RADIOTRONICS

IN THIS ISSUE

COVER

A ceramic Tetrode with Integral Boiler manufactured by the English Electric Valve Co. Pty. Ltd.

Vol. 31, No. 3 July, 1966

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AN AWV PUBLICATION

REGISTERED IN AUSTRALIA FOR TRANSMISSION BY POST AS A PERIODICAL

INTEGRATED CIRCUIT APPLICATION NOTE

Application of the RCA-CA3000 Integrated-Circuit DC Amplifier

The RCA-CA3000 dc amplifier is a monolithic silicon integrated circuit supplied in a 10-terminal TO-5 package. This stabilised and compensated differential amplifier has push-pull outputs, high-impedance (0.1-megohm) inputs, and gain of approximately 30 dB at frequencies up to one megacycle per second. Its useful frequency response can be increased to several tens of megacycles per second by the use of external resistors of coils.

Because full gain-control capability is inherent in the CA3000, it can be used as a signal switch (with pedestal), a squelchable audio amplifier (with suppressed switching transient), a modulator, a mixer, or a product detector. When suitable external components are added, it can also be used as an oscillator, a one-shot multivibrator, or a trigger with controllable hysteresis. Within its specified frequency range, it is an excellent limiter, and can handle input signals up to about 80 millivolts rms before significant cross-modulation or intermodulation products are generated.

CIRCUIT DESCRIPTION

The circuit diagram and terminal connections for the CA3000 dc amplifier are shown in Fig. 1. The circuit is basically a single-stage differential amplifier (Q_2 and Q_4) with input emitter-followers (Q_1 and Q_5) and a constant-current sink (Q_3) in the emitter-coupled leg. Push-pull input and output capabilites are inherent in the differential configuration.

The use of degenerative resistors R_4 and R_5 in the emitter-coupled pair increases the linearity of the circuit and decreases its gain. The low-frequency output impedance between each output (terminals 8 and 10) and ground is essentially the value of the collector resistors R_1 and R_2 in the differential stage.

OPERATION OF CIRCUIT

The CA3000 is designed for operation from a wide range of supply voltages. Operation from either one or two power supplies is feasible, as illustrated by the typical biasing techniques shown in Fig. 2. However, operation from two supplies is recommended because fewer external bias networks are required and, therefore, less power is consumed.

The maximum voltage that can be applied across the voltage V_{CC} plus negative supply voltage circuit (positive supply V_{EE}) is 16 volts. The maximum voltage capability (V_{CE}) of the differential pair is limited to 8 volts. Extra care must be used to ensure that these values are not exceeded when the circuit is used to drive inductive loads.

The operating-current conditions of the differential pair are determined by the base-bias circuit and emitter resistance of the emitter-coupled constant-current sink (Q_3), as well as by the voltage between terminals 2 and 3. Each possible current condition is manifested by (1) a distinct set of dc operating characteristics with differing temperature characteristics, (2) a particular value of gain having its own temperature dependence, and (3) a particular dynamic output-voltage capability. For each value of voltage between terminals 2 and 3 (V_{EE} when terminal 2 is grounded), there are four possible operating modes, as described in Table I.

Table I—Operating Modes for CA3000 DC Amplifier

Mode	*Shorted Terminals*	*Condition of Diodes*	*Q_3 Emitter Resistor*
A	none	in	R_9+R_{10}
B	5-3	out	R_9+R_{10}
C	4-3	in	R_9
D	5-4-3	out	R_9

The operating characteristics for these modes of operation are summarised in Table II for various two-supply configurations with terminal 2 grounded and with V_{EE} values of —3 and —6 volts dc.

Table II shows that the positive supply voltage can be adjusted for each mode of operation and for each value of negative supply so that the nominal dc output

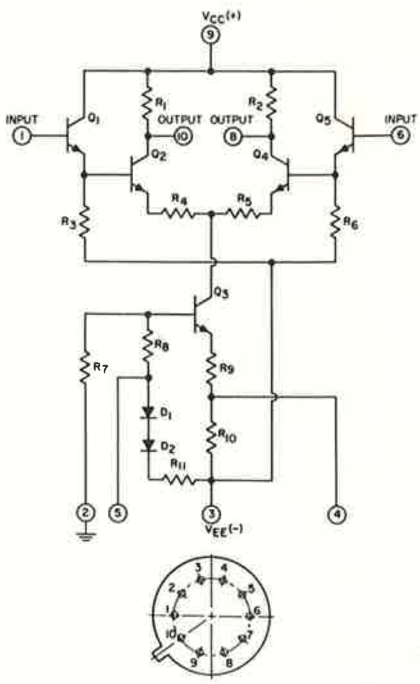


Fig. 1— Schematic diagram and terminal connections for the CA3000 integrated-circuit DC amplifier.

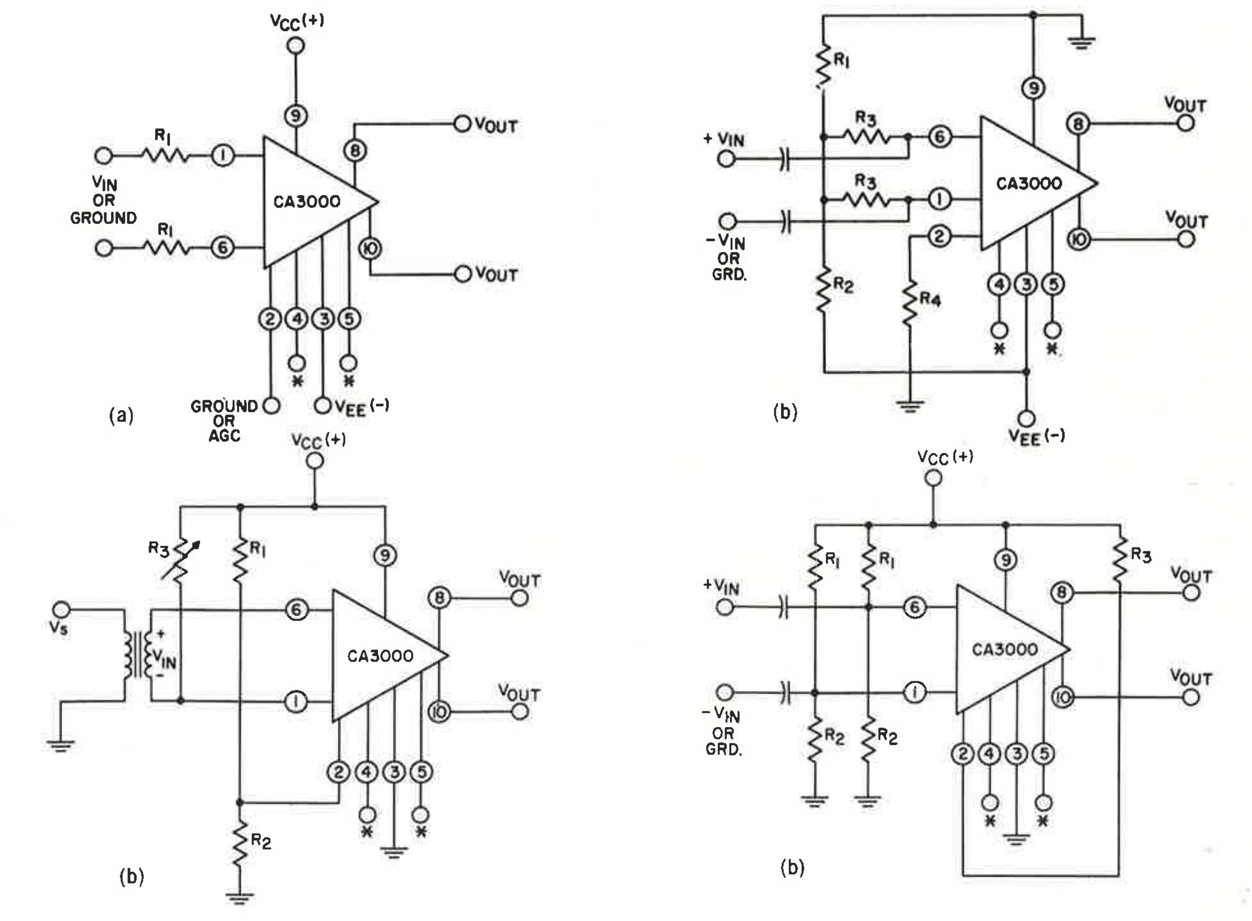
Table II—Design Characteristics of CA3000 Operating Modes

DC Supply Volts Positive V_{cc}	DC Supply Volts Negative $-V_{EE}$	Operating mode	Single-ended midband voltage gain — dB G_{VS}	DC output volts (Terminal 8 or 10 to ground) V_{odc}	Positive voltage swing V_{omax}*	Negative voltage swing V_{omin}*	Total power dissipation — mW
6	−6	A	31.2	+2.3	+3.7	−3.8	40
6	−6	B	27.3	+4.3	+1.7	−5.7	36
6	−6	C	34.6	−1.5 (saturated)	+7.5	0	61
6	−6	D	32.4	+1.0	+5.0	−2.4	47
3.7	−6	A	31.2	0	+3.7	−1.4	33
1.7	−6	B	27.3	0	+1.7	−1.4	25
10.6 (over rating)	−6	C	34.6	0	+10.6	−1.5	83
5.0	−6	D	32.4	0	+5.0	−1.5	43
3	−3	A	27.5	+1.2	+1.8	−2.6	8.8
3	−3	B	16.6	+2.6	+0.4	−4.1	7.4
3	−3	C	32.6	−1.5 (saturated)	+4.5	0	14
3	−3	D	24.4	+1.9	+1.1	−3.3	8.5
1.8	−3	A	27.5	0	+1.8	−1.5	7.2
0.4	−3	B	16.6	0	+0.4	−1.5	8.4
5.3	−3	C	32.6	0	+5.3	−1.5	19
1.1	−3	D	24.4	0	+1.1	−2.6	6.2

* V_{omax} and V_{omin} are the ac swing extremities above and below V_{odc}.

voltage is zero. (Although the V_{CC} value required for mode C for a V_{EE} of −6 volts dc is in excess of the maximum rating, operation within ratings can be achieved with slightly negative values of output voltage.) The use of these adjusted values of positive supply provides two advantages: (1) direct interstage coupling can be effected in a single-ended configuration, and (2) negative feedback can be introduced from a single output back to the appropriate input. For low-level applications in mode D with a negative supply voltage V_{EE} of −3 volts dc and a positive supply voltage V_{CC} of 1.1 volts dc, the CA3000 has a gain of 24.4 dB, a dissipation of 6.2 milliwatts, an output capability of 2.2 volts peak-to-peak, and a dc output-voltage reference level of zero.

The information in Table II can be modified for single-supply designs by simple addition and/or subtraction of dc



* Connection of terminals 4 and 5 depends on mode of operation.

Fig. 2—Typical biasing arrangements for the CA3000 for operation from (a) two separate voltage supplies, or (b) a single voltage supply.

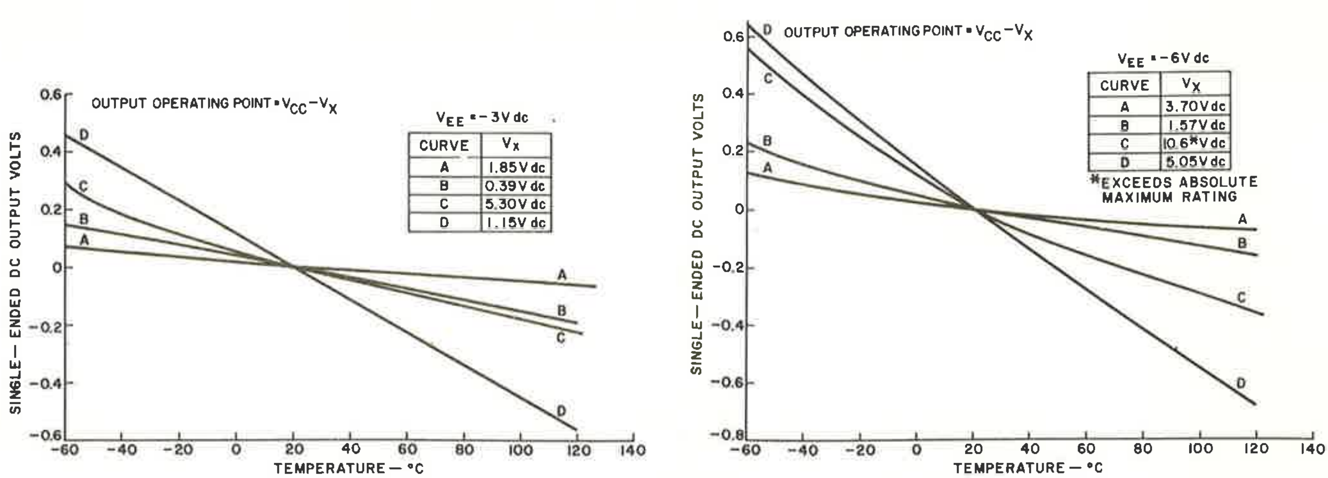


Fig. 3—Theoretical curves of dc output voltage as a function of temperature for negative-supply voltages of —3 and —6 volts dc (calculated for $\beta=35$ at 20°C).

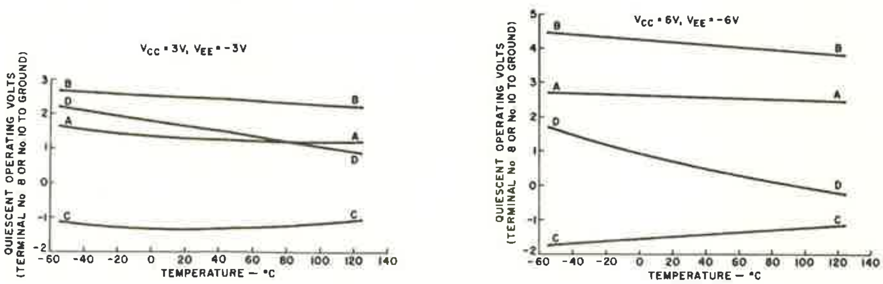


Fig. 4—Measured curves of DC output voltage as a function of temperature for negative-supply voltage of —3 and —6 volts DC.

values. For example, the correct information for a single supply of 12 volts dc for operating mode A can be obtained from the conditions shown in the table for mode A for $V_{CC}=6$ Vdc and $V_{EE}=-6$ Vdc by the addition of 6 volts to the values shown for V_{CC}, V_{EE}, Vo_{dc}, Vo_{max}, and Vo_{min}. (It should be noted that the required voltage levels at the input terminals 1 and 6 and at terminal 2 are also 6 volts higher.

As mentioned previously, the four operating modes exhibit different temperature characteristics. Fig. 3 shows theoretical curves of dc output voltage as a function of temperature for each operating mode for negative supply voltages V_{EE} of —3 and —6 volts dc. The experimental curves shown in Fig. 4 are in excellent agreement with the theoretical curves except in the case of mode C. In this mode, the differential-pair transistors Q_2 and Q_4 were driven into saturation as a result of the use of symmetrical supplies ($V_{CC}=V_{EE}$) for the experimental data. The discrepancy could be corrected by use of somewhat higher values of positive supply voltage

Fig. 5 shows theoretical curves of gain as a function of temperature for the four operating modes with V_{EE} values of —3 and —6 volts dc. With the diodes in (modes A and C), the gain decreases for both values of V_{EE}. With the diodes out (modes B and D), on the other hand, the gain increases with temperature for a negative supply of —3 volts dc, but decreases with temperature for a negative supply of —6 volts dc. With the diodes out, there is a value of negative supply (approximately —4.5 volts dc) for which the gain is independent of temperature. Fig. 6 shows measured values of single-ended and push-pull gain for mode A with symmetrical power supplies of ±6 volts dc. (This configuration is used in the remaining discussion because it provides the maximum sinusoidal output capability, as shown in Table II, and because of the convenience of ±6-volt dc supplies.)

The typical single-ended voltage-gain/frequency-response curve of the CA3000 for dc supplies of ±6 volts in operating mode A is shown in Fig. 7, together with the test circuit used for voltage-gain measurements. The Mode responses of the CA3000 are virtually independent of source impedance up to 10,000 ohms because of the emitter-follower inputs. The curves in Fig. 8 show that gain and bandwidth are virtually independent of temperature for operation in mode A with ±6-volt dc supplies.

Fig. 9 shows agc characteristics for the CA3000 for an input frequency of one kilocycle per second, together with the agc voltage-gain test circuit. When the agc voltage at terminal 2 is varied from 0 to —6 volts, the amplifier gain can be varied over a range of 90 dB.

Fig. 10 shows the test circuit used to measure common-mode rejection, together with curves of common-mode rejection as a function of frequency and temperature. Typical rejection is 97 dB at a frequency of one kilocycle per second. Fig. 11 shows the test circuit used to measure the dc unbalance of the amplifier (referred to the input), together with a curve of the input offset voltage as a function of temperature. Typical input offset voltage (with an assumed push-pull differential gain of 37 dB) is 1.5 millivolts. Fig. 12 shows curves of input bias current, input impedance, and dynamic output voltage as functions of temperature.

APPLICATIONS

Crystal Oscillator—The CA3000 can be used as a crystal oscillator at frequencies up to one megacycle per second by connection of a crystal between terminals 8 and 1 and use of two external resistors, as shown in Fig. 13(a).

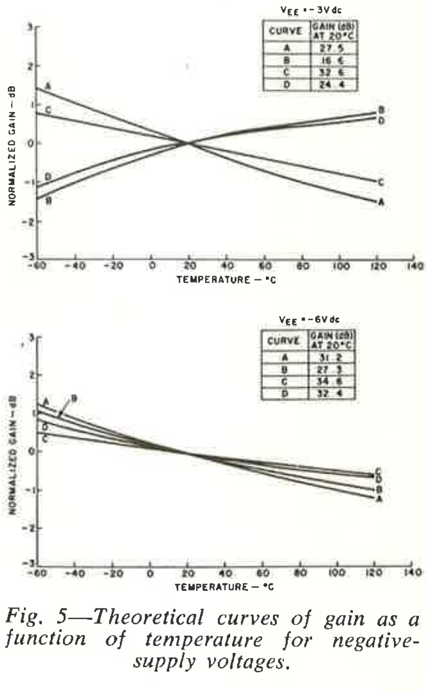


Fig. 5—Theoretical curves of gain as a function of temperature for negative-supply voltages.

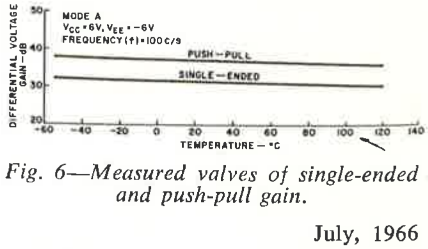


Fig. 6—Measured valves of single-ended and push-pull gain.

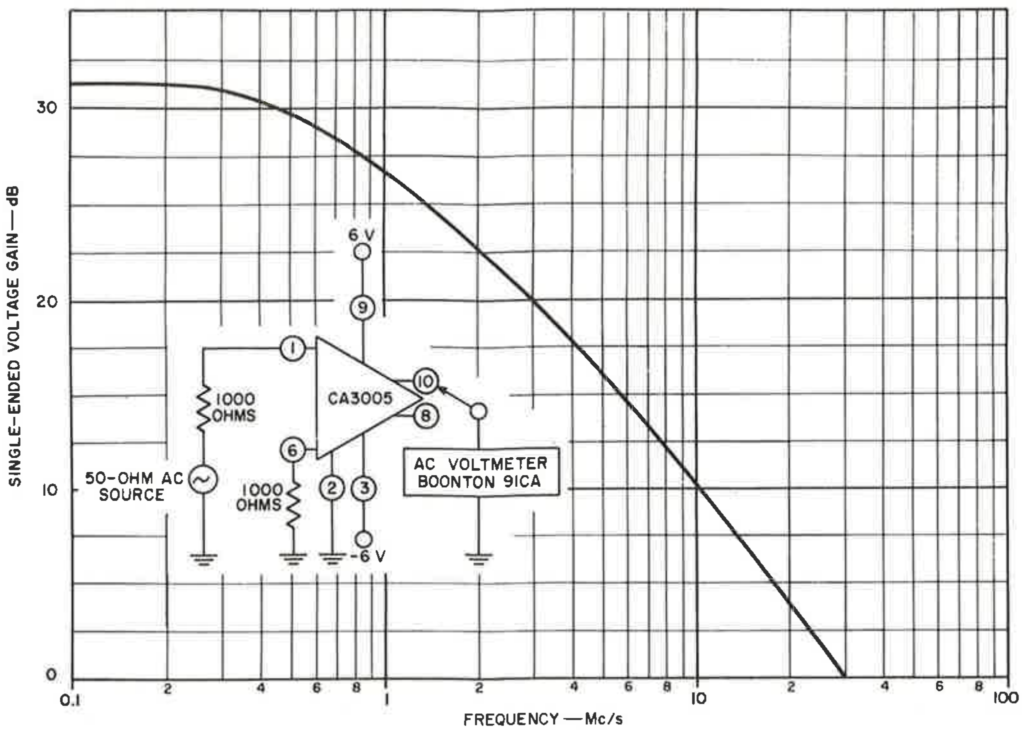


Fig. 7—Single-ended voltage gain of CA3000 as a function of frequency in test circuit shown.

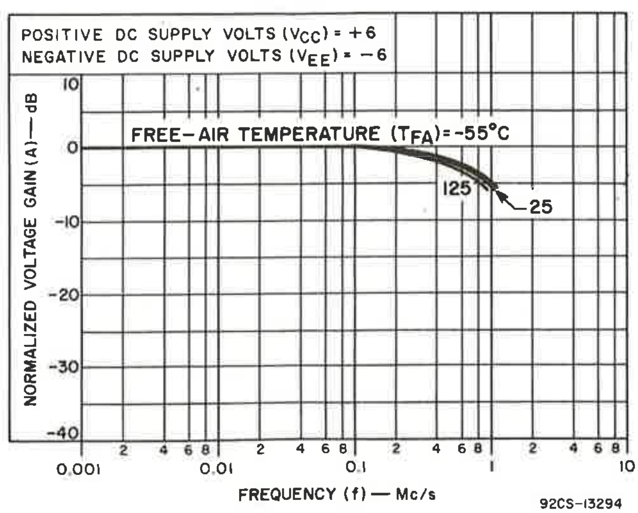


Fig 8—Normalized gain-frequency curves for CA3000 at three different temperatures.

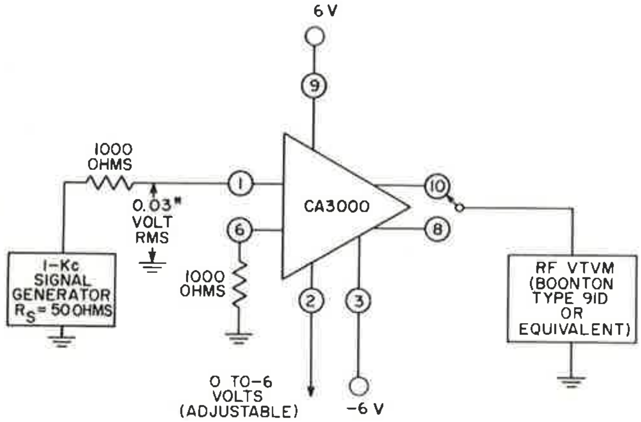
The output is taken from the collector that is not connected to the crystal (in this case, terminal 10). If a variable-feedback ratio network is used, as shown in Fig. 13(b), the feedback may be adjusted to provide a sinusoidal oscillation. Output waveforms for both circuits are also shown. The frequency in each circuit is 455 kilocycles per second, as determined by the crystal. The range of these crystal oscillators can be extended to frequencies of ten megacycles per second or more by use of collector tuning.

Modulated Oscillator—If a low-frequency signal is connected to terminal 2, as shown in Fig. 14, the CA3000 can function as an oscillator and produce an amplitude-modulating signal. The waveform in Fig. 14 shows the modulated signal output produced by the modulated oscillator circuit when a one-kilocycle-per-second signal is introduced at terminal 2 and a high-pass filter is used at the output.

Low-Frequency Mixer—In a configuration similar to that used in modulated-oscillator applications, the CA3000 amplified may be used as a mixer by connection of a carrier signal at the base input of either differential-pair transistor (terminal 1 or 6) and connection of a modulating signal to terminal 2 or 5.

Cascaded RC-Coupled Feedback Amplifier—The two-stage feedback cascade amplifier shown in Fig. 15 produces a typical open-loop midband gain of 63 dB. This circuit uses a 100-picofarad capacitor C_1 to shunt the differential outputs of the first stage. This capacitor staggers the high-frequency roll-offs of the amplifier and thus improves stability.

The gain-frequency characteristic of the feedback amplifier is shown in Fig. 16(a) for a feedback resistance R_f approaching infinity. The low-end roll-off of the amplifier is determined by the interstage coupling. Because agc may be applied to the first stage, the amplifier



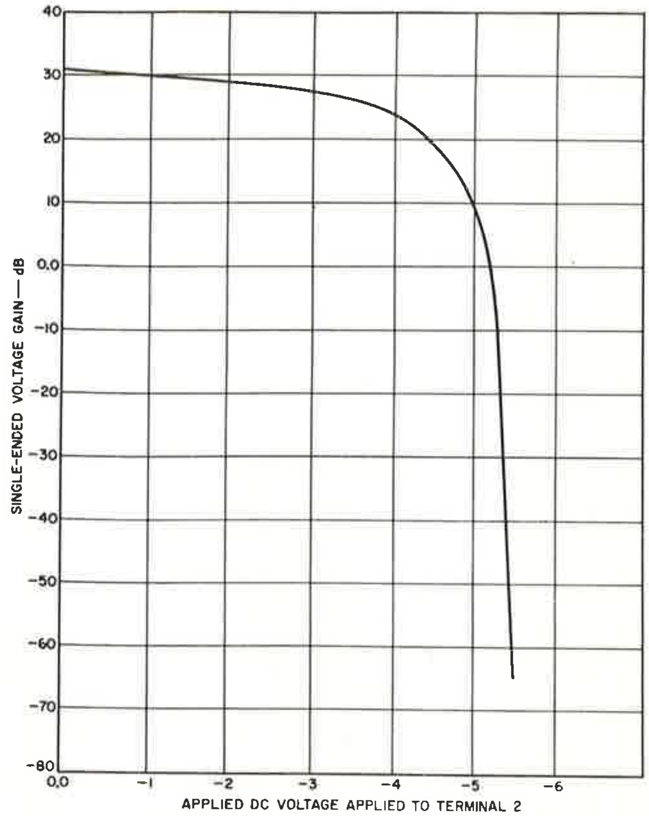
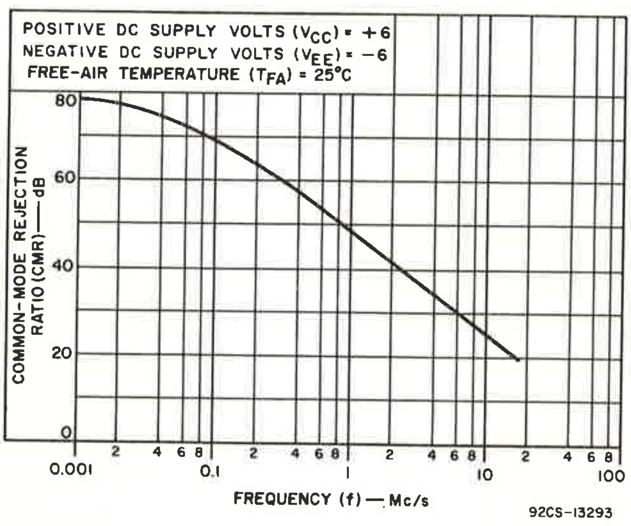
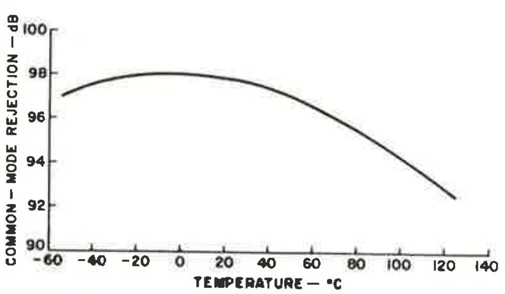


Fig. 9—AGC characteristics of CA3000 in test circuit shown at frequency of one kilocycle per second.





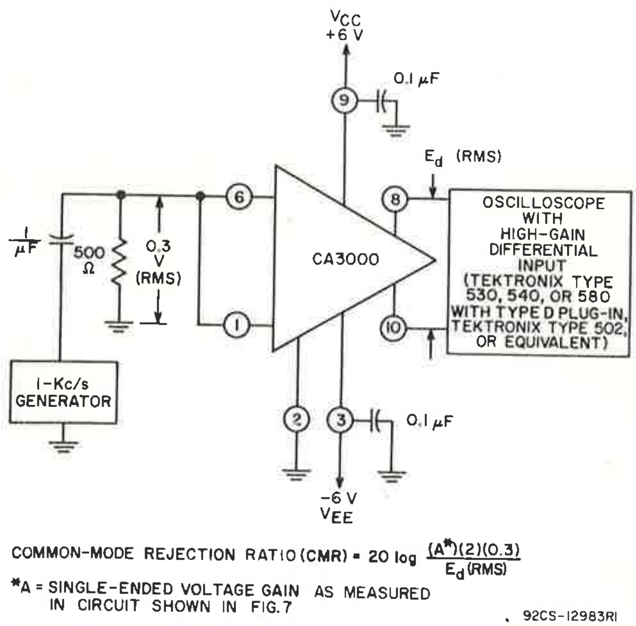


Fig. 10—Common-mode rejection of CA3000 as a function of frequency and of temperature in test circuit shown.

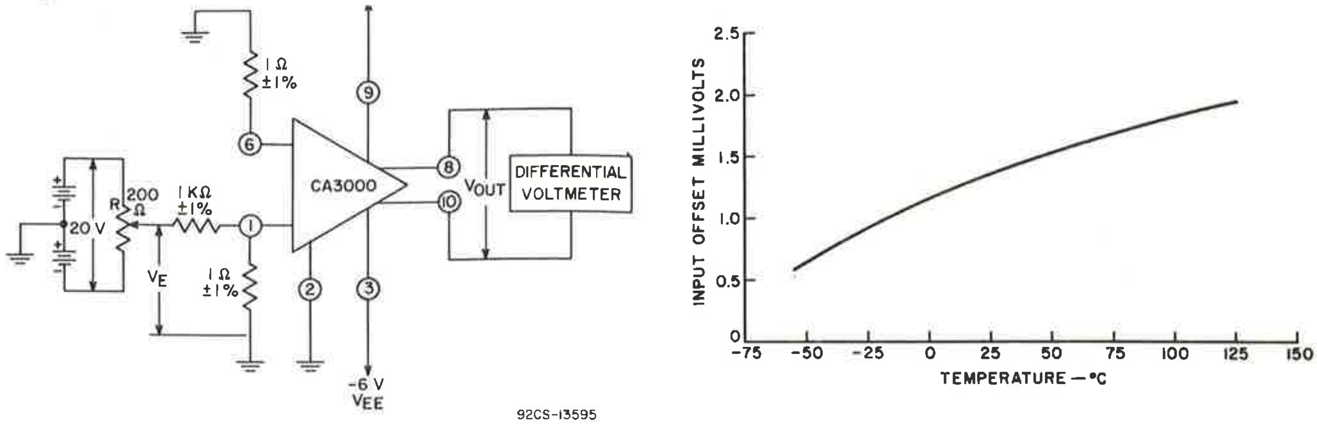


Fig. 11—Input offset voltage of CA3000 as a function of temperature in test circuit shown.

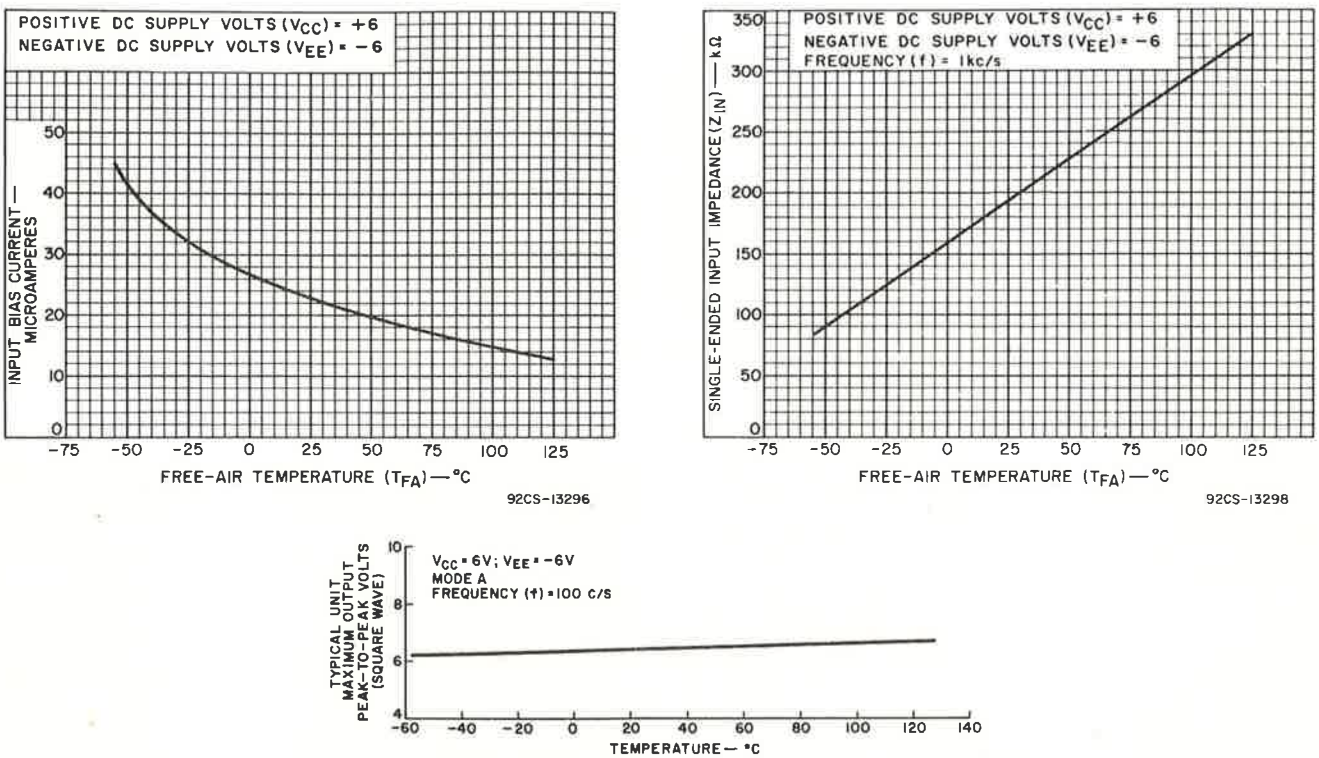


Fig. 12—Input bias current, input impedance, and dynamic output voltage of CA3000 as functions of temperature.

of Fig. 15 may be used in high-gain video-agc applications under open-loop conditions. If feedback is used to control the gain, agc may still be applied successfully.

Fig. 16(b) shows the agc characteristics for the two-stage amplifier under open-loop and two closed-loop conditions at a frequency of one kilocycle per second. As shown in Fig. 16(a), the open-loop bandpass is 18 cycles to 135 kilocycles per second; under closed-loop conditions, the bandpass is 1.3 cycles to 2 megacycles per second for 40 dB gain and 0.13 cycle to 6.6 megacycles per second for 20 dB gain. The negative feedback thus improves low-frequency performance sufficiently so that the use of the small coupling capacitors C_2 and C_3 involves little sacrifice in low-frequency response. If three or more CA3000 amplifiers are cascaded, the low-frequency roll-offs must be staggered as well as those at the high end to prevent oscillation. A three-stage cascade has a midband gain of approximately 94 dB.

Narrow-Band Tuned Amplifier—Because of its high input and output impedances, the CA3000 is suitable for use in parallel tuned-input and tuned-output applications. There is comparative freedom in selection of circuit Q because the differential amplifier exhibits inherently low feedback qualities provided the following conditions are met: (1) the collector of the driven transistor is returned to ac ground and the output is taken from the non-driven side, and (2) the input is adequately shielded from the output by a ground plane.

The CA3000 has an output capacitance of approximately 9 picofarads at a frequency of 10 megacycles per second. This capacitance will resonate a 28-microhenry coil at this frequency and give a minimum Q of 4.55 when the collector load resistor is the only significant load. With this low Q, stagger tuning may be unnecessary for many broadband applications.

Fig. 17 shows the CA3000 in a narrow-band, tuned-input, tuned-output configuration for operation at 10 megacycles per second with an input Q of 26 and an output Q of 25; the response curve of the amplifier is also shown. The 10-megacycle-per-second voltage gain is 29.6 dB, and the total effective circuit Q is 37. There is very little feedback skew in the response curve. The CA3000 can be used in tuned-amplifier applications at frequencies up to the 30-megacycle-per-second range.

Schmitt Trigger—The CA3000 can be operated as an accurate, predictable Schmitt trigger provided saturation of either side of the differential amplifier is prevented (hysteresis is less predictable if saturation occurs). Non-saturating operation is accomplished by operation in mode B (terminals 3 and 5 shorted together) in the configuration shown in Fig. 18. Large values are required for exter-

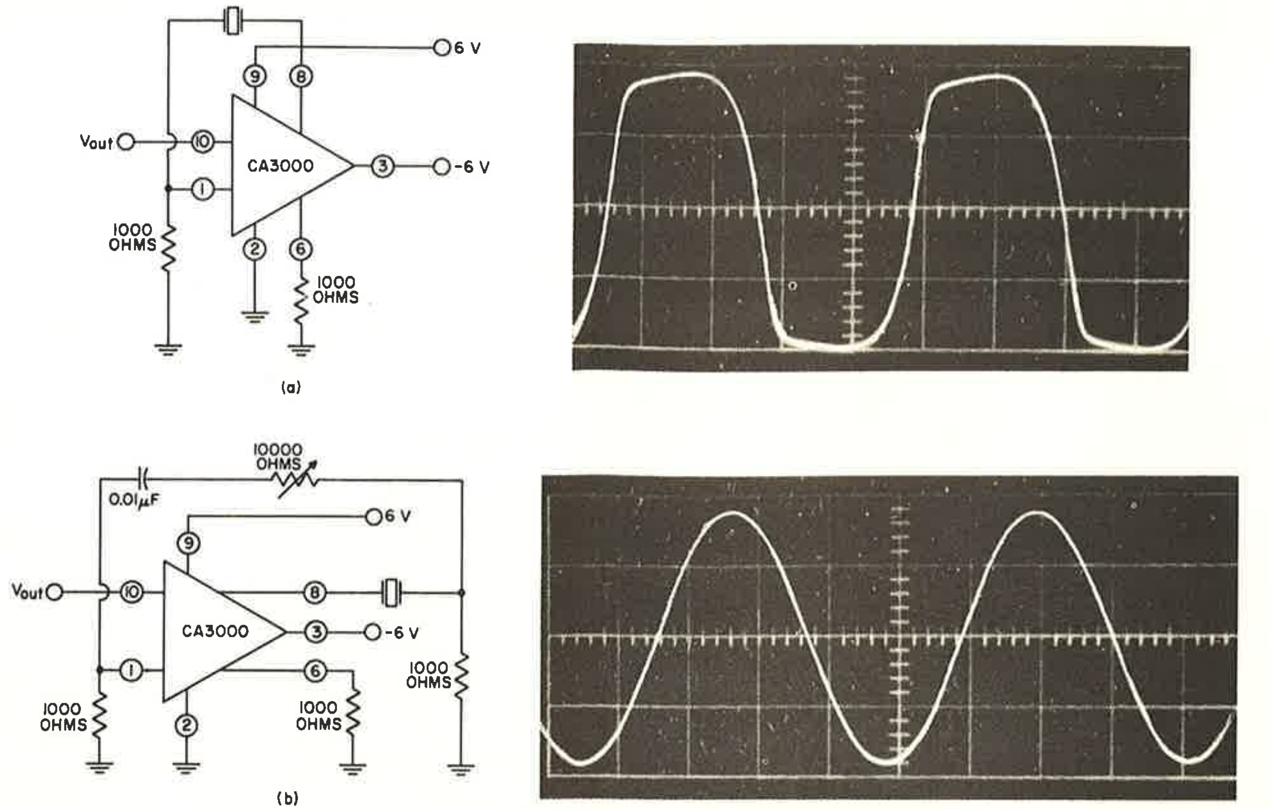


Fig. 13—Schematic diagrams and output waveforms of (a) crystal oscillator and (b) crystal oscillator with variable feedback.

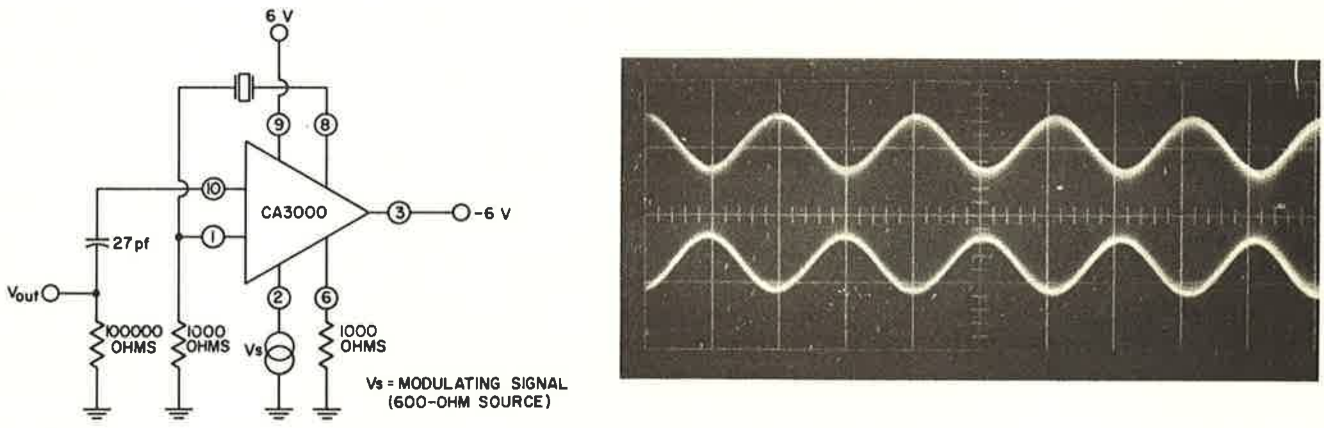
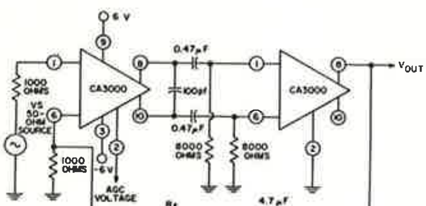


Fig. 14—Schematic diagram and output waveform of CA3000 modulated oscillator.

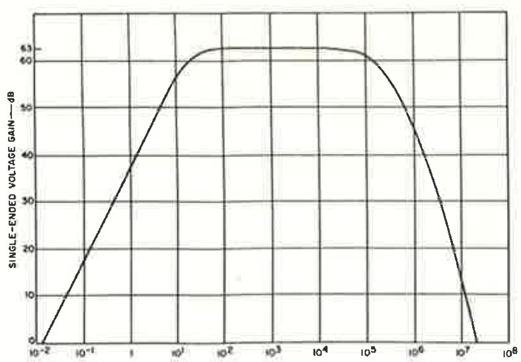


FOR $V_{AGC} = 0$

R_f	G_V
∞	63 db (OPEN LOOP)
100000 OHMS	40 db
9000 OHMS	20 db

Fig. 15—Cascaded RC-coupled feedback amplifier using two CA3000 circuits.

(a)



(b)

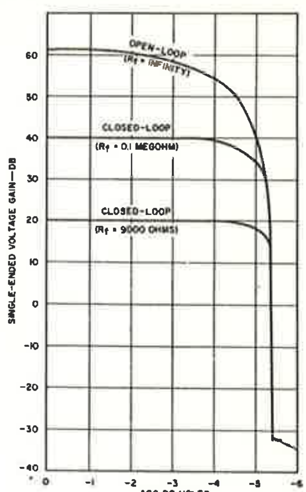


Fig. 16—(a) Gain-frequency and (b) agc characteristics of feedback amplifier shown in Fig. 15.

nal resistors R_1 and R_2 because they receive the total collector current from terminal 10. Because of the high impedances, resistor R_2 is actually a parallel combination of the input impedance (approximately 0.1 megohm) of the CA3000 and the 0.25-megohm external resistor. The Schmitt-trigger design equations (for $\alpha = 1$) are summarised below. In these equations, Q_2 and Q_4 are the differential-pair transistors, Q_1 and Q_5 are the emitter-follower transistors, and Q_3 is the constant-current sink.

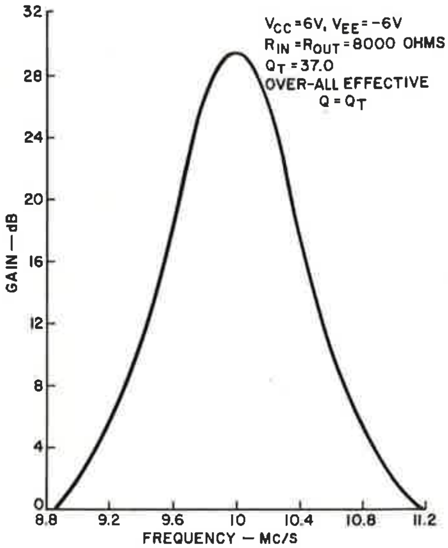


Fig. 17—Schematic diagram and response curve for 10-Mc/s tuned-input, tuned-output, narrow-band amplifier using CA3000.

STATE I: Q_2 off, Q_4 conducting (not saturated)

$$V^{9}_{I} = \frac{V_{CC}(R_2) - V_{EE}(R_1+8000)}{R_1 + R_2 + 8000}$$

where 8000 ohms is the output impedance of Q_4 (obtained from the published data). For R_1 = 27000 ohms and $V_{CC} = V_{EE}$ = 6 Vdc.

$$V^{9}_{I} = \frac{6V(R_2) - 6V(35000)}{R_2 + 35000} \quad (A)$$

$$R_2 = (R_1 + 8000)\frac{V_{EE} + V^{9}_{I}}{V_{CC} - V^{9}_{I}}$$

$$R_2 = (35000)\frac{6V + V^{9}_{I}}{6V - V^{9}_{I}} \quad (B)$$

$$V^{8}_{I} = V_{CC} - I_E\,(8000)$$

where I_E = collector current of transistor Q_3 $\approx$0.48 milliampere in operating mode B with V_{EE} = −6 volts dc.

$$V^{8}_{I} = 2.14\ V \quad (C)$$

$V_{FI} \equiv$ Firing voltage for transition from state I to state II

$$V_{FI} = V^{9}_{I} - 0.053 - 100\ I_E \text{ at } 25°C$$

$$V_{FI} = V^{9}_{I} - 0.101\ V \text{ at } 25°C \quad (D)$$

STATE II: Q_2 conducting (not saturated), Q_4 off

$$V^{8}_{II} = V_{CC}$$

$$V^{8}_{II} = 6\ V$$

$$V^{9}_{II} = \frac{(V_{CC} - I_E 8000)R_2 - V_{EE}(R_1 + 8000)}{R_1 + R_2 + 8000} \quad (A)$$

$$V^{9}_{II} = \frac{2.14\ V(R_2) - 6\ V(35000)}{R_2 + 35000} \quad (B)$$

$V_{FII} \equiv$ Firing voltage for transition from state II back to state I

$$V_{FII} = V^{9}_{II} + 0.053 + 100\ I_E \text{ at } 25°C$$

$$V_{FII} = V^{9}_{II} + 0.101\ V \text{ at } 25°C \quad (C)$$

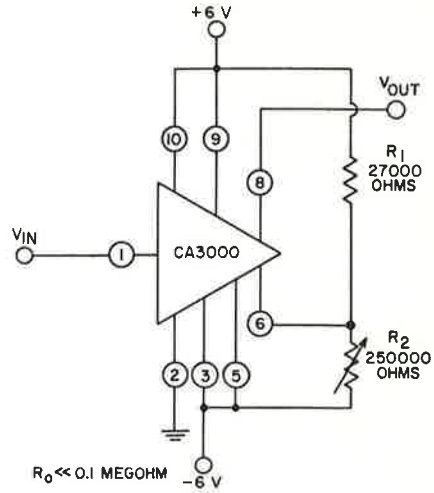


Fig. 18—Schematic diagram for Schmitt trigger using CA3000.

HYSTERESIS VOLTAGE

$$V_{HYS} = V_{FI} - V_{FII}$$

$$= \frac{3.86\ V\ (R_2)}{R_2 + 35000} - 0.202\ V \text{ at } 25°C$$

From the calculations for state I, it is evident that either V^{9}_{I} or R_2 must be a known design value. Because R_2 is a composite value, V^{9}_{I} is the more reasonable choice. The ability of these equations to predict the Schmitt-trigger performance is evidenced by the comparison of calculated and experimental data in Table III.

Table III—Comparison of Calculated and Experimental Data for Schmitt Trigger

Condition	*Parameter*	*Calculated*	*Experimental*
1) $V^{9}_{I} = -2V$	V_{FI}	−2.1V	−2.2V
	V_{FII}	−3.19V	−3.2V
	V_{HYS}	+1.09V	+1.0V
2) $V^{9}_{I} = -1V$	V_{FI}	−1.10V	−1.0V
	V_{FII}	−2.51V	−2.45V
	V_{HYS}	+1.41V	+1.4V
3) $V^{9}_{I} = 0$	V_{FI}	−0.101V	0
	V_{FII}	−1.83V	−1.8V
	V_{HYS}	+1.73V	+1.8V
4) $V^{9}_{I} = +1V$	V_{FI}	+0.9V	+1.0V
	V_{FII}	−1.15V	−1.0V
	V_{HYS}	+2.1V	+2.0V
5) $V^{9}_{I} = +2V$	V_{FI}	+1.9V	+2.0V
	V_{FII}	−0.472V	−0.5V
	V_{HYS}	+2.43V	+2.4V

(WITH ACKNOWLEDGEMENT TO RCA)

SUPER RADIOTRON 25TP4 PICTURE TUBE

The 25TP4 is a directly viewed glass picture tube having an aluminised screen $20\frac{7}{8}''$ x $16\frac{3}{8}''$ with a minimum projected area of 327 square inches. It employs 110° magnetic deflection and low voltage electrostatic focus. Integral implosion protection is provided by a formed rim band and tension band around the periphery of the tube panel. Mounting lugs have also been included in the tube design, thus eliminating the need for complicated mounting and implosion protection equipment in the receiver.

GENERAL

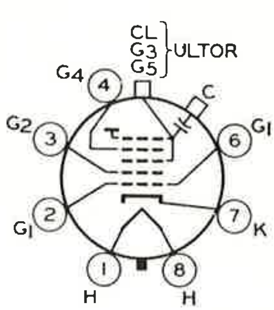
Heater Voltage	6.3 volts
Heater Current	0.6 amps.
Direct Interelectrode Capacitances:	
Cathode to all other electrodes	5 pf
Grid 1 to all other electrodes	6 pf
External conductive coating to anode:	
Maximum	2500 pf
Minimum	2000 pf
Faceplate	Filterglass
Light Transmission	40%
Phosphor	Aluminised P4 Sulphide
Fluorescence	White
Phosphorescence	White
Focusing Method	Electrostatic
Deflection Method	Magnetic
Deflection Angles (approx.):	
Diagonal	110°
Horizontal	100°
Vertical	84°

Tube Dimensions:			
Overall Length	15.813	±0.281	inches
Greatest Width	22.500	±0.125	inches
Greatest Height	18.250	±0.125	inches
Diagonal	25.781	±0.125	inches
Neck Length	5.125	±0.125	inches

Screen Dimensions (min.):	
Horizontal	20.875 inches
Vertical	16.375 inches
Diagonal	24.250 inches
Area	327 sq. in.
Electron Gun	Unipotential
Bulb	J203½-A1
Bulb Contact	JEDEC J1-21
Base	JEDEC B7-208

SOCKET CONNECTIONS

Pin 1—Heater
Pin 2—Grid No. 1
Pin 3—Grid No. 2
Pin 4—Grid No. 4
Pin 5—Blank
Pin 6—Grid No. 1
Pin 7—Cathode
Pin 8—Heater
Bulb Contact—Anode



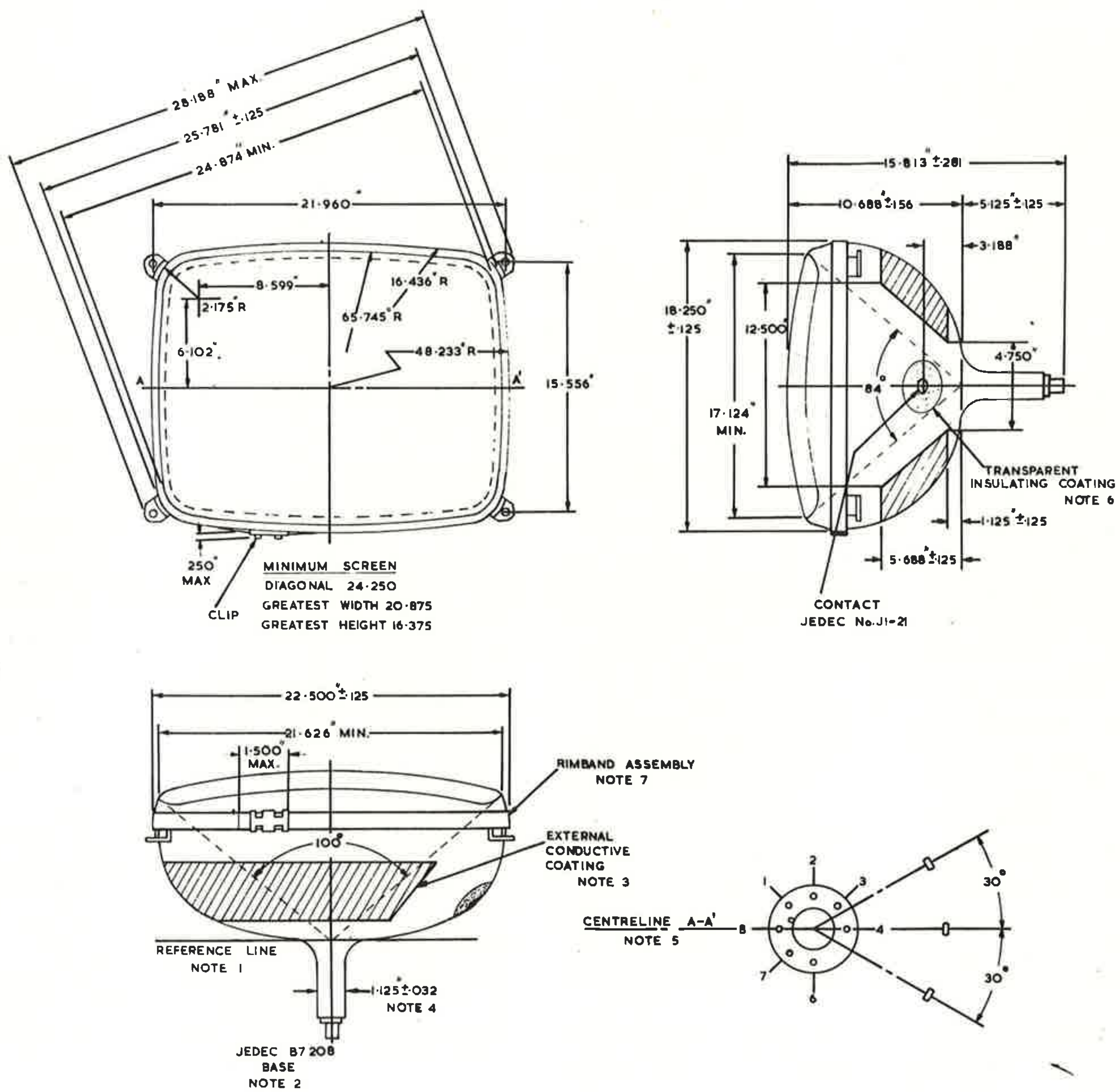
RATINGS, DESIGN MAXIMUM SYSTEM

(Unless otherwise specified, voltage values are positive, and measured with respect to cathode)

Maximum Anode Voltage .... 22,000 volts
Minimum Anode Voltage .... 12,000 volts

Maximum Grid No. 4 Voltage .... +1100, —550 volts
Maximum Grid No. 2 Voltage .... 550 volts
Minimum Grid No. 2 Voltage .... 200 volts

Grid No. 1 Voltage:
- Maximum Negative Value .... —154 volts
- Maximum Negative Peak Value .... —220 volts
- Maximum Positive Value .... 0 volts
- Maximum Positive Peak Value .... 2 volts



Maximum Heater-Cathode Voltage, Heater Negative with respect to Cathode:	
During Warm-up, 15 secs.	450 volts
After Warm-up Period	200 volts
Maximum Heater-Cathode Voltage, Heater Positive with respect to cathode	200 volts

TYPICAL OPERATION, GRID DRIVE SERVICE

(Unless otherwise specified, all voltage values are positive with respect to cathode)

Anode Voltage	18,000 volts dc
Grid No. 4 Voltage*	0-400 volts dc
Grid No. 2 Voltage	400 volts dc
Grid No. 1 Voltage	—36 to—94 volts dc

TYPICAL OPERATION, CATHODE DRIVE SERVICE

(Unless otherwise specified, all voltage values are positive with respect to Grid No. 1)

Anode Voltage	18,000 volts dc
Grid No. 4 Voltage*	0-400 volts dc
Grid No. 2 Voltage	400 volts dc
Cathode Voltage	36 to 78 volts dc

MAXIMUM CIRCUIT VALUE

Grid No. 1 Circuit Resistance	1.5 megohms

* The grid No. 4 (or grid No. 4 to grid No. 1) voltage required for optimum focus of any individual tube will be a value between 0 and 400 volts independent of anode current. It will remain essentially constant for values of anode (or anode to grid No. 1) voltage and grid No. 2 (or grid No. 2 to grid No. 1) voltage within the ranges shown for these items.

NOTES

NOTE 1. Yoke Reference Line is determined by plane surface of flared end of JEDEC Reference Line Gauge No. 126 when seated on funnel of tube. With minimum neck length tube, the PM centring magnet should extend no more than 2¼" from Yoke Reference Line.

NOTE 2. Lateral strains. on the base pins must be avoided. The socket should have flexible leads permitting movement. The perimeter of the base wafer will be inside a 1¾" diameter circle concentric with the tube axis.

NOTE 3. External conductive coating forms supplementary filter capacitor and must be grounded.

NOTE 4. Neck diameter may be a maximum of 1.168" at the splice.

NOTE 5. Base pin No. 4 aligns with centreline A-A' within 30° and is on the same side as anode contact J1-21.

NOTE 6. To clean this area, wipe only with a soft, dry lintless cloth.

NOTE 7. The Rimband assembly must be grounded.

BOOK REVIEWS

"ELECTRICAL INSTRUMENTS IN HAZARDOUS LOCATIONS," E. C. Magison, Plenum Press, 225 pages. Size 9" x 6".

This title is a publication of the Instrument Society of America, and the scope of the book is indicated by the title. Much of the book is based on work done by the ISA's Recommended Practice Committee, and it presents information needed for an understanding of the requirements for the safe and economical use of electrical instrumentation in dangerous areas. The book is an excellent summary and collection of pertinent material on the nature and avoidance of electrical hazards. As the title of the book suggests, this is to some extent a specialised book. It is always of interest to become aware of some of the problems that others in different branches of an industry have to contend with, and how one sets about overcoming them. Although a specialised book, this title is at the same time a fascinating book to take a casual browse through, and should certainly find a place in any library.

"FEEDBACK CIRCUIT ANALYSIS," S. S. Hakim, Iliffe Books Ltd., 392 pages, 316 diagrams. Size $8\frac{3}{4}$" x $5\frac{1}{2}$".

Dr. Hakim has made quite a name for himself over the last few years as a technical author of some stature. His "Junction Transistor Circuit Analysis" and "Transistor Circuits in Electronics" have already been reviewed in these pages, and this latest title is a further fine contribution. In this comprehensive evaluation of the various effects of feedback on the circuit performance of an amplifier, the classical theory and its limitations are examined in detail. From this examination a more generalised theory, accounting for the effects associated with transistorised linear electronic circuits is developed. In addition, the book investigates the problem of stability and its relation to the closed-loop transient response, open-loop frequency response and the driving point impedance of the feedback circuit. This title is expected to be of great service to final-year electronics graduates, post-graduates and practising electronic engineers seeking a sound understanding of the feedback problem and its various circuit implications.

"SOVIET RADIO ENGINEERING." Translation of "Izvestiya VUZ. Radiotekhnika". Periodical. The Faraday Press Inc.

This is one of a growing number of publications that are appearing to fill the needs of scientists and engineers in keeping in touch with progress in the Soviet Union. This particular publication consists of a cover-to-cover translation of the original Russian publication, which is a well-known magazine of some authority. Whilst a regular and complete translation of this type must be of interest to those English-speaking people who are interested in the art, it is likely that most of their reference thereto will be through a library. The high cost of translation and other factors result in an annual subscription of 115 US dollars, which puts it well beyond the pockets of most of us. As a step towards the further exchange of information, however, the production of this publication is a very commendable step.

NEWS & New Releases

Ceramic Tetrode with Integral Boiler*

The first commercial ceramic tetrode with an integral vapour cooling system has been introduced into the extensive range of power tubes available.

This new tetrode, the CY1170J, is intended for use in audio amplifiers, linear amplifiers, class 'C' amplifiers and oscillators. It will operate at full ratings up to 30 Mc/s and will deliver an output power of 82.5kW under class 'C' unmodulated conditions.

The advanced concept of combining ceramic construction with the recently developed 'built-in' method of vapour cooling results in major advantages to both designer and equipment user.

*Illustrated on cover of this issue.

Ceramic provides for :

* More robust construction
* Higher reliability
* Increased operating frequency
* Higher temperature processing

Integral Cooling allows:

* Higher maximum anode dissipation
* Space saving
* Lower initial costs
* Simple external cooling services
* Greater reliability

ABRIDGED DATA

Anode Dissipation (Class C Telegraphy)	60kW (max.)
Anode Voltage	15kV (max.)
Frequency (for full rating)	30Mc/s (max.)
Output Power	82.5kW

AN ELECTRONIC PRECISION TIMER

T. D. LAMB

This article is based on an entry to the Victorian Science Talent Search of 1965, and has been specially prepared for this magazine. The object of the project was to investigate the theoretical and practical aspects of building an electronic device to measure small time intervals, up to one or two seconds, with an accuracy of one ten-thousandth of a second.

Two ways of doing this immediately come to mind. One method would be to measure the decay of voltage across a charged capacitor in relation to time. The voltage across the capacitor at any point in time will depend on the value of the capacitor, the potential to which it has been charged, the total resistance across the terminals (discharge resistance in parallel with the leakage resistance), and the lapse of time since the commencement of the discharge. If the values cited are known, then the voltage across the capictor may be translated in terms of time.

The discharge of the capacitor is exponential, and is governed by the expression:

$$V = E\ (1 - \epsilon - t/RC$$

where ϵ is the base of natural logarithms ($\simeq$ 2.718).

t is the time in seconds after commencement of the discharge.
E is the initial voltage to which the capacitor is charged.
R is the discharge resistance in ohms.
C is the capacitance in farads, and
V is the potential difference in volts across the capacitor at time = t.

Whilst the method described is simple, it is not possible to use it to give the required degree of accuracy. Not only would it be necessary to determine the values of the capacitor and discharge resistor to five significant figures, but the initial voltage and the voltage after the time t would also need to be measured with a similar degree of accuracy. Further difficulty would arise from the fact that arrangements to measure the voltage across the capacitor precising at the required point in time would not be simple, and that the measuring device itself will have a finite impedance. Further, under a given set of conditions, the accuracy of measurement would vary according to the position on the discharge curve at which the measurement was made. These considerations mean that this method is unsuitable for this project.

A second method involves counting the number of pulses present in the output of a good quality oscillator during the time to be measured. In order to make a measurement to one ten-thousandth of a second, the oscillator frequency will need to be at least 10KHz, and preferably higher, because the counter can count only complete oscillations. If n pulses are counted from an oscillator operating at a frequency of Hz over a period equal to t seconds, then t = n/f. Because division by f is required, the method will be simplified by making f an integral power of 10, e.g., 10 KHz or 100 KHz.

This method is much more attractive. Only two quantities need be known with good accuracy, the oscillator frequency and the number of oscillations during the measurement period. The oscillator is comparatively easy to arrange, but we are left with the counter problem.

Leaving the counter for a moment, there is a further possibility which is actually used to a large extent in industry, and which is partly derived from the two methods already mentioned. This method is to apply a pulse train for the duration of the measuring period to a capacitor. The pulses would be unidirectional and would be applied to the capacitor, say, through a diode. Each pulse provides some charge for the capacitor, so that during the application of the pulse train, the charge on the capacitor will increase in step-like fashion. The final charge on the capacitor with all factors known will be a measure of the time during which the pulse train was applied.

Reflection will show that this method has some of the disadvantages of the first-described system. Further, it is suitable only when the time to be measured is short in relation to the time constant of the charging circuit, because as the capacitor approaches the fully-charged condition, it becomes increasingly difficult to detect and measure changes in voltage thereon.

Finally, therefore, it was decided to pursue the second method described above, and to construct an interval timer with the stated degree of accuracy. The timer was also required to be inexpensive, simple to operate, lightweight and of robust construction. It was also decided to use an oscillator frequency of 100 KHz. Consideration then turned upon the type of counter to be used.

COUNTING DEVICES

The first type of counter investigated was of the electro-mechanical type, in which an armature drives a drum-type counter, the count increasing by one for each energisation of the armature. These counters are familiar and are widely used in telephone work. It is clear that they could not be expected to operate at a high speed, but as it was thought that they could be useful in the low-speed sections of a timer, two were obtained from surplus stores.

It was first necessary to determine the probable maximum counting rate of these counters. This was done by setting them up in a circuit such that the mechanical reaction when the counter armature was energised was used to break a circuit supplying current to the operating coil. In this way the counter was caused to form part of a vibrator system, and several checks were made of the number of pulses over measured intervals of ten seconds. This showed that the maximum counting rate the units could accommodate was of the order of 16 pulses per second. A small disadvantage of this type of counter is that it cannot easily be reset to zero.

For example, we may be interested in measuring the duration of light pulses, as in the case of some types of camera shutter checking and adjusting apparatus. Here the likely form of the switch would be a transistor actuated by a light-sensitive cell. In a similar way a microphone could be used as part of a sound-sensitive switch. One application the author had in mind was the measurement of the speed of a bullet. In this case one possibility is to arrange two spaced wires in the path the bullet is to follow and arrange that the bullet break the wires in turn. Light-gauge foil could be used instead of wire.

The ideal type of counter in this application is naturally the decade counter. This type of counter counts from one to nine input pulses, the state of the count at any time being shown by a variety of methods. On receipt of a tenth impulse, the counter resets to zero and supplies a pulse to a further counter which counts in tens, and so on. Any desired number of decade counters can thus be set up in a counting chain to count as many places either side of a decimal point as one may require. A suitable circuit for such a decade counter was found in "Miniwatt Germanium and Silicon Transistors and Diodes," page 33. This counter has a maximum counting speed of 20,000 counts per second, the counting state is indicated by a calibrated milliammeter, and the counter can be manually reset to zero.

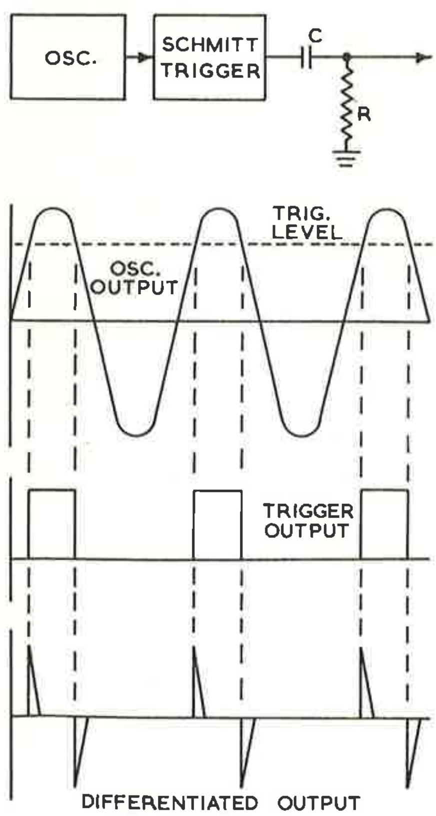


Fig. 1—Basic counting pulse-train generation.

PULSE GENERATOR

The next point to consider was the pulse generator. This would consist basically of an oscillation generator and pulse shaping means to provide a pulse train of a type suitable for actuating the counter. It had already been decided to use an oscillator frequency of 100 KHz. Crystals for use in crystal-controlled oscillators are readily available at this frequency, and have an accuracy of $\pm 0.005\%$, that is, 5 parts in 100,000. The customary small trimmer could be used in conjunction with the crystal to adjust the precise oscillator frequency, and to calibrate an harmonic of the output against a known standard, e.g., against station WWVH, which transmits at a frequency of 10 MHz with an accuracy of one part in 10^7. It was considered that with the aid of a communication receiver, it would be possible to beat the 100th harmonic of the crystal oscillator against WWVH and to adjust the oscillator for a beat frequency of 50 Hz or less. This would mean that the oscillator was then adjusted to the required 100KHz within about one half cycle, or an accuracy of one part in 200,000.

The output of the oscillator is a sine wave, and is therefore unsuitable for application directly to a counter. The output of the oscillator must therefore be shaped into more suitable form. There is a variety of circuits which are capable of filling this shaping function. One of these is the well-known Schmitt Trigger. This circuit is basically a level-sensing circuit, which turns "on" rapidly when the input exceeds a predetermined value, and turns "off" again when the level falls below that value. From a sinusoidal input, e.g., from the crystal-controlled oscillator, the Schmitt Trigger will therefore produce a train of unidirectional pulses with a frequency equal to that of the applied sine wave. The width of the pulses will depend on the level at which the circuit triggers.

All that requires to be done now is to produce a series of short pulses at the same repetition rate, and this can be done by differentiating the rectangular output of the trigger circuit. A simple RC circuit of short time constant will provide the differentiation, and will produce, from the rectangular output of the trigger circuit, a series of short positive and negative-going pulses corresponding to the "on" and "off" times of the trigger circuit. This is shown in Fig. 1.

The differentiated output consists of both positive and negative pulses, whilst only one polarity is required to actuate the counting circuits. In many cases the counting circuits will not respond to the incorrect and unwanted polarity pulses. If desired, pulses of the unwanted polarity can easily be removed by using a diode to clip them, e.g., connected across the resistor in the RC differentiating circuit.

TIMING SWITCH

Having disposed of the question of counters and the pulse train for application thereto, the remaining basic matter to be discussed is the switch through which the timing pulses will be applied to the timer during the timing period. The requirements for this switch indicated that it be of the electronic type, but the exact form of the switch could be one of several, dependent on the application of the timer itself.

For a higher speed counter, the obvious choice was an electronic counter. Either valves or transistors can be used in counting circuits to operate at a variety of speeds according to the characteristics of the devices used. Because of the requirement to make the unit light and portable, the choice of a transistorised counter would seem to be indicated. However, at the time the project was carried out, the author was in secondary school and the cost of the project had to be kept down. Some inexpensive transistors of the audio type were available, but none of the faster switching types that would be needed for a high speed counter. For the first counter, therefore, an electronic valve counter was used, followed by transistor counters.

THE BASIC PLAN

The basic plan of the timer has now evolved. It will consist of a 100 KHz oscillator, with Schmitt Trigger and differentiating means to provide a train of pulses. This train of pulses will be applied through a timing switch to a chain of decade counters. The first stage of the counter chain will use electron valves because the transistorised counter will not count at the speed required. It is true that the transistorised counter could be modified to work at a higher speed, but this would need a different type of transistor which is not available to the author at this time. The output of the valve counter will be at 10 KHz, so that the next three counter stages will be transistorised. The output from the last transistor counter will be at 10 Hz, and will be fed to one of the mechanical counters.

In this way the mechanical counter will be used to indicate seconds and tenths of seconds, the milliammeters connected to the transistorised counters will indicate hundredths, thousandths and ten-thousandths of seconds, and the neon indicators on the electron valve counter stage will indicate one hundred thousandths of seconds. A block diagram of the time unit is shown in Fig. 2.

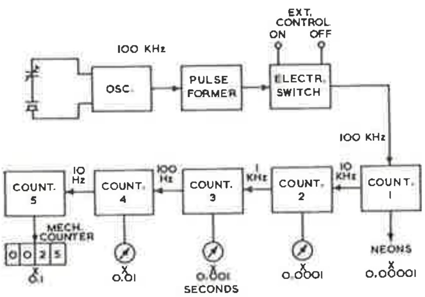


Fig. 2—Block diagram of the interval timer. Counter 1 is an electron valve counter, counters 2 to 4 are transistorised counters, and counter 5 is an electro-mechanical counter.

Having discussed some of the basic problems involved, it now remains to be seen how the unit was finally constructed. Each stage of the project will be dealt with in sequence, followed by a few remarks on the unit in general.

OSCILLATOR

A 100 KHz crystal vacuum-mounted in a B7B envelope was purchased for the project. Because an harmonic content up to the 100th harmonic was required to facilitate adjustment and calibration against WWVH, an orthodox harmonic crystal oscillator circuit was used, based on a circuit given in "Amateur Radio Handbook," page 470. It is true that other methods of adjusting and calibrating the crystal oscillator could have been used, but this seemed to be the simplest and cheapest with the facilities available.

The circuit of the oscillator is shown in Fig. 3. The screen grid of the valve functions as the anode of a Pierce-Colpitts oscillator, with electron coupling to the plate of the valve. The oscillator was built on part of a chassis carrying the counting arrangements. It was provided with a coaxial socket for taking out a calibration signal, and was found to have a very rich harmonic content. In fact, had a communication receiver not been available, it would have been possible to calibrate the 15th harmonic of the oscillator against station 3AK, 1.5 MHz. It was found possible in this way to adjust the oscillator for a beat against 3AK of about 1Hz, which meant that the fundamental of the oscillator was adjusted to within 1/15th of 1 Hz, that is, an accuracy of one part in one and a half million.

The trimmer in the oscillator was actually connected in series with the crystal. The capacitor is at a potential of about 150 volts above ground. For this reason, and to reduce capacitance effects, the trimmer should be adjusted with a tool made of insulating material.

SCHMITT TRIGGER AND VALVE COUNTER

There were a number of reasons why it was not possible to complete the project to the extent of making and incorporating the electron valve counter mentioned earlier. However, several suitable circuits are available in the literature and no difficulty would be expected in finding a circuit and making it work. In general these circuits use neon indicators of the count, although it is also possible to use some of the special counter indicators in which numbers are successively energised in a special neon tube.

Because time and other considerations precluded the incorporation of the valve counter, the Schmitt Trigger also was not constructed, although it was planned along the lines shown in Fig. 4. It will be seen that this is a fairly conventional circuit and presents no special problems. The value of the coupling capacitor bringing in the output of the 100 KHz crystal-controlled oscillator could be determined experimentally for satisfactory operation.

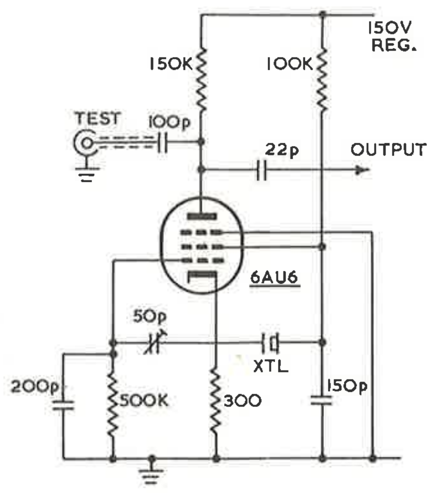


Fig. 3—Circuit diagram of the 100 KHz crystal oscillator.

The operation of the three transistorised counter stages and the final mechanical counter hinges of course on the output from the Schmitt Trigger and the valve counter in the original arrangement proposed. In the absence of these two units it was decided to get the remainder of the counter operative by using a multivibrator to count down from the 100 KHz output of the crystal-controlled oscillator to 10 KHz, the required input frequency for the first of the transistorised counters. This would mean that the timer would be operative to the fourth decimal place instead of the fifth place. However, the remainder of the proposed circuitry could be inserted at a later date if required, and if circumstances permitted.

The multivibrator used for the count-down function is shown in Fig. 5. Here again the circuit is conventional. Like the Schmitt Trigger it was arranged to operate from the regulated 150-volt supply to provide as much stability as possible. The potentiometer provided in the circuit allows the running frequency of the multivibrator to be adjusted within a range of frequencies. The required operational frequency of 10 KHz lies within this range of frequencies, and when the multivibrator is connected to a source such as the 100 KHz oscillator, adjustment of the potentiometer allows the multivibrator to be locked in by every tenth pulse from the oscillator and so run at 10 KHz with an accuracy approaching that of the oscillator. Correct locking of the multivibrator can be checked, say by using an audio oscillator, or better still by using an oscilloscope.

In the initial stages of the tests of this multivibrator, the input from the 100 KHz oscillator was applied to the left-hand (in the diagram) grid of the valve. This was found to provide too large an input, which gave uncertain operation of the multivibrator. The input was then changed to the plate of the valve, a position requiring a higher input, and this resulted in satisfactory operation.

Referring back to the block diagram of Fig. 2, the absence of the Schmitt Trigger pulse shaping circuit and the valve counter means that the output of the 100 KHz oscillator was fed directly to the count down multivibrator, and the electronic switch was then placed between this multivibrator and the first of the transistorised counter circuits. This will clarify the difference between the plan as originally proposed and what actually happened.

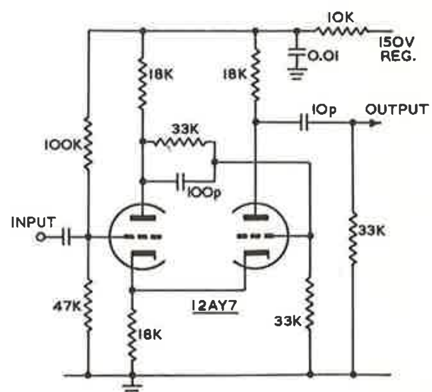


Fig. 4—Proposed Schmitt Trigger circuit for pulse-shaping the 100 KHz oscillator output.

TRANSISTORISED DECADE COUNTERS

As previously mentioned, these units are based on a published circuit, and it is therefore not proposed to present the circuit again here. Three such counters are used in this timing unit, all three being identical. A block diagram of one of the decade counters is included here in Fig. 6. Each counter consists of four binary flip-flops and a feedback network as shown. The flip-flops are arranged so that they turn "on" without providing a pulse of the required polarity to effect the following flip-flops in the arrangement, but when a flip-flop turns "off," a pulse is provided to turn "on" the next unit in the chain. The counter therefore divides the input pulses into two, hence the name.

The use of four flip-flops of this type produces a device which is actually capable of counting to 2^4 or 16. However, only a count to ten is required. The counting redundancy is overcome by a suitable arrangement of the feedback, as will now be seen. Let us assume that all the four flip-flops in the counter are in the "off" condition. The application of a pulse of suitable polarity at the input of the counter will turn circuit A "on." The next suitable pulse will turn a "off" and B "on." Successive pulses will actuate the counter along the programme shown in Table I until after the seventh input pulse, circuits A, B and C will be "on." The eighth pulse will turn "off" A, B and C and turn "on" D.

But at this stage the feedback arrangements mentioned come into play, and feedback from stage D turns "on" circuits B and C again immediately, so that after the eighth pulse, B, C, and D are "on." Then the ninth pulse turns "on" A also, whilst the tenth pulse then tu;ns "off" all four stages, so resetting the counter to zero, the original condition. It will be seen therefore that the provision of the feedback turns the counter from a sixteen-fold counter to a decade counter.

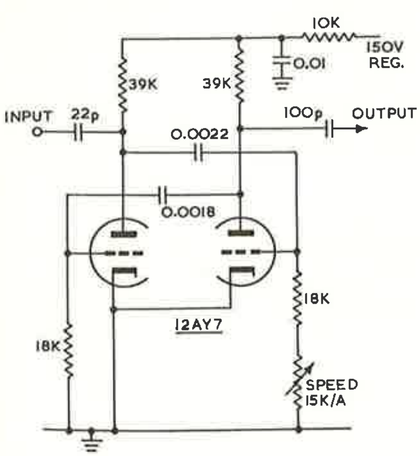


Fig. 5—Multivibrator used to count down from 100 KHz to 10 KHz.

When stage D is turned "off," it produces another output pulse which may be applied to a further decade counter of the same type. This counter would then advance by one digit, indicating a count of 1 in the next decade. Any number of decade counters can be connected in this way. An arrangement of x counters will allow a count to be made to $10^x - 1$.

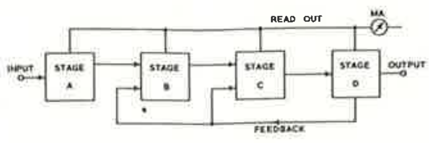


Fig. 6—Block diagram of one of the transistorised counters.

In the case of the counter being discussed, the indication of count is given by means of a milliammeter. Each of the counting steps of the counter corresponds to a unit of current flowing through the meter, so that suitable calibration of the meter will enable a direct reading to be made of 0 to 9. Resetting of the counter manually may be effected by the use of a switch which causes all stages to return to "off."

Because this project was required to proceed at low cost and meters are expensive items, it was decided to use a single meter that was at hand. This was done by permanently wiring low values of resistors in the read-out circuits of the three transistor decade counters. The single meter was then arranged with a four-pole switch so that it could be successively connected across each of the resistors mentioned to measure output current from the counter. A timing reading would then be taken by moving the switch in turn to each of the three active positions and noting the current reading at that point. The arrangement by which the meter was switched across fixed resistors avoided the possibility of generating surges that might have disturbed the count. Because the fixed resistors function as shunts for the meter, it is important that they all have the same value within the closest possible tolerance, so that calibration of the meter will hold good at each counter.

The calibration of the meter can be carried out experimentally merely by providing the required number of pulses to a counter and observing the meter movement. Series or shunt resistance can then be provided for the meter as required so that the "nine" count corresponds to full deflection or thereabouts and the "zero" count corresponds to minimum reading.

The output pulse from stage D of each counter which is available for switching the next decade counter is positive going. However, the amplifier provided for the mechanical counter and described elsewhere requires a negative pulse to actuate it. These negative pulses were obtained from the alternative collector in the stage D flip-flop through a 0.01 mfd capacitor.

TABLE I

Decade Counter Programme

Pulse No.	Condition			
	A	B	C	D
1	ON	—	—	—
2	—	ON	—	—
3	ON	ON	—	—
4	—	—	ON	—
5	ON	—	ON	—
6	—	ON	ON	—
7	ON	ON	ON	—
8	—	ON	ON	ON
9	ON	ON	ON	ON
10	—	—	—	—

MECHANICAL COUNTER CIRCUIT

The last counter in the counting chain is an electro-mechanical unit, which requires a substantial current to operate it. The operating coil has a DC resistance of 2,500 ohms, and is probably designed for an operating current of the order of 20 ma. It was clear that the output of the next preceding counter, a transistorised unit, would not operate this electromechanical counter directly, and that an amplifier would be required. It was decided to use a valve amplifier, based on a 6GW8 triode-pentode valve. The idea here was to use the pentode section to operate the counter and the triode section to amplify the input pulse from the preceding counter.

It was thought that to obtain satisfactory operation of the counter, it would be necessary to operate the pentode section cut off, and then drive it into conduction to energise the counter coil. This seemed to be the surest way to do it, although other methods may have worked equally well. Early trials of such a circuit did not work very well, and it was thought that this was due to the fact that the pentode was self-biased and therefore not well cut off, and also due to the fact that the input pulse was of short duration.

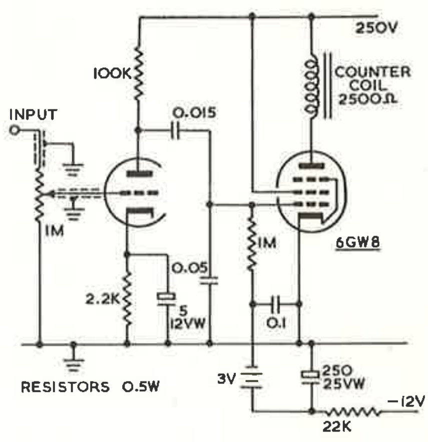


Fig. 7—Circuit diagram of the valve amplifier used to drive the final electro-mechanical counter.

The circuit finally evolved is shown in Fig. 7, and it will be seen that fixed bias was finally decided on for the pentode, derived from the 12-volt supply, with decoupling to prevent possible feedback into the rest of the counting chain. The 12-volt supply was supplemented with two dry cells, giving a total bias of —15 volts. The 0.05 microfarad capacitor connected between the grid of the pentode and ground as intended to store the input pulse and prevent differentiation of the pulse in the RC coupling.

Because the amplifier has substantial gain, the input leads were shielded to prevent triggering of the circuit by stray impulses. A potentiometer allowed adjustment for a satisfactory operating level to be made.

POWER SUPPLY

One power supply was built to power the entire unit, both transistors and valves. In the early stages the current drain at the various outlets was not known for certain, so fairly generous arrangements were made. The power supply used various components that were to hand, and may therefore appear a little unorthodox. The circuit diagram of the power supply is given in Fig. 8. Of interest is the method of obtaining the —100 volts bias supply, which is taken off across the filter choke L2. This was actually the energising winding of an old dynamic speaker, with a DC resistance of 1500 ohms.

The 150-volt regulated supply is obtained with the aid of two parallel-connected type VR150/30 gaseous regulator valves. It is not normally recommended to operate this type of valve in parallel, but the insertion of two low-value resistors respectively in the plate leads effectively balances the current flowing in the two valves, and the arrangement operates quite successfully. Although both the plate resistors are shown in the diagram as 330 ohms, and should be satisfactory at or about that value, in actual fact the values of the two resistors were varied slightly in order to compensate for slight differences between the two regulator valves and to get equal current through them. However, this would not appear to be strictly necessary.

The fact that the unit was made up of parts at hand accounts for the fact that one of the main filter capacitors is made up of two units in parallel. The same reason accounts for the parallel connection of the 7 ohm and 10 ohm resistors in the 12 volt DC supply.

ELECTRONIC SWITCH

The whole of the interval timer has now been described, with the exception of the electronic switch which is to "gate" the oscillator input to the timer during the interval to be measured. As previously mentioned, this switch could take any one of several forms, dependent on the application of the timer. It is in this field that the user will have to rely on his own ingenuity. It has already been stated that one of the applications the author had in mind when making the timer was the measurement of the speed of a bullet, and this was to be done by measuring the time taken by the bullet in flight to traverse a known distance. The electronic switch used, and its application to this specific problem, will therefore be described as an example of application of the timer.

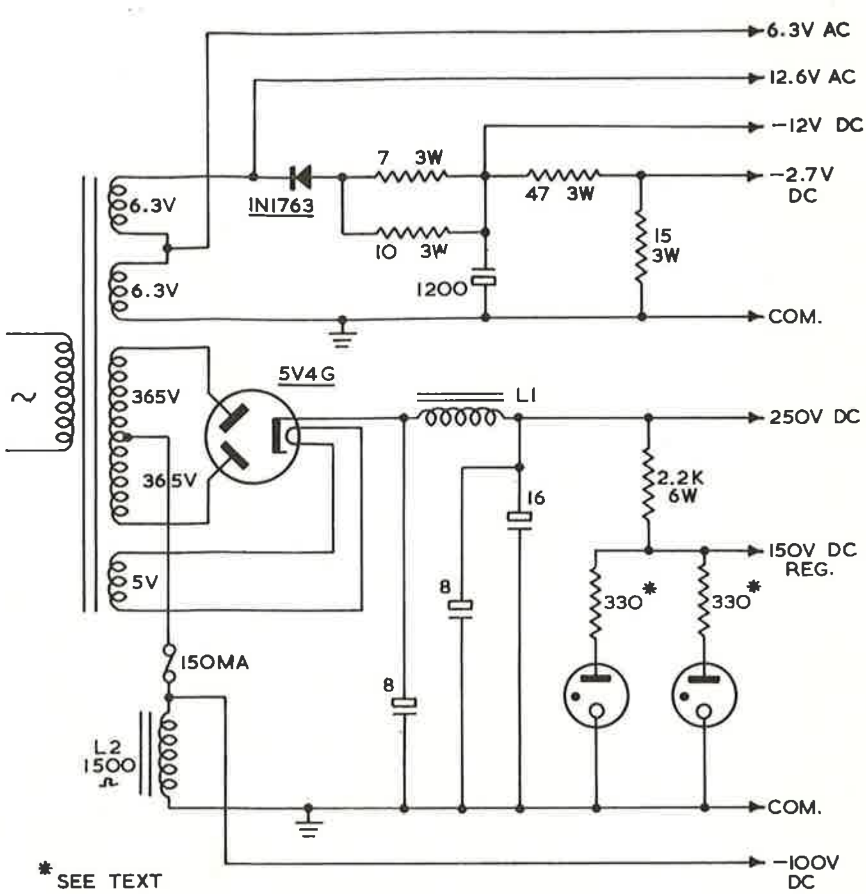


Fig. 8—Circuit diagram of the power supply.

It was decided to use a valve as the switch. The switch was required to be turned "on" by the breaking of a first circuit, and then to be turned "off" by the breaking of a second circuit. A pentode valve was selected, the suppressor grid of which was to be the switching element. The arrangement used is shown in Fig. 9. The theory of this switch is that it would be cut off in the quiescent condition, by means of a large negative bias applied to the suppressor grid. Then when the first circuit, indicated by the link 1, is broken, the valve conducts. The subsequent breaking of the second circuit, indicated by the link 2, would then restore the negative bias to the valve and return it to the cut off condition.

The arrangement described allows for the measurement of the time taken by an object to move a certain distance. It will however be clear from the diagram that either of the links shown can also be used as a simple switch to determine the time for which the timer is operative. A number of variations of the basic idea shown will also be obvious.

In measuring bullet speed, the rifle was held in a wooden crotch and so arranged that when it was fired, a fine wire forming link 1 was broken. After travelling the stipulated distance, the bullet struck a prepared target, consisting of a wooden panel on which was arranged a zig-zag pattern of thin foil, forming the link 2. On striking the target, the bullet severed the foil somewhere on the target and so broke the number 2 circuit.

CONSTRUCTION

Provided customary precautions are taken, the construction of a unit of this type presents no special difficulties. The power supply and the other sections using electron valves were constructed on the customary chassis, and space was left on the chassis for mounting the transistorised units. These latter units were constructed on small insulating panels, each

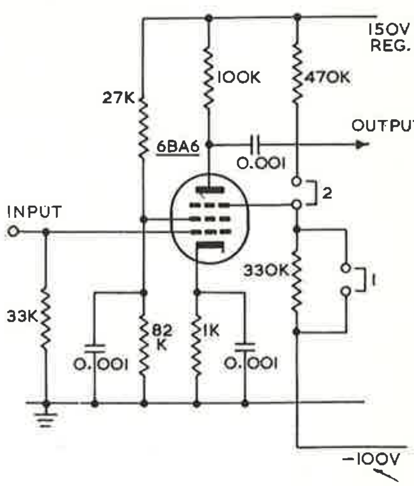
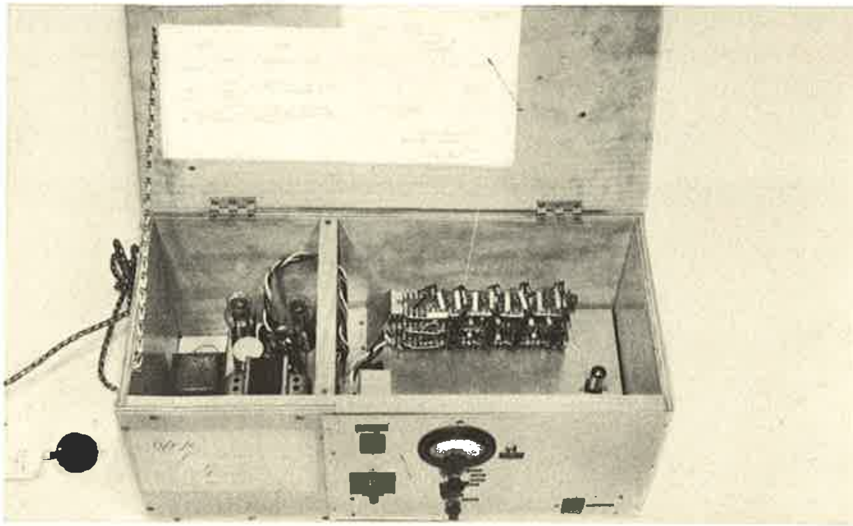


Fig. 9—Circuit diagram of the electronic switch.

forming an integral unit. The unit was housed in a wooden case with the controls, mechanical counter and milliammeter mounted in the front panel.

SUMMARY

A spokesman from AWV has commented on the project as follows:—"This forms an example of the interest and tenacity shown by some of our younger experimenters who, although hindered by lack of time and the expense of components, find a way around their problems. When one considers that the three transistorised counter units alone required some 69 semi-conductors, 52 capacitors and 120 resistors, it will be seen that this was no small project. It would be easy to say that it would have been possible to transistorise the whole unit, for example, but the basic premise was that the unit was to be made of parts that were already available or which could easily be obtained. In this context it is felt that the project merits commendation."



General view of the timing unit in its case with the power supply chassis at the left and the timer proper at the right.

Radiotronics is published quarterly by the Wireless Press for Amalgamated Wireless Valve Co. Pty. Ltd.

This publication is available at a cost of 50c per copy from the Sales Dept., Amalgamated Wireless Valve Co. Pty. Ltd., Private Mail Bag, Ermington, N.S.W.

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