO's (where this is feasible) is sheer reduction in size, which, however, can be significant. LC filtering tends to require bulky components; RC filtering, although physically small, is insufficient to satisfy fully the considerations discussed above.

To satisfy those applications in which a sinusoidal VCO is necessary, the oscillator was re-packaged and a filter employing capacitors and miniature audio chokes was added (see Figs. 3.6 and 3.7). Proper matching of the oscillator to the filter required the use of a grounded-collector transistor stage to provide the VCO with a low output impedance. This VCO has less than 5% harmonic distortion and has an output impedance of approximately 6 kΩ.

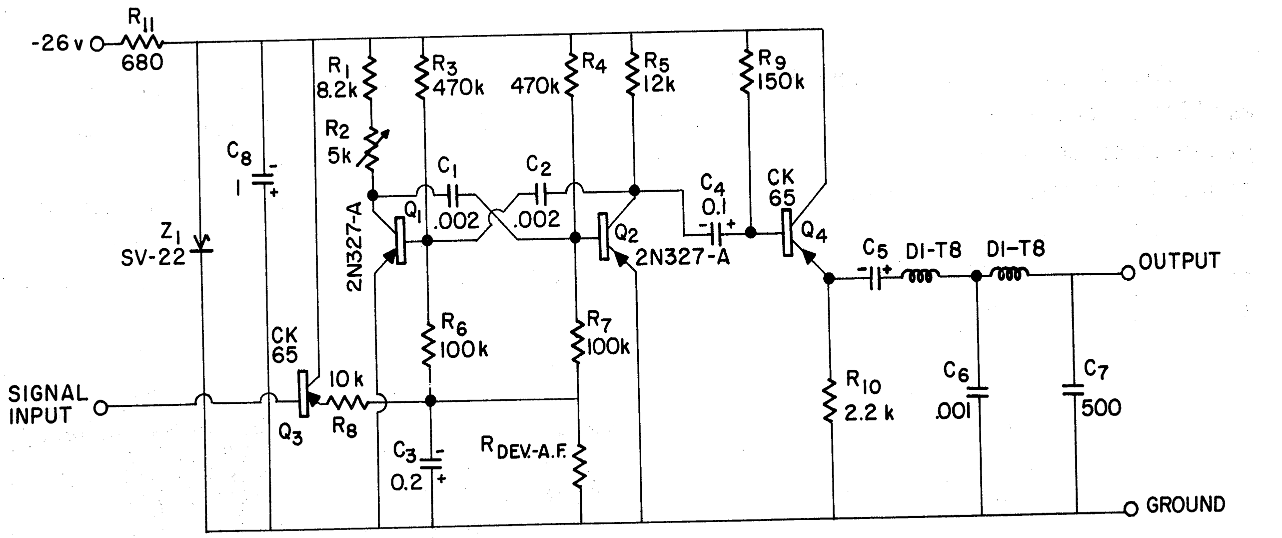


Fig. 3.6. VCO with LC filtered output.

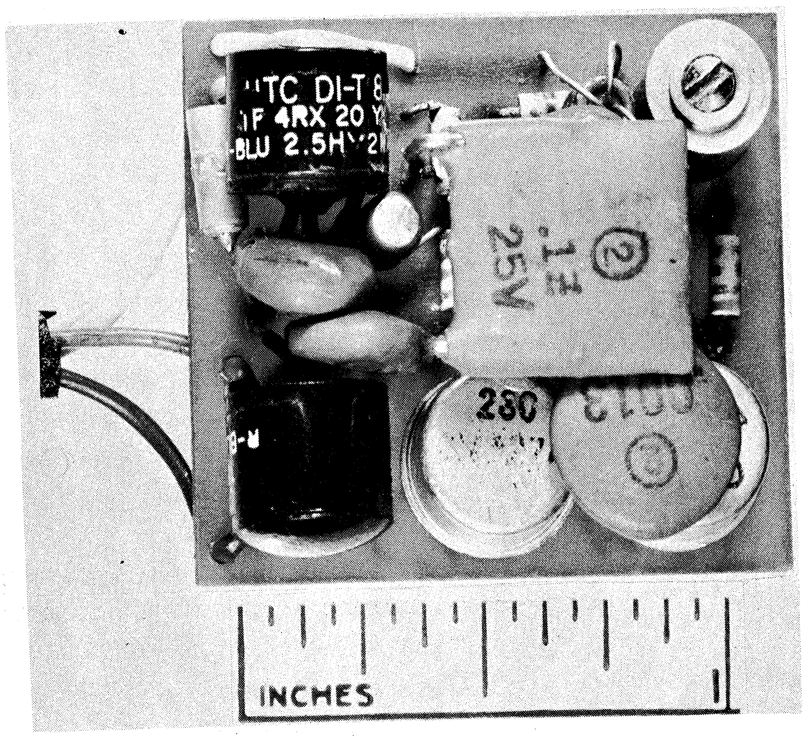


Fig. 3.7. VCO with LC filtered output.

4. RANGE-CHANGING CIRCUIT

The range-changing circuit is a high current-gain silicon semi-conductor amplifier designed to energize a relay when the circuit input voltage reaches a predetermined dc level called the switching point. In the case of the Langmuir probe instrumentation, the relay contacts are used in an appropriate manner to add a shunting resistor at the input of the current detector and thus change its current sensitivity. Figure 4.1 is the circuit schematic and Fig. 4.2 is a photograph of the circuit.

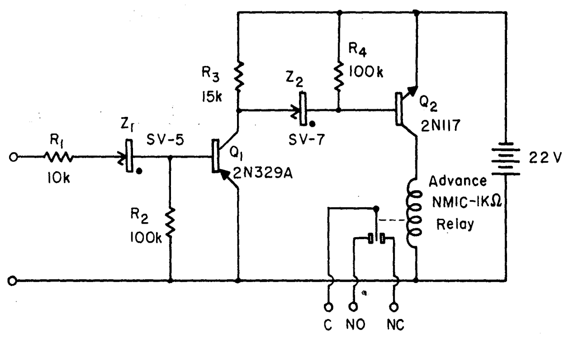


Fig. 4.1. Range-switching circuit.

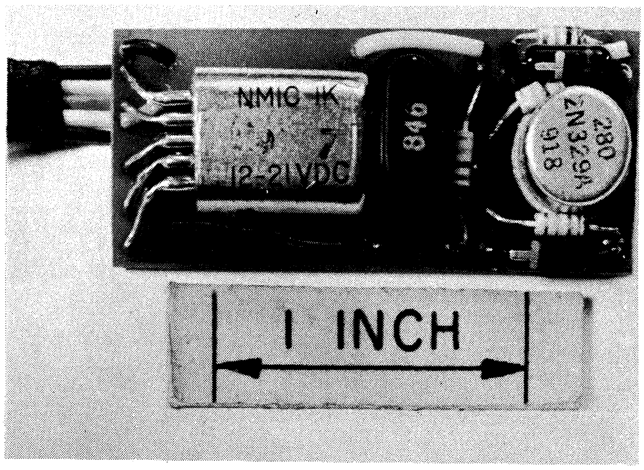


Fig. 4.2. Photograph of circuit board.

The switching operation may be described as follows. When the input voltage is below the zener region of reference diode Z_1, virtually no base or collector current flows in Q_1. Then the collector of Q_1, and the base and emitter of Q_2 all reside at $-E_{bb}$, resulting in no current through the relay.

As the input voltage approaches the switching point, Z_1 approaches its breakdown voltage and begins to allow Q_1 base current. The resulting collector current drops the collector voltage and Z_2 approaches its zener region allowing Q_2 base current and consequently collector or relay current. At the switching point, relay current reaches its pull-in value and switching occurs.

Figure 4.3 shows a typical curve of relay current vs. input voltage. The steep slope insures an accurately defined switching point. The relay current levels off at about 20 ma (over twice the pull-in value) when the limited supply voltage no longer allows Q_2 collector current to follow base current increases.

Some important characteristics of the switching circuit are discussed below.

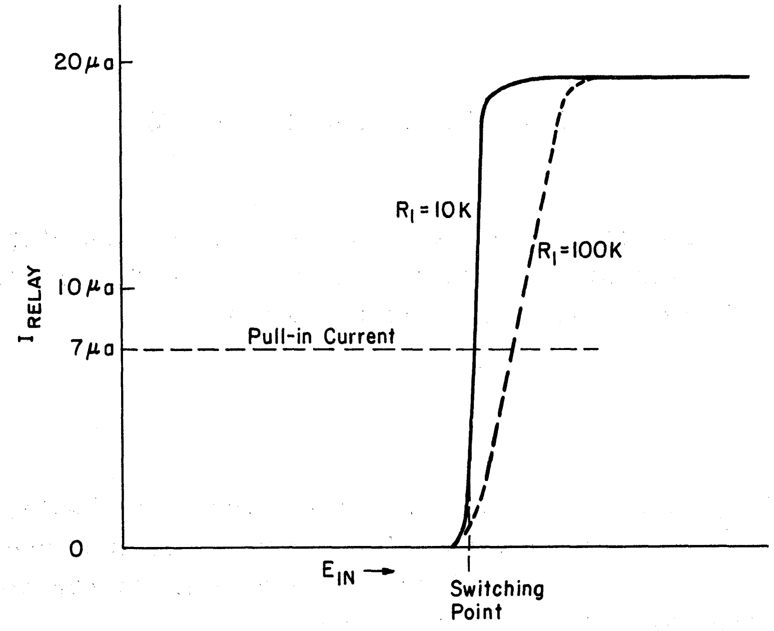


Fig. 4.3. Input voltage vs. relay current for range-switching circuit.

The switching point is fixed by selecting Z_1 so that it begins to break down at the desired switching point. For example, a 5.0-volt zener diode provides a switching point of approximately 4.0 volts which is stable to within ± 1% from 0° to 80°C, assuming the supply source maintains normal terminal voltage at the extremes of this temperature range. This stability is made possible primarily by the use of silicon rather than germanium transistors, which tend to have high I_{co} and avalanche more readily at higher temperatures.

Current drawn from the driving (input) voltage source is about 10 ua at the switching point (7 ma relay current), indicating a current gain of about 700 through the circuit. Above the switching point, the current drawn from the input depends upon the value of R_1 and the input impedance of the Q_1 amplifier stage. The input impedance is 4 k ohms and R_1 = 10 k, making the total input impedance 14 k to all increases in voltage above the switching point. Larger values of R_1 can be used if decreased loading of the source is desired; however, this has the same effect as decreasing the current gain of the amplifiers Q_1 and Q_2, resulting in a less steep input voltage vs. I_{relay} characteristics (Fig. 4.3) and less switching point stability. The largest practical value of R_1 would be approximately 100 k, which would result in a curve somewhat like that shown dotted on Fig. 4.3.

Below the switching point, the current from the power supply is less than 1 ua, making supply voltage on-off control unnecessary.

The circuit shown in Fig. 4.1 operates on a negative input voltage but can be made to operate on a positive input by interchanging Q_1 and Q_2, reversing the polarities of Z_1 and Z_2, and reversing the supply voltage polarity.

5. SAWTOOTH VOLTAGE GENERATORS

Two different sawtooth voltage generators have been developed for use in Langmuir probe instrumentation. These make possible different output waveforms and have different power supply requirements.

GENERATOR NO. 1

Figures 5.1, 5.2, and 5.3 illustrate the first device which supplies a clipped sawtooth output waveform, sweeping through a maximum range of from -7 volts to +7 volts with an output voltage divider arranged for probe purposes to provide two amplitudes, ± 2 volts and ± 6 volts.

The circuit operation is outlined as follows. The emitter output of a unijunction oscillator,[2,3] Q1, provides a sawtooth waveform at a frequency determined primarily by C1 and R4. This output is amplified by Q2 and is directly coupled to the emitter follower stage Q3 which provides a low impedance output. Zener diode, Z_1, establishes the zero reference such that the output waveform varies equally above and below zero with the emitter of Z_1 taken as zero reference.

The sawtooth output is clipped at both extremes to provide known reference levels and to maintain amplitude stability independent of supply voltage and temperature changes. The clipping circuit uses a silicon zener diode for each output polarity and a germanium diode for each zener diode to prevent shorting of the output by forward-conduction of the zener. This provides the output waveform shown in Fig. 5.4.

Several important characteristics are noted below.

1. The frequency is approximately 1 cps for C_1 equal to 33 μfd.

2. Linearity tolerance of the straight-line portion of the output is less than 5%.

3. Temperature changes tend to cause the waveform to shift upward or downward slightly, thus changing the length of the upper or the lower reference level but maintaining the same slope and reference-to-reference amplitude.

4. The output impedance of the generator is about 2 k ohms plus the resistance added by that part of the divider which is in series with the output.

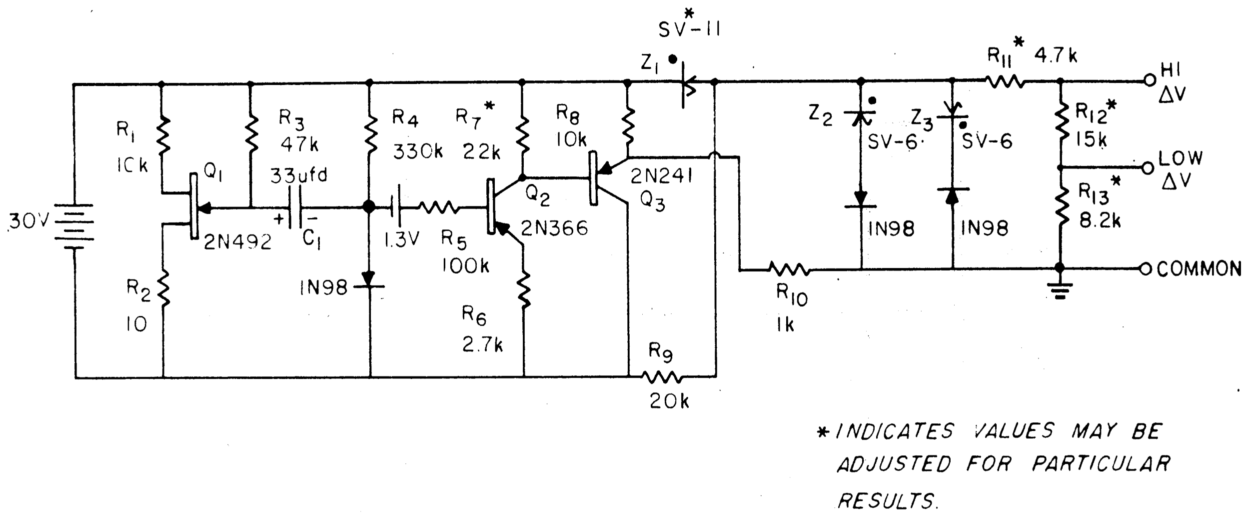


Fig. 5.1. No. 1 sawtooth voltage generator

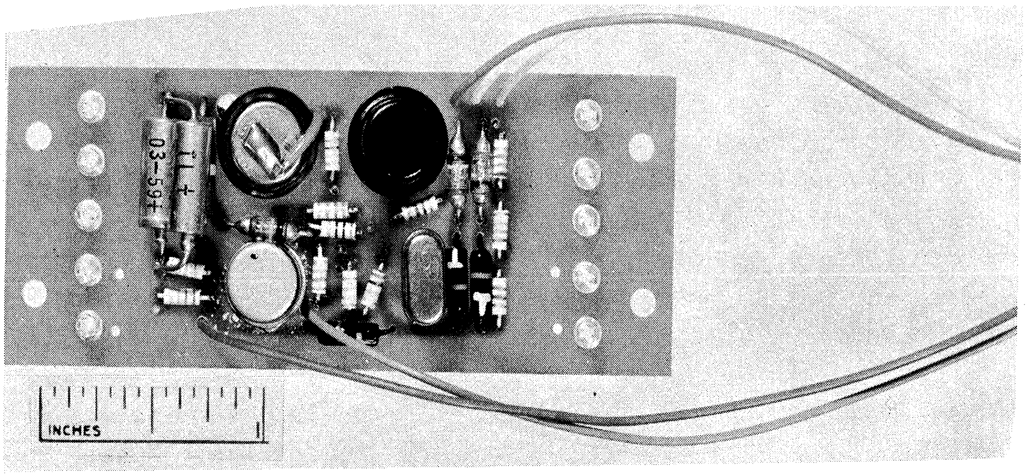


Fig. 5.2. Photograph of No. 1 generator.

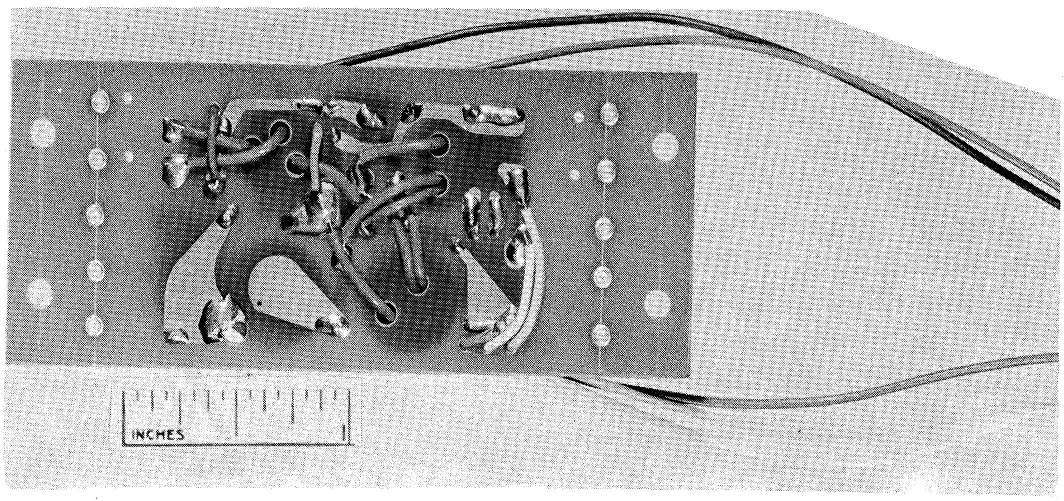


Fig. 5.3. Photograph of No. 1 generator (etched board view).

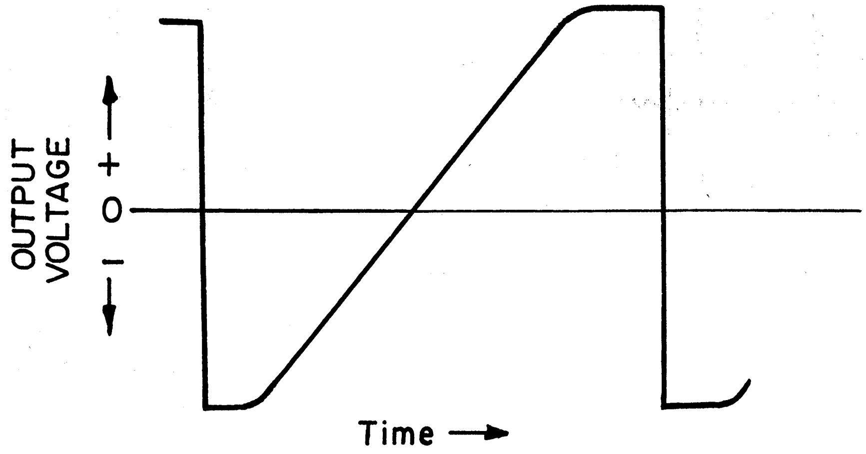


Fig. 5.4. Typical output from No. 1 generator.

GENERATOR NO. 2

A second generator accomplishes the more difficult task of producing a pure sawtooth output having extremely good linearity, amplitude stability, and freedom from drift of the waveform about the zero reference. Furthermore, if the output is to be symmetrical about ground, this circuit may use supplies with a common ground, eliminating the need for separate batteries as required on the first sawtooth voltage generator discussed earlier. This improved performance is obtained by the use of the more expensive silicon semiconductor devices and by requiring both a negative and a positive supply source.

Good amplitude stability (eliminating the need for clipping) is the result of two features of the circuit. (1) The supply voltages are zener-diode-regulated to reduce bias level changes due to drift in the source voltage. (2) A silicon transistor and a thermistor used in the amplifier stage reduce the drift in bias level due to temperature changes.

Operation of the circuit (Fig. 5.5) is as follows. The sawtooth output of Q_1 used in a unijunction-oscillator circuit is directly coupled to the emitter follower stage Q_3, which isolates the oscillator from the low input impedance of amplifier stage Q_4. Emitter follower stage Q_5 presents a low output impedance to the load. Q_2 is a PNP silicon transistor, which acts as a constant current generator in charging C_1, resulting in a more linear sawtooth output. Zener-diode regulators, Z_2 and Z_3, are selected so that their difference in zener voltage provides sufficient bias for Q_2 (and therefore for Q_3) to allow maximum swing of output without significant distortion of the output sawtooth. Z_1 is then selected for amplitude symmetry of the output wave about the ground level or zero reference.

Thermistor T is selected to minimize changes in output amplitude with temperature change. Amplitude stability better than 1% can be obtained through the range of 0°C to 80°C.

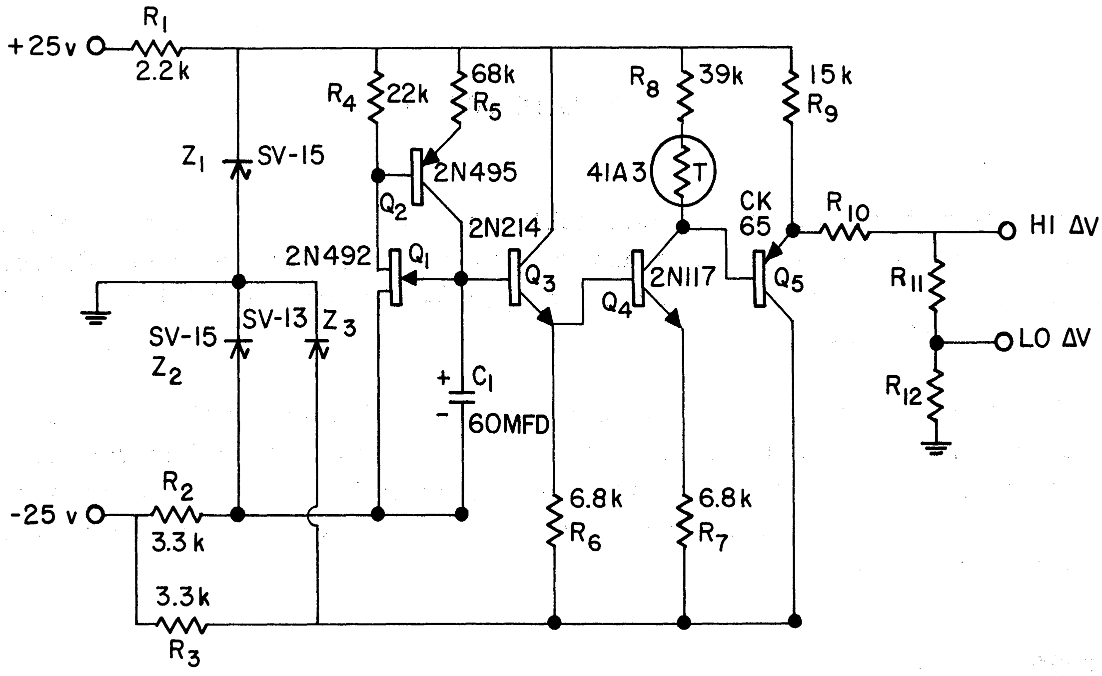


Fig. 5.5. No. 2 sawtooth voltage generator.

The maximum output amplitude is about 25 volts peak to peak, or ± 12-1/2 volts with respect to ground. R_7 and R_8 are adjusted to provide necessary voltage amplification for a given silicon transistor Q_3. Supply voltage changes of + 25% to -25% cause less than 1% amplitude change and less than 1% zero drift.

A typical output waveform is shown in Fig. 5.6. The linearity tolerance of the device is reduced to less than 2% by use of the silicon transistor Q_2 mentioned previously.

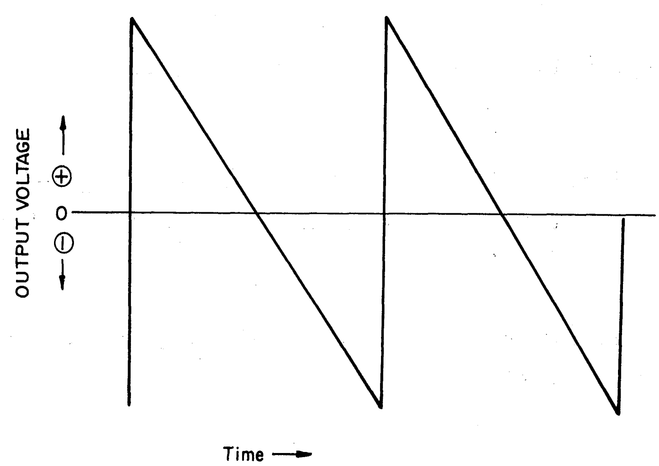


Fig. 5.6. Typical output of No. 2 sawtooth generator.

6. DC TO DC CONVERTER

The converter employs a common collector configuration and a toroid transformer, and in probe use is intended to supply through use of zener diodes several regulated 22-volt outputs from a single 6-volt source consisting of four zinc-silver oxide cells.

The circuit shown in Fig. 6.1 employs two half-wave rectifiers and RC filters supplying plus and minus 40 volts at 40 ma. The required number of regulated outputs are then obtained from these sources by means of separate zener diodes. Further filtering if necessary, is achieved by the 1-ufd capacitors across each regulated output.

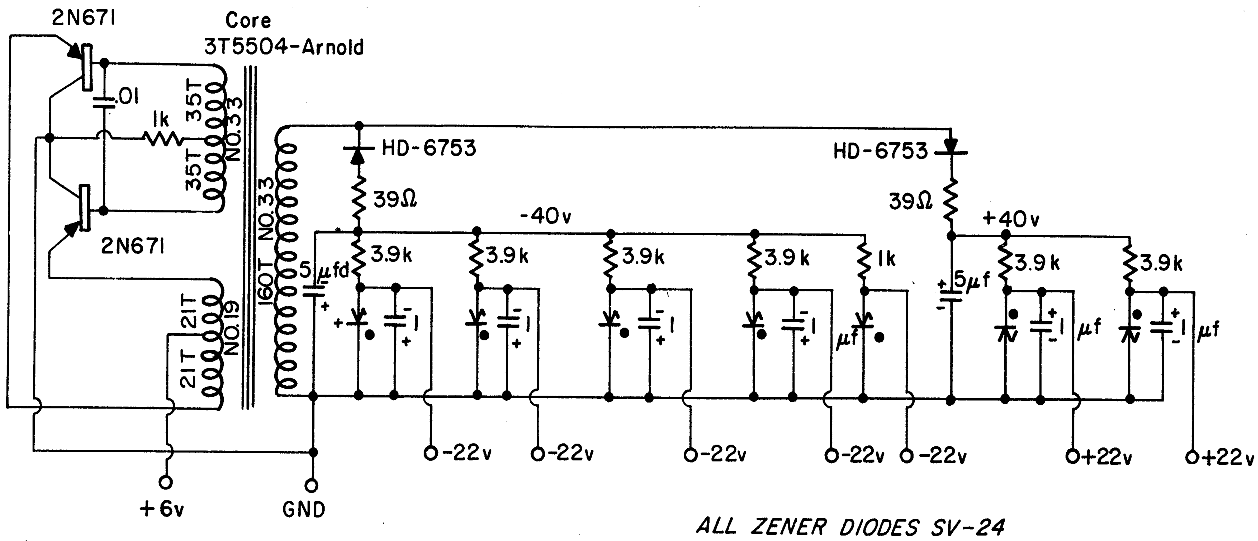


Fig. 6.1. Converter and regulators.

Under full load conditions (40 v at 40 ma, -40 v at 40 ma) the output power is 3.2 watts, while the input power is 4.2 watts (6 volts at 700 ma), resulting in a conversion efficiency of 75%.

Figures 6.2, 6.3, and 6.4 are photographs of three views of the converter. The regulator board is not shown.

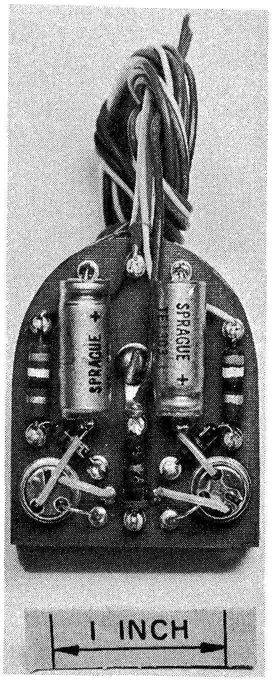


Fig. 6.2. Converter (front view).

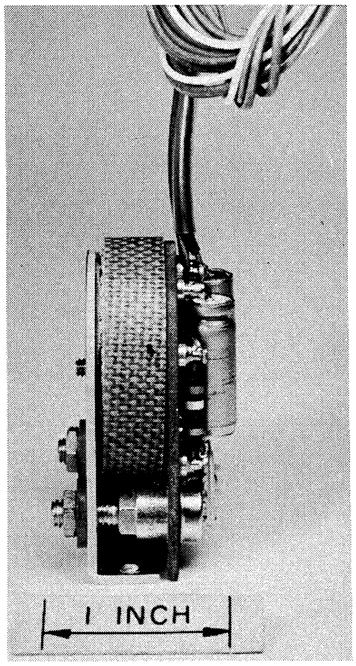


Fig. 6.3. Converter (side view).

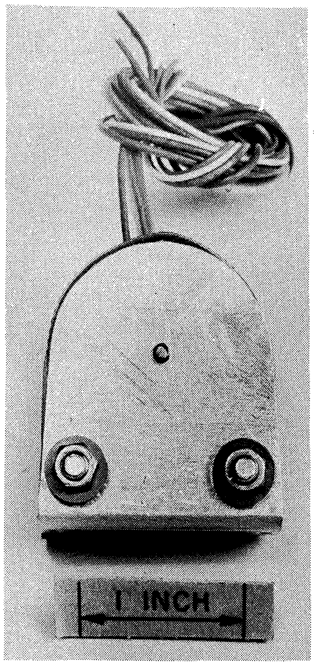


Fig. 6.4. Converter (rear view).

7. TIMER

The timer circuit shown in Fig. 7.1, uses a unijunction transistor[1,2] to time and trigger two transistors in a flip-flop multivibrator circuit in which the collector load of one transistor has been replaced by a miniature relay whose contacts are used to initiate or interrupt the operation of other circuits. The on and off times may be controlled independently by suitable adjustment of R_4 and R_6. Operating from 12 V or 24 V, periods of a few milliseconds to about 50 seconds may be obtained. A complete discussion of the operation of the timer may be found in Refs. 3 and 4.

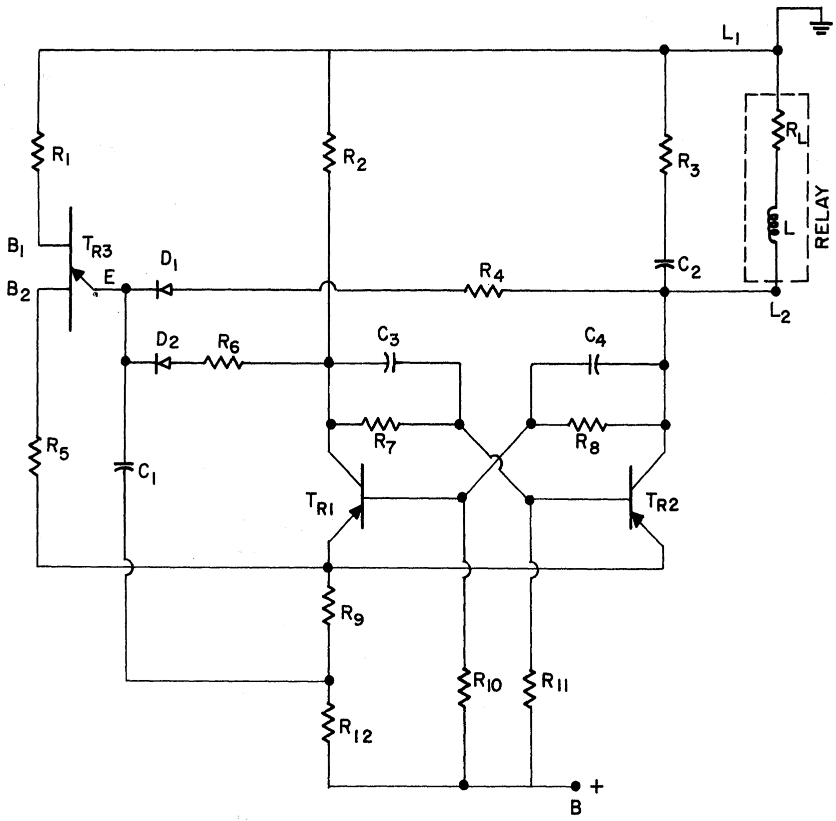


Fig. 7.1. Timer.

The timer has been packaged in two etched circuit configurations. Figures 7.2 and 7.3 show front and back views of the original packaging. Figures 7.4 and 7.5 show two views of a later version in which two timers were placed on one circuit board.

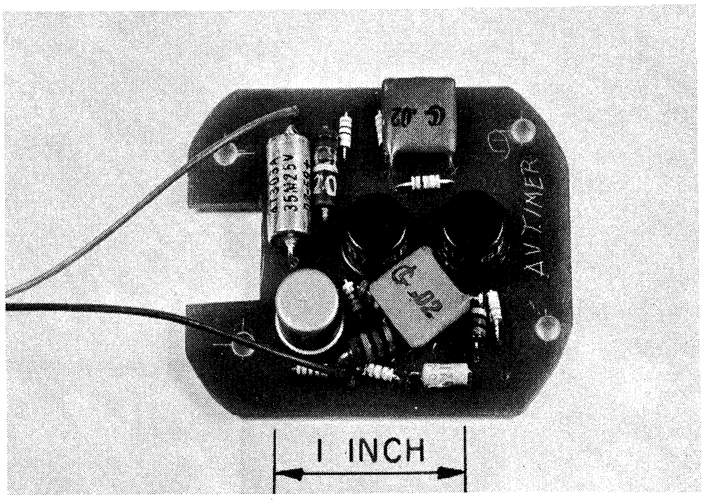


Fig. 7:2. Timer (front view).

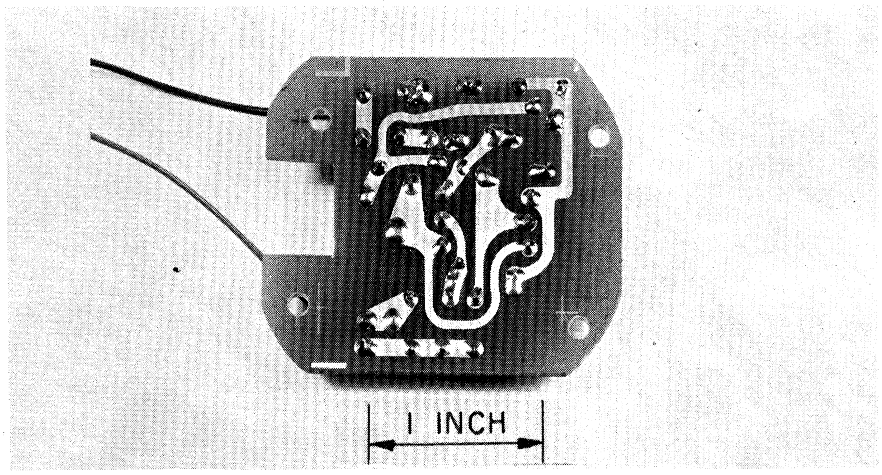


Fig. 7.3. Timer (rear view).

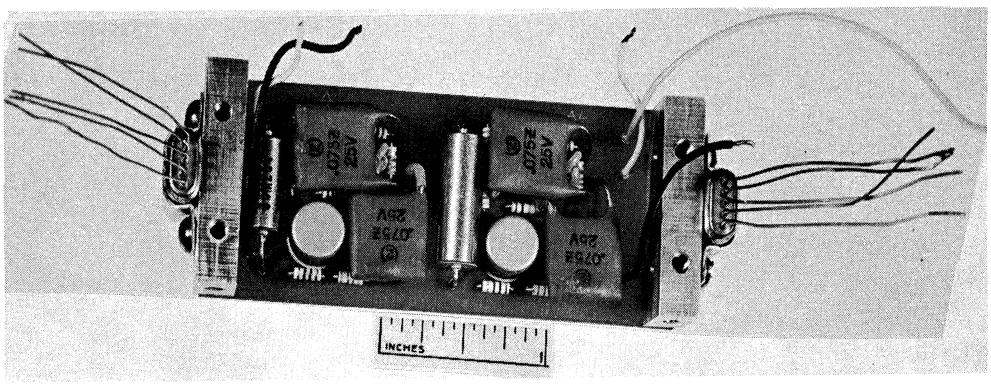


Fig. 7.4. Two timers on single board.

8. CIRCUITS BEING DEVELOPED

Several devices are being developed for use with future instrumentations and are reported here to indicate work in progress.

IONIZATION GAGE EMISSION REGULATOR

This transistorized circuit will maintain a constant predetermined emission current by regulation of the gage filament voltage. It may be adjusted to provide 2 to 10 volts at 1 to 5 amperes and operates from a 12-volt dc source.

SWEEP FREQUENCY OSCILLATOR

This transistorized oscillator is arranged so that its frequency is swept linearly in four steps, over the range of 100 kc to 4.5 mc. Voltage-controlled capacitors are employed as the frequency sweeping elements.

VOLTAGE STORAGE UNIT

This device, intended for the short-time storage of periodically and slowly varying dc potentials, utilizes a 30 segment (nonshorting) motor-driven (1 cps) commutator. During the read-in time, the commutator wiper completes one revolution, and a 0.5 Mfd ceramic capacitor on each segment is charged from the information source to the signal level present at the time of wiper passage and is thus stored. Thus the function to be stored is represented by 30 voltage levels corresponding to the signal level at the time the wiper last passed that particular segment.

During read-out, the wiper is disconnected from the information source and is connected to the high impedance input of a silicon transistor grounded collector stage. The stored voltage output is then taken at the emitter of this stage, and may be fed as a series of pulses to, for example, a voltage-controlled oscillator.

RANGE-CHANGING CIRCUIT

This circuit may be used to change range resistors (sensitivity) automatically of, for example, a current measuring system so that the output is maintained between a predetermined upper and lower bound. When the output reaches one of these bounds, a transistorized circuit is triggered and discharges an electrolytic capacitor into one of two Ledex rotary solenoids mounted back to back on a common shaft with a wafer switch which has a number of range resistors connected to it in such a manner that switching in one direction increases the value of the range resistor and switching in the other direction decreases it. This capacitor discharge causes solenoid action and the wafer is stepped in the proper direction to insert a different range resistor which has the correct value to bring the system output back within bounds.

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