RADIOTRONICS

IN THIS ISSUE

COVER:

Receiving valves being tested at AWV, Rydalmere, N.S.W.

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COMPLEMENTARY DRIVEN STEREO 10 WATT AUDIO AMPLIFIER

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The stereo audio amplifier design described in this article consists of two main sections. A direct coupled output section and a preamplifier with variable tone controls. Power output per channel is 10 watts which is supplied at a low level of distortion and with a frequency range extending within 3dB from 40Hz to 18 KHz. It has been designed to operate into a 15Ω load and when used with a suitable speaker system, e.g., the MSP bass reflex bookshelf enclosure, will be found to have a sufficient reserve of power to produce sound pressure levels which are satisfactory for a wide variety of listening conditions and programme material.

General Considerations

The ratio of peak to average sound pressure levels for programme material is in the order of twenty to one. With average speaker efficiencies, the normal level of input power to the speaker is about 100 mW and assuming a margin of approximately five to one to allow the younger listeners to make more noise occasionally, the maximum required is about 500 mW of average power. This level can require 10 watts peak and the system must produce such power without distress. This design has been developed just for such a purpose and has very "clean" overload characteristics.

Tone controls are provided to enable adjustment of frequency response, together with a simple but effective form of balance control. The correct adjustment of this or any form of balance control is simplified if a stereo-mono switch is available. The use of such a switch in the mono position will allow the production of identical signals in both channels from monophonic as well as stereo records when played with a stereo pickup.

Modern ceramic pickups for use on stereo records are robust, being little affected by high humidity and have output levels in the order of 100 mV. The complete audio system sensitivity of 300 mV for 10 watts per channel covers the pickup's output level with a margin for variations. Higher level pickups can be used but to avoid overload in the input preamplifiers the peak input signal must not exceed a maximum of 4 volts.

Note: When connecting the pickup care should be taken to ensure proper phasing or polarity. Incorrect connection may reduce output and/or seriously affect frequency response and distortion. Reference should be made to the pickup manufacturer.

Preamplifier

The complete circuit of the preamplifier is shown in Fig. 1. The first section of each channel is the pickup preamplifier and achieves correct sensitivity and equalization by the method of "matching time constants," i.e., by matching the time constant of the pickup circuit to that of the input circuit of the preamplifier. This method results in moderate levels of input

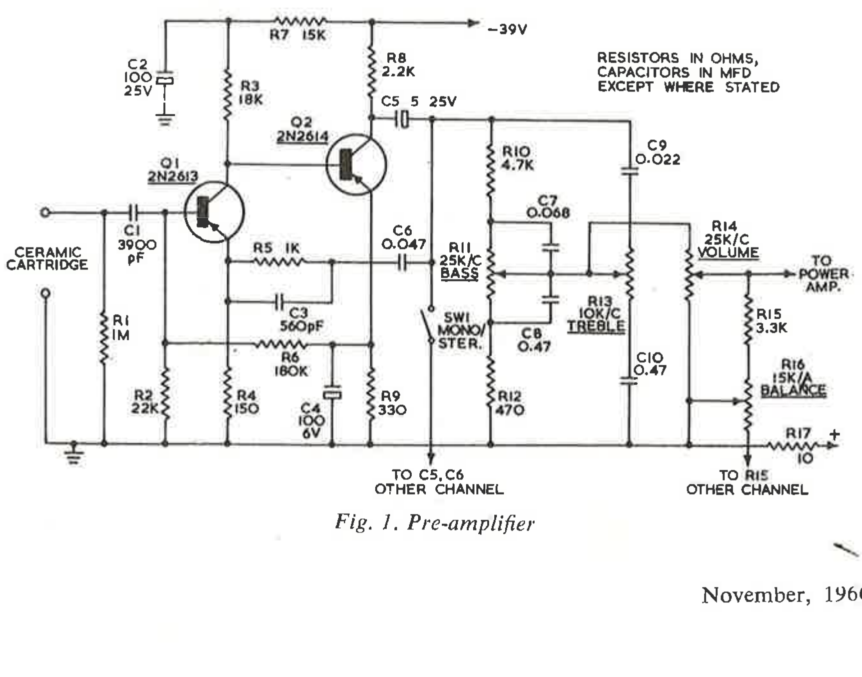


Fig. 1. Pre-amplifier

impedance, obviating problems associated with high impedance circuits, such as thermal stability, dynamic range, hum pickup, etc., and possible noise problems associated with low input impedance.

The output of ceramic pickups can be represented by a generator in series with a small capacitor, and a suitable load must be used to obtain a linear signal feed to the preamplifier. Also equalisation must be incorporated in the preamplifier so that a satisfactory signal is applied to the main amplifier. To obtain the required pickup correction and RIAA compensation, two RC time constants have been incorporated in the preamplifier. The first, at the input, consists of the pickup's input capacitance and the 20 kΩ input impedance of the amplifier. This time constant maintains constant output above 10KHz. For frequencies below 1.5 KHz another time constant is incorporated in the feedback circuit of the input pair of transistors—the combination of the 1kΩ resistor and the series capacitor between the collector of Q2 and the emitter of Q1. The 560 pF shunt capacitor across the 1 kΩ resistor provides stability in the feedback loop at high frequencies.

Tone control circuits follow the preamplifying stage and consist of two frequency selective attenuators. One for the low frequencies and the other for the higher frequencies. The fixed insertion loss at mid frequencies is 10 times in voltage and when coupled to the main amplifier provides a satisfactory output level. The output level from the preamplifier when used with a ceramic pickup is 35 mV which is sufficient to drive the main amplifier to the full output of 10 watts.

Power Amplifier

The second section of the amplifier is a direct coupled amplifier system capable of supplying 10 watts of audio power over almost the full frequency range to the 15Ω speaker on each channel. The output section of the power amplifier as shown in Fig. 2 is basically a complementary driver stage directly coupled to a single ended push-pull output stage. Both the driver stage and the output stage are operated under class B conditions.

The driver transistor, Q4, which directly drives the complementary stage also is directly coupled to the input amplifier Q3. The emitter circuit of Q3 is returned through a resistor to the collector of Q7, which is the output section of the amplifier. All sections are thus directly coupled and the input stage has a voltage, which is related to the supply rail potential, applied to its base. This voltage is fixed for any given supply potential and is used as a reference point for the system.

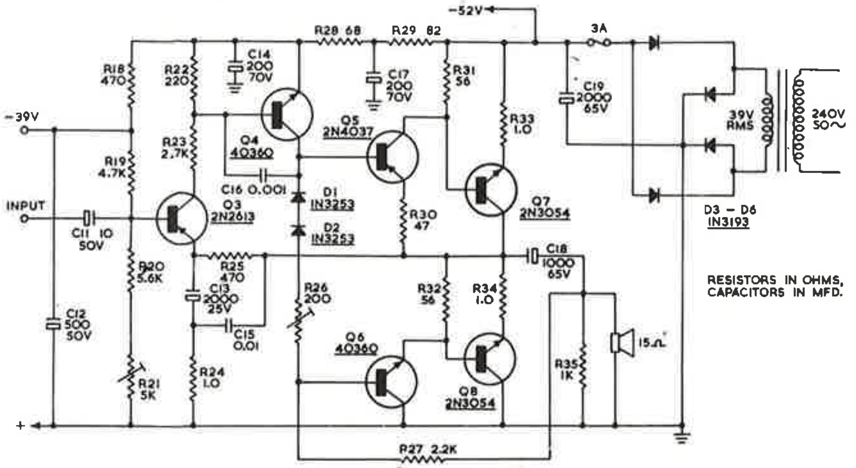


Fig. 2. Main amplifier

Frequency response of the main amplifier is determined by the values of C13 and C15. C13 determines the low frequency roll off point and C15 the high frequency point. Due to the high cut off frequency of transistors Q3, Q4 and Q5, the capacitor C16 has been included to stabilise these three transistors against parasitic oscillation at ultrasonic frequencies.

Except for the quiescent current of the output stage the system is inherently temperature stable. In the case of the output stage stability has been obtained by providing temperature compensation in the collector circuit of Q4 as shown in the circuit and a 1.0 ohm resistor in each of the output transistor's emitter circuits.

Power Supply

The power supply is simple and due to the class B operation of the output stage, must have a low internal impedance. The power transformer is rated at approx. 40 VA and has only two windings, both being untapped for simplicity. A full bridge rectifier is used and the peak current limiting resistance has been incorporated in the transformer winding resistances. This resistance should have an effective value ≮ 1.5Ω. A filter capacitor of approximately 2000μF is satisfactory since the amplifier is insensitive to ripple at this point.

Operating Considerations

Following the completion of the amplifier construction a few adjustments are necessary. These are detailed below:—

1. Before switching on, set R26 to minimum value and R21 to about midpoint, then switch on the amplifier. The output transistor's quiescent current must be adjusted to a value within the range of 30 to 50 mA, by a simple adjustment of the resistor R26. This should be checked after the amplifier has been running for some time.
2. Maximum unclipped output is obtained by adjusting the resistor R21 for equal clipping of the output waveform. Although ideally this adjustment requires the viewing of the output waveform on a c.r.o. an approx. result can be obtained if the voltage at the collector of Q7 is set to 50% of the unloaded power supply potential.
3. When the speakers are in the required position, set the "Stereo-mono" switch to "mono" and adjust the balance control. If phasing is correct, then as the balance control is rotated from one extreme to the other extreme, a distinct movement of the sound source from one speaker to the other is produced. If not, reverse the connections to one speaker only. Finally, adjust the balance between the speakers until the sound source is about midway between the speakers. Then switch to "stereo" and the system is ready for use*

Specification

The performance of each channel of the stereo amplifier is given in the following tables and is followed by two performance graphs. Unless otherwise stated the input mains voltage is 240 volts and the test frequency and reference points on the frequency response graphs is 1,000Hz. All relevant measurements are in r.m.s. values.

**Further discussion on phasing and balance will appear in a subsequent issue of Radiotronics.*

1. Power Amplifier

Nominal output power rating ... 10 Watts
Maximum ambient operating temperature ..... 50° C
Output power for THD of 10% 13 Watts
Output power for THD of 5% ... 12 Watts
Output power before clipping ... 10 Watts
Distortion before clipping 0.45%
Sensitivity at full unclipped output 35 mV
Input impedance ..... 3.0 Kohms
Supply current at an output of 10 watts ..... 400 mA
Overall feedback ..... 18 dB
Noise level below full unclipped output ..... 70 dB
Total thermal resistance, case to air (Q7, Q8) ..... 18.0° C/Watt

The graph shown in figure 3 indicates the distortion at various levels of power output at 1,000Hz. Figure 4 shows the amplifier's frequency response at all power levels up to 10 watts, measured with the tone controls set in the "flat" position. The response is substantially the same for all power reference levels.

2. Preamplifier

Nominal pickup output level* 300 mV
Output voltage at above input 35 mV

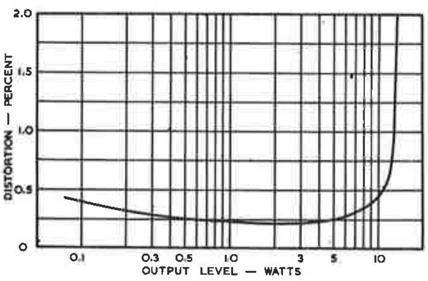


FIG 3

Bass control range at 100Hz* +9, —15 dB
Treble control range at 10 kHz* +15, —7 dB

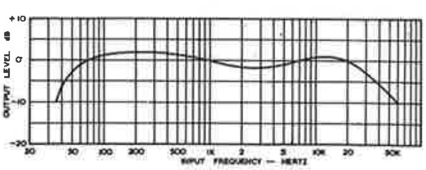


FIG 4

Maximum input voltage before overload 4.0 Volts
Distortion at 10 watts through complete system ..... 0.5% Max.

*Input voltage applied through a 680 pF capacitor

Amplifier Components List for Complete Stereo System

Resistances

All are ½-watt power rating and values are in ohms with 10% tolerance.

Circuit Reference Number	*Component Value*	*Component Description*	*Quantity Required*
R1	1.0M	Carbon	2
R6	180K	Carbon	2
R2	22K	Carbon	2
R3	18K	Carbon	2
R7	15K	Carbon	2
R20	5.6K	Carbon	2
R10, R19	4.7K	Carbon	4
R15	3.3K	Carbon	2
R23	2.7K	Carbon	2
R8, R27	2.2K	Carbon	4
R5, R35	1K	Carbon	4
R12, R18, R25	470	Carbon	6
R9	330	Carbon	2
R22	220	Carbon	2
R4	150	Carbon	2
R29	82	Carbon	2
R28	68	Carbon	2
R31, R32	56	Carbon	4
R30	47	Carbon	2
R17	10	Wire Wound	2
R24, R33, R34	1.0	Wire Wound	6

Potentiometers

R11	25K ohm	Type C Log Taper. Dual matched
R13	10K ohm	Type C Log Taper. Dual matched
R14	25K ohm	Type C Log Taper. Dual matched
R16	15K ohm	Wire Wound Linear. Single
R21	5K ohm	Preset Carbon Trimpot
R26	200 ohm	Preset Wire Wound Trimpot

Capacitors

Circuit Reference Number	*Component Value*	*Component Tolerance*	*Component Description*	*Voltage Rating Minimum*	*Quantity Required*
C19	2000 μF	—10, + 50%	Electro*	65 V.W	1
C13	2000 μF	—10, + 50%	Electro*	25 V.W	2
C18	1000 μF	—10, + 50%	Electro*	65 V.W	2
C12	500 μF	—10, + 50%	Electro	50 V.W	2
C14, C17	200 μF	—10, + 50%	Electro	70 V.W	4
C2	100 μF	—10, + 100%	Electro	25 V.W	2
C4	100 μF	—10, + 100%	Electro	6 V.W	2
C11	10 μF	—10, + 100%	Electro	50 V.W	2
C5	5.0 μF	—20, + 100%	Electro	25 V.W	2
C8, C10	0.47 μF	20%	Polyester	100 V	4

C7	0.068 μF	20%	Polyester	100 V	2
C6	0.047 μF	20%	Polyester	100 V	2
C9	0.022 μF	20%	Polyester	100 V	2
C15	0.01 μF	20%	Polyester	100 V	2
C1	0.0039 μF	20%	Polyester	100 V	2
C16	0.001 μF	20%	Polyester	100 V	2
C3	560 pF	20%	Ceramic	100 V	2

NOTES—The above electrolytic values are minimum values and can be increased if desired. The units marked with an asterisk (*) would need to be chassis mounting type due to their physical size. The voltage ratings are minimum values and can be increased up to 100 per cent. if required.

Transistors and Diodes

Circuit Reference Number	*Type Number*	*Remarks*
Q1	2N2613	—
Q2	2N2614	—
Q3	2N2613	—
Q4	40360	—
Q5	2N4037	—
Q6	40360	—
Q7	2N3054	C/W mounting accessories and mica washer
Q8	2N3054	C/W mounting accessories and mica washer
D1, D2	1N3253	C/W flag type heat sinks
D3, D4, D5, D6	1N3193	

Sundry Components

Power transformer:

Primary—240 volts 50 Hz
Secondary—40 volts 0.8 amp. } M.S.P. Part No. H2649

Fuse:

Cartridge fuse at 3 amp. rating.

Speaker enclosure:

M.S.P. type 52873/8TAC/15 and 52874/4MBC/15 speakers mounted in an A.W.A. enclosure type PA680 or PA682.

Chassis:

Constructed of 16-gauge aluminium with output transistors mounted upon top of chassis. Diodes D1 and D2 are each mounted in a flag heat sink which is fixed on the chassis between their associated output transistors. A $\frac{1}{8}$-inch air space should be provided between the power transformer and the chassis.

IMPORTANT ANNOUNCEMENT

Effective immediately we are pleased to announce that in response to the wishes of our many readers we are again accepting subscriptions to this publication.

Radiotronics will be published four times each year during the months of February, May, August and November and subscriptions will only be received for one calendar year January/December, at a rate of $2.00 per annum.

Readers requiring individual copies are advised that these are available at 50 cents per copy from: Sales Department, Amalgamated Wireless Valve Co. Pty. Ltd., Private Mail Bag, Ermington, N.S.W.

Design Considerations for Speaker Enclosures

In the quest for the ideal loudspeaker enclosure several factors need to be compromised. Among these are cost, size, aesthetic appeal and, of course, sound quality. An enclosure is described which meets the general requirements of the Hi-Fi enthusiasts yet is relatively cheap and easy to construct.

TECHNICAL REQUIREMENTS

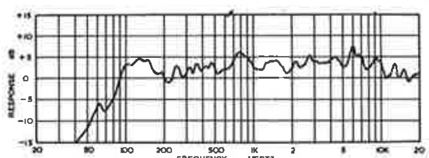
The technical requirements for a loudspeaker enclosure system to be able to give satisfactory reproduction of musical and oral sound are—

- Adequate frequency response, extending well into the high as well as the low frequency areas.
- Smooth response, free from sharp peaks and dips, which usually indicate sub-resonances.
- Damping of the fundamental resonance, which is one cause of boom in closed and reflex enclosures.
- Adequate freedom from excessive transient hangovers due to sub-resonances, such as rim resonance and standing waves in the cone.
- Low harmonic distortion and intermodulation distortion due primarily to non linearities in suspension and magnetic flux.
- Adequate power handling and sensitivity for the user's conditions and equipment.

The enclosure combination described was designed using these principles. The speakers selected for the enclosure are M.S.P's new ceramic Hi-Flux speakers 52873/8TAC/15 for the woofer-mid-range and 52874/4MBC/15 tweeter. The 8" cone has a reasonable area to compromise for the change in resonance when the speaker is placed in a small enclosure, while sensitivity derived from using this speaker is still quite high. The 15Ω voice coil impedance is standard for most Hi-Fi amplifiers in use.

Examination of the impedance responses of the 8" unit revealed a resonance of 70 Hz with a Q of 0.7. This when placed in the enclosure, which was tuned by a ducted vent and damped with absorbent material, resulted in a response which was very smooth from the bass end through the middle and into the highs. It was decided to cut off the 8TAC at 3KHz and to bring in the tweeter, whose range is up to 20KHz, using the crossover network as illustrated. The resulting frequency response as shown was taken in a room with a microphone in front of the tweeter, and satisfies the first and second requirement for good reproduction. Damping of the fundamental resonance is achieved magnetically by using a ceramic magnet, which gives high flux (11,000 gausses), and acoustically by three layers of "Innerbond" loosely attached to the rear panel. This results in clean, crisp transients, free from annoying hangover which "distorts" the sound. In testing, no excessive higher frequency hangover was found to exist, due to the cone design and the materials used in the construction of the loudspeaker and enclosure. Voice coil and suspension design ensured that harmonic distortion was minimised, while the use of two speakers with crossover limited the usually prevalent and prominent intermodulation distortion. The large high-flux ceramic magnet and the use of a one inch voice coil diameter ensured high efficiency and heat dissipation, while the reflex action of the enclosure boosts the bass response and also ensures minimum cone excuesion in the bass region, enhancing the power capabilities of the system and allowing a full 10 watt input to be used.

The resulting performance on listening test was very pleasant, the response giving an impression of totality to the reproduction while not being peaky or ringing. Upon comparison with other (more expensive) units the results were most favourable, the smoothness being the most apparent quality.



CONSTRUCTION

Construction is simple, using butt or mitred joints, and cleats to secure front and rear panels. The rear panel is braced to avoid flexure and vibration, while the port duct is simply rigid tubing glued into the front panel. Alternatively, the duct may be constructed from ½" timber to give the same total duct length (2½"), the vent being a rectangular aperture 3" x 2". Timber should be at least ½" thick, veneered particle board giving a most professional finish. Fret cloth needs to be of the open-weave type, coloured hessian giving a modern look enhancing any decor. The enclosure need not be air tight, good solid joinery being all that is required. The chamfer lowers the air loading on the light tweeter cone while increasing the dispersion of the high frequencies, eliminating piercing sounds often associated with tweeters.

The crossover network is an important part of the system since it controls the response over an octave. The inductor by itself cuts the 8TAC off at 6KHz, but an additional effect is decrease in the upper mid range (3-6KHz). This is offset by the use of the 10uF capacitor, which as well as reducing the cut-off to 3KHz and making it sharper, resonates with the voice coil inductance and lifts the upper-mid-range response to the mid-range (0.1-1KHz) level, greatly improving the resultant performance.

The size of the enclosure has been designed as a small floor-mounting enclosure or a medium size book-shelf type, operating in either application with equal satisfaction.

AVAILABILITY OF PARTS

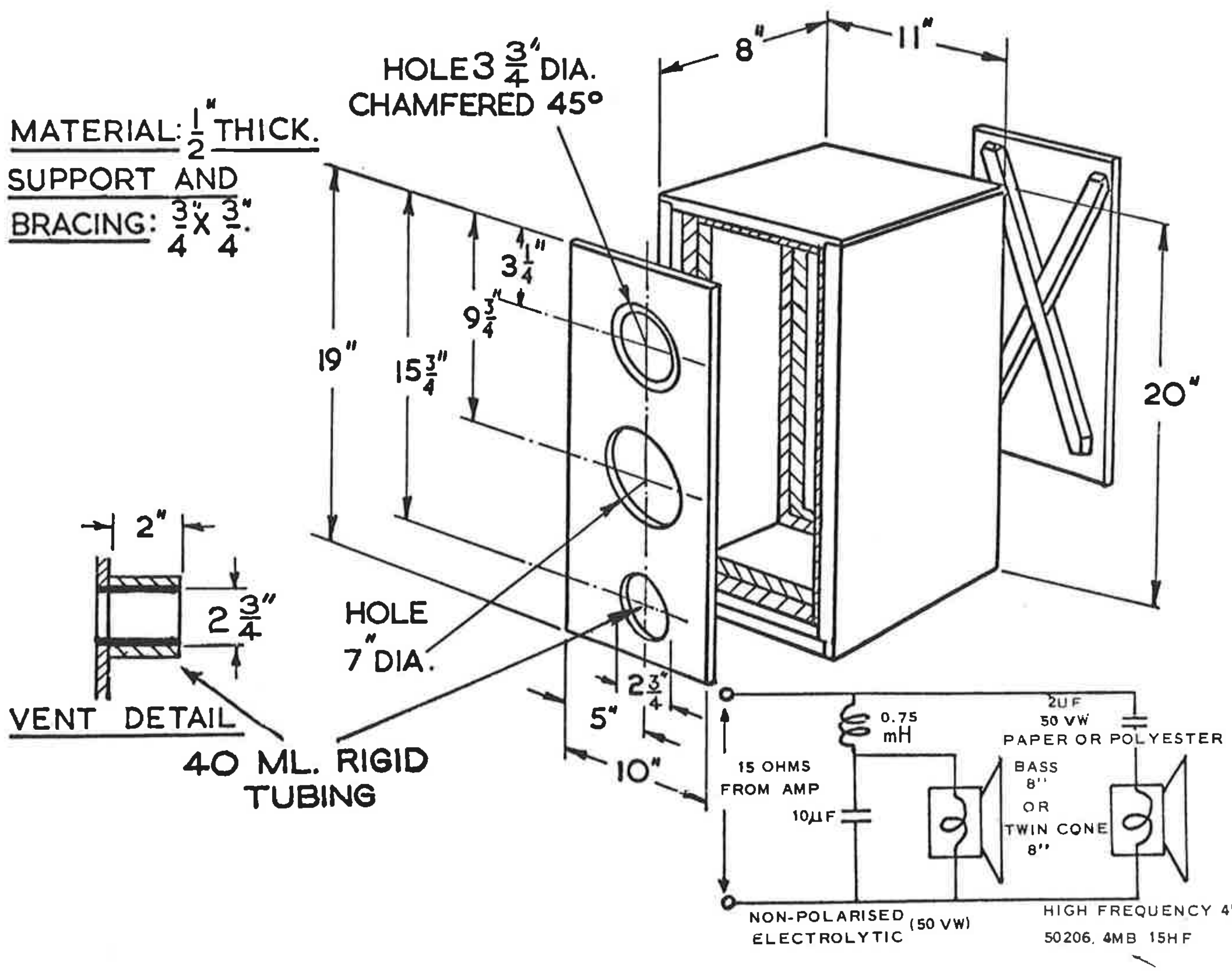
The speakers may be ordered through recognised trade and retail channels.

Capacitors 2 uf . . . 2—10μF 160 V polyester in parallel may be used, or 2 uF 200V AEE type W48 may be ordered through the trade stores.

10 uF, non polarised electrolytic capacitors are not at present available. However, 2X 20 uF, 100 VW type ET3CNP may be ordered through the trade channels. Used in series these give the required value. The inductor is 0.75 mH, air cored. This may be wound on a bobbin 1" x 1" x 1¼" long using 185 turns of 20 gauge enamelled wire. These are available as M.S.P. Part No. 53940 through recognised trade sources. "Innerbond" is available from Wonder Wool Pty. Ltd., 87 James St., Leichhardt, N.S.W., or through major electrical suppliers.

If use of a twin-cone speaker is desired, thereby eliminating the tweeter and crossover network, a suitable replacement is M.S.P. type 52855/8TACX/15.

For those who would prefer to purchase a complete enclosure at very reasonable cost, two versions of this box are available as types PA680 and PA682 from the Consumer Products and Engineering Products divisions of Amalgamated Wireless (Australasia) Pty. Ltd. This enclosure has been specifically designed to complement the solid state amplifier introduced in this issue, and together with a suitable turntable, they form an audio system to satisfy the most demanding enthusiast.



BOOK REVIEWS

"FUNDAMENTALS OF RELIABLE CIRCUIT DESIGN," Mel Xlander. Iliffe Books Ltd. 2 Volumes, 196 pages, 138 pages. Size: $8\frac{1}{2}''$ x $5\frac{1}{2}''$.

The purpose of this two-volume book is to introduce students and technicians to basic circuit design and to provide them with practice in solving the type of problems which occur in industry. This is one of the few books available that bridge the gap between the theoretical and practical approach to the subject. Emphasis is placed on the intuitive understanding of electronic components rather than on memorising formulae, and only the basic and most frequently encountered types of problems are examined, so that a sound foundation in the fundamentals of reliable circuit design can be established.

Volume 1 introduces the five basic elements in electronics—voltage sources, current sources, resistance, capacitance and inductance — on which all components, however complex, rely. Elementary circuits are designed from these basic linear elements and the laws governing the relations between the elements in the circuits are discussed. Although ideal circuits are considered first, in order to acquaint the reader with the basic tools of circuit analysis, the question of tolerance is introduced at an early stage. This leads to the effect that such tolerances have on circuit design and to the policy of "worst-case design."

Volume 2 considers the diode and the transistor as ideal components and as non-linear devices. Several specimen data sheets are supplied for both diodes and transistors, and the reader is shown how to interpret these specifications when designing circuits. Gradually, a series of electronic circuits is built up and the functions and inter-relationships of the components are discussed, again showing the effect of their various tolerances. Eventually this leads to the design and drawing up of engineering specifications for circuits.

"MICROWAVE VALVES," C. H. Dix and W. H. Aldous. Iliffe Books Ltd. 269 pages. Size: $8\frac{3}{4}''$ x $5\frac{1}{2}''$.

In recent years, solid-state devices have taken over many of the functions previously carried out by thermionic valves, and this trend is currently continuing. However, for the generation and amplification of power at microwave frequencies, valves are at present the obvious choice and usually the only possibility.

Although many books have been written on microwave valves, the majority are for advanced students in this field and are specialised in their scope. This book is intended to fill a gap in the existing literature by presenting readers educated to graduate or British H.N.C. level with a readily absorbed account, which includes only essential mathematics, of the basic physical processes and operation of microwave valves.

The approach to the subject is a fundamental one, starting with a description of the motion of electrons in electric and magnetic fields and the properties of the various types of RF circuits and transmission lines that are used in the devices. Microwave triodes are discussed, but the emphasis is on beam devices, both linear and crossed field, and in describing these the space-charge wave approach is used consistently. Further chapters cover the formation and focusing of electron beams, the noise properties of microwave devices, construction and applications. Suggestions for further reading at the end of each chapter provide the source of an extensive further bibliography.

NEWS & New Releases

New EEV TR/Limiter Cell Type BS808

English Electric Valve Company Limited announces a new X-Band TR/Limiter cell combination, type BS808, with peak operating powers of up to 200 kW. It requires no external trigger and is designed for use in radar equipment which uses very sensitive crystals or tunnel diode amplifiers.

The BS808 consists of a gas-discharge TR cell and a solid-state limiter fitted together. It is electrically and mechanically interchangeable with a standard X-Band TR cell.

Important features of the BS808 are:—

Very low spike leakage—less than 0.01 erg/pulse (measured at 40 kW peak power, 1.0 μSec pulse, 1000 p.p.s.).

Full passive protection.

Elimination of Crystal burn-out.

EEV Klystrons Power Pye U.H.F. TV Transmitters For Switzerland

English Electric high-power klystrons, type K370, have been chosen by Pye T.V.T. Ltd. for the U.H.F. television transmitters that they are supplying for the Swiss Post and Telegraph Administration transmitting station at Mount Rigi in Central Switzerland.

The two Pye 10 kW transmitters to be installed at the Rigi station, are valued at approximately £100,000.

The EEV K370 klystron has a peak synchronising signal power output of 10 kW, saturated output typically 1 dB higher, with a frequency range of 470-610 Mc/s. The minimum gain of the K370 is 40 dB. These klystrons are vapour and forced-air cooled and have been supplied to Pye complete with circuit assemblies comprising mount, focusing coils, cavities and boiler units.

Full information on the EEV K370 U.H.F. power klystron and others specifically designed for television service, may be obtained from the address below.

A COMPACT MULTISTAGE INDUSTRIAL PROCESS CONTROL TIMER

D. N. V. MacDonald,* Affil. I.R.E.E. (Aust.)

The unit described in this short note may be of interest to readers, both as an example of the industrial application of electronics and as a unit which may be useful to themselves. Even if the precise form of the unit as described herein does not suggest the use that a particular reader may have in mind, the modification of the basic unit to provide a variety of timed programmes is a simple matter and will present no trouble to the experimenter.

This particular timing unit was designed to fulfil a requirement in the AWV Rydalmere Factory which called for a four-stage sequence of adjustable times but, as already mentioned, it could find an application wherever a compact versatile timer is required for the control of complex switching functions or where the requirement is for a series of accurately timed intervals.

OPERATION

The circuit shown in the accompanying diagram consists essentially of a grid-controlled tetrode thyratron, type 2050A, operated in a timing circuit, in which a series of resistance-capacitance elements form the timing elements and control the thyratron. The load for the thyratron is a dc relay, controlling an industrial-type stepping switch which has a strong similarity to the stepping selectors used in automatic telephone exchanges. A typical modern stepping switch is shown in figure 1, which is reproduced by courtesy of Automatic Electric Telephones Pty. Ltd., 86 Holdsworth Street, Woollahra, N.S.W. The stepping switch in this case has six switching arcs, in each of which a wiper arm is driven over a series of switch contacts. One of the switching arcs or "levels" serves as part of a "homing" circuit, which operates to return the switch to a normal rest or "home" position. Switches of this kind are, in general, as in this case, operated by a series of dc pulses wherein each pulse steps the switch forward to the next contact.

*Kinescope Manufacturing Section, AWV.

The stepping switch contacts in the circuit diagram are shown in the rest or "home" position. For reasons which will become apparent later in the discussion, the operational contacts of the switch bank, that is, the contacts which actuate and control the timing functions, are alternate contacts in the bank. The sequence of operations from the position shown in the diagram will, therefore, be home, stage 1, step on, stage 2, step on, stage 3, step on, stage 4, and then the remaining contacts are all wired for step on. In the contacts here referred to as step on contacts, the arrangement is such that when the switch arrives at such a contact it automatically and immediately steps forward to the next contact. Ignoring for the moment the timing function that takes place at the four stages, the

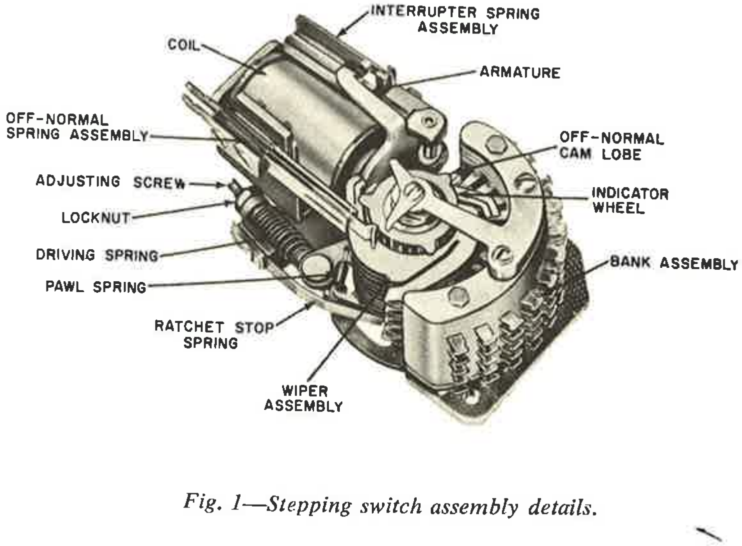


Fig. 1—Stepping switch assembly details.

basic operation is that once the stepping switch is moved forward from the "home" position, it will step sequentially through all the contacts until it arrives back at the "home" position, where it will rest until the next timing sequence is initiated.

The basic operational requirement of the switch just described arises from the following considerations, in which again the timing function of the four stages is temporarily ignored and the function is regarded as a simple step on. The number six group of contacts on the stepping switch is the HOMING LEVEL. This group of contacts, together with the stepping coil and interrupter contacts shown below it, controls the whole switch except in the four functional stages. It will be seen that, with the exception of contact 1, all contacts on level six which do not correspond to one of the four functional positions are commoned and complete a circuit from the energising supply through the interrupter contacts and the stepping coil.

It should be noted that level 6 is a "bridging" level, i.e., the wiper is a make-before-break type, which minimises contact arcing when the level is used for the homing function. The arrival of the switch at one of these contacts therefore energises the stepping coil, which causes the drive mechanism to engage. At this point, the interrupter contacts are opened, thus releasing the coil and allowing the drive spring to step the switch to the next position. The absence of a connection to contact one of level six provides a "home" position, because when the switch arrives at this position no circuit is made through the stepping coil, and the switch must await the initiation of a further cycle of operation. It will be clear that if a connection is provided to contact 12, then the switch will continuously sweep through the series of functions, automatically recycling itself. In this unit, the recycling connection is provided by the START button, which releases the switch from the "home" position and starts the cycle. It will be noted from the circuit diagram that the start button is wired in series with one of the "OFF-NORMAL" contacts on the stepping switch. This provides that the start button will only function when the stepping switch is in the "home" position, which is the characteristic desired in this particular application. If the button is wired directly to the coil, then in addition to the start function, pressing it momentarily during a cycle will cause the stepping switch to skip to the next stage, a feature which may be desirable in other applications of this device.

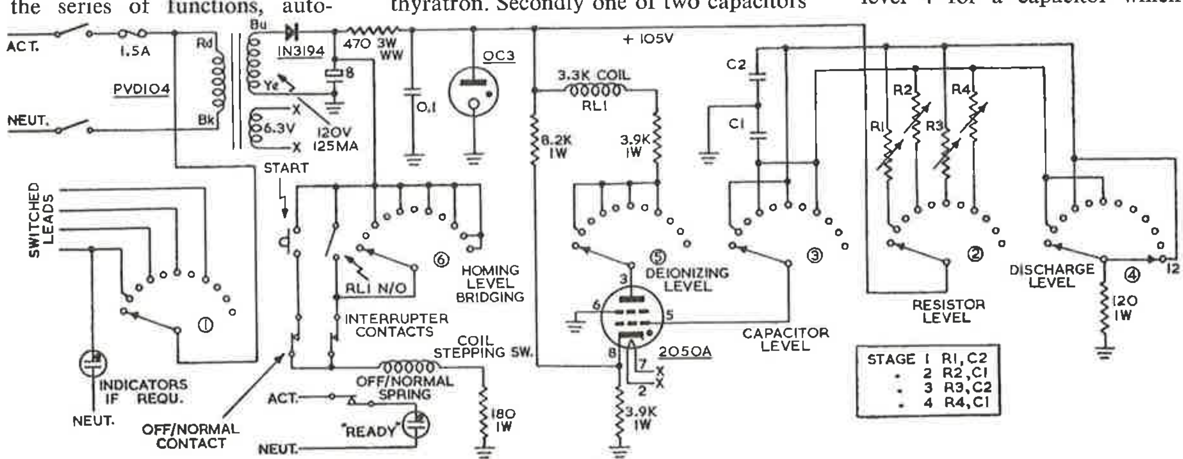
Switch level one provides switched leads at the timing positions, which leads are used as required to control the external functions that are to be timed. Indicators for these leads may be provided if required. A further indication is already provided in the READY lamp, which is energised through an auxiliary contact set on the stepping switch, this contact set being open except when the switch is in the home position. It will be clear from the circuit diagram that with the stepping switch used and the circuit arrangement adopted, provision could have been made for five timed functions; however, only four were required in this case.

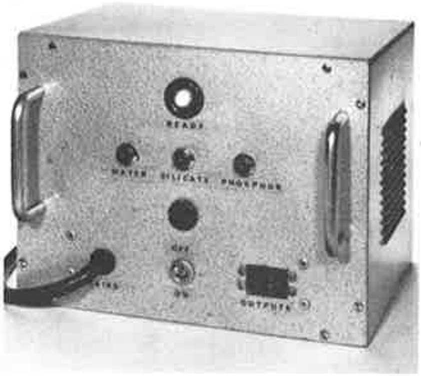
It will now be necessary to consider the timing function that takes place at each of the four timed positions. From the diagram it will be seen that at each of these positions, which are contacts 2, 4, 6 and 8 on the bank, several operations take place. Firstly, a circuit is completed through the control relay coil from the power supply to the plate of the thyratron. Secondly one of two capacitors C1 or C2 is connected from the grid of the thyratron to ground. Thirdly a preset potentiometer is connected from the power supply to the capacitor at the grid of the thyratron so as to charge the capacitor.

The thyratron is biased by reason of the fact that the cathode is returned to the junction of two resistors which form a fixed voltage divider across the supply. Because the valve is not drawing current at this stage, the voltage at the cathode can be calculated from the values of the two resistors. With the plate connected to the supply and the charging capacitor connected to the grid of the valve, the timing function is provided by the time taken for the capacitor to charge substantially to the same potential as the bias potential at the cathode of the valve whereupon the valve fires, energises the control relay which in turn actuates the stepping switch coil, and steps the switch to the next position. The timing duration is determined by the value of the charging capacitor and the preselected value of the charging resistor.

Some explanation is required of the arrangement of the resistors and capacitors in the timing circuits, as it will be seen that whereas four different preset resistor values are used, only two capacitors are used, and these are both the same value. Two factors dictate this arrangement, which in other cases could be varied to suit conditions. Firstly, with the value of capacitor chosen, the required timing delays can be achieved within the range of the potentiometers. Secondly, the arrangement is such that each capacitor is alternately switched into the grid circuit of the valve on successive timing periods, the reason for which will now be seen.

Arrangements are provided in switch level 4 for a capacitor which has just





General view of the timer housed in a small instrument case. The titles shown on the unit refer to stages in the process being conducted in the factory.

played a part in the timing function to be discharged, so that when it is next switched into the timing circuit, there will be no residual charge remaining which could affect the timing of the circuit, thus ensuring uniform operation. The use of the two capacitors simplifies the arrangement.

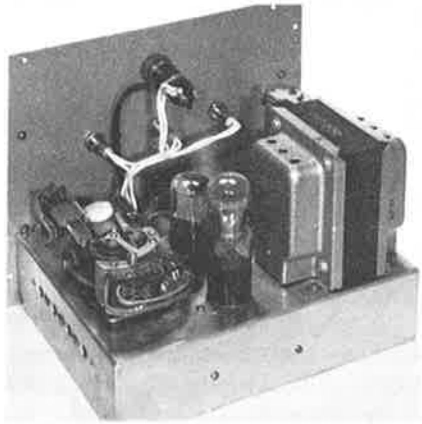
It was mentioned earlier in this note that the functional contacts of the stepping switch, that is, those contacts which operate external functions which have to be timed, are alternate contacts on the bank. The reason for this is seen in level 5, the DEIONIZING LEVEL, where, between timing functions, the plate supply to the thyratron is interrupted, and a blank position is provided between each operational position. This arrangement ensures that the thyratron has ample time to deionize and become available for the next timing function, and that sufficient time is allowed for the actuating contacts of the control relay RL1 to restore themselves ready for the next impulse.

POWER SUPPLY

The choice of a power supply for such a unit will be influenced by specific conditions. In this case a simple mains-operated power supply was provided, consisting of a step-down mains transformer, half-wave rectifier and very simple filtering arrangements. So that mains voltage fluctuations will not influence the timing accuracy, the supply to the thyratron and the timing circuits is stabilised by means of an OC3 gaseous regulator valve. Current-limiting resistors are used in series with the actuating coils of the stepping switch circuitry. In this case it will be seen that the primary fuse protection afforded the unit also protects the external circuits which are controlled through the bank level number one.

SUMMARY

This note has described a unit admirably adapted to the purpose in mind, and of great reliability. Several such units have been in daily operation in the factory for over eighteen months without failure. Whilst the specific arrangement used in the factory is the one that has been described here, it is clear that there are many variations on the basic arrangement that could be made to suit particular applications. The four timing stages provided in this unit allow for four timing intervals of up to ten seconds, the timing intervals could be increased up to a practical limit of several minutes per stage. The limitation on the switching arrangement is only that dictated by the number of contacts available on the stepping switch.



The general layout of the unit.

ADDITIONAL FEATURES

By fitting an extra normally open contact to the dc control relay, the unit can be arranged to provide a series of impulses at timed intervals, rather than the timed stages as at present. Further, the use of a relay (control) with a faster release time than the standard type used would permit, if necessary, the use of all the contact positions on the stepping switch as actuating contacts.

REFERENCE

Rotary stepping Switches, Technical Bulletin 961-473, Automatic Electric Co., Illinois, U.S.A., 1961.

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INTEGRATED CIRCUITS APPLICATION NOTE

Application of the RCA-CA3002 Integrated-Circuit IF Amplifier

The RCA-CA3002 integrated-circuit if amplifier is a balanced differential amplifier that can be used with either a single-ended or a push-pull input and can provide either a direct-coupled or a capacitance-coupled single-ended output. Its applications include RC-coupled if amplifiers that use the internal silicon output-coupling capacitor, video amplifiers that use an external coupling capacitor, envelope detectors, product detectors, and various trigger circuits.

The CA3002 features all-monolithic silicon epitaxial construction designed for operation at ambient temperatures from −55 to 125°C, and contains a built-in temperature-compensating network for stabilization of gain and dc operating point over this operating-temperature range. It is supplied in a 10-terminal TO-5 low-silhouette package.

Because the CA3002 is a balanced differential amplifier fed from a constant-current source, it makes an excellent controlled-gain amplifier. The gain-control function may be extended to include video gating, squelching, and blanking applications. Envelope detection can be achieved by suitable biasing of the emitter-base diode of the output emitter-follower transistor. Product detection can be obtained by re-insertion of the carrier at the base of the constant-current-source transistor. Various trigger and waveform-generating circuits can also be achieved by the addition of suitable external components.

CIRCUIT DESCRIPTION AND OPERATING MODES

Fig. 1 shows the circuit diagram and terminal connections for the CA3002 integrated circuit. The circuit is basically a single-stage differential amplifier (Q_2 and Q_4) with input emitter-followers (Q_1 and Q_5), a constant-current sink (Q_3) in the emitter-coupled leg, and an output emitter-follower (Q_6). A single-ended input is connected to terminal 10 or a push-pull input to terminals 10 and 5. A single-ended output is direct-coupled at terminal 8 or capacitance-coupled at terminal 6. Terminals 5 and 10 must be provided with dc returns to ground through equal external base resistors. The emitters of the differential pair (Q_2 and Q_4) are connected through degenerative resistors (R_3 and R_4) to the transistor current source (Q_3). The use of these resistors improves the linearity of the transfer characteristic and increases the signal-handling capability.

Transistor Q_1 provides a high input impedance for the if amplifier. Transistor Q_5 preserves the circuit symmetry, and also partially bypasses the base of Q_4. Additional bypassing can be obtained by connection of an external capacitor between terminal 5 and ground. The emitter-follower transistor Q_6 provides a direct-coupled output impedance of less than 100 ohms.

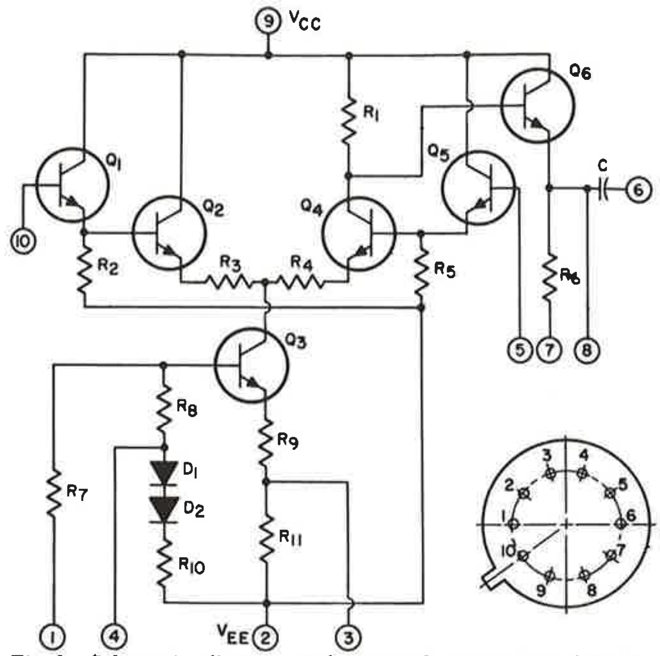


Fig. 1—Schematic diagram and terminal connections for the CA3002 integrated-circuit if amplifier.

When voltage supplies are connected to the CA3002, the most positive voltage must be connected to terminal 9 and the most negative voltage to terminal 2 (internally connected to the substrate and case). The CA3002 may be operated from various supplies and at various levels. Operation from either single or dual power supplies is feasible. When two supplies are used, they may be either symmetrical or non-symmetrical. When both positive and negative voltage supplies are used, external components can be minimized, as shown in Fig. 2(a). For single-supply applications, a resistor divider and a bypass capacitor must be added externally, as shown in Fig. 2(b). The current through R_2 and R_3 should be greater than one milliampere. Except in applications that use inductive drive, equal external base resistors must be added at terminals 5 and 10 to provide base-current returns. Terminal 7 can be connected to ground, or to the negative supply if a larger negative-going voltage swing is desired at any operating point.

For either single or dual supplies, the operating current in transistor Q_3 is determined by the bias voltage between terminals 1 and 2. The more negative point of this bias voltage must be connected to terminal 2. For dual-supply systems, terminal 1 is usually referenced to ground.

For any given bias voltage (V_{EE} when terminal 1 is grounded), four operating modes are possible, as described in Table I. In general, each mode is characterized by (1) a distinct dc operating point with a characteristic temperature dependence, and (2) a particular value of gain that has a distinct temperature dependence.

When the diodes are utilized in the bias circuit (modes A and C), the current is essentially dependent on the temperature coefficient of the diffused emitter resistors R_9 and R_{11}, and has a tendency to decrease with increasing temperature at a rate independent of the negative supply voltage. The

Table I—Identification of CA3002 Operating Modes

Operating Mode	*Shorted Terminals*	*Condition of Diodes*	*Q_3 Emitter Resistor*
A	none	in	$R_9 + R_{11}$
B	4-2	out	$R_9 + R_{11}$
C	3-2	in	R_9
D	4-3-2	out	R_9

temperature coefficient of the diffused collector resistor R_1 is the same as that of the emitter resistor, and a constant collector-voltage operating point results at the collector of transistor Q_4. However, the operating point at output terminal 8 is modified by the base-emitter voltage drop of transistor Q_6 and its temperature dependence. Typical variation of the output operating point with temperature is shown in Fig. 3 for the four operating modes for V_{EE} supplies of -3 and -6 volts. The voltage between terminals 8 and 9 is denoted by V_x. In mode B (with the diodes out of the bias circuit), it should be noted that the output operating point is constant with temperature because the change in the collector operating point is cancelled by the change in the base-emitter voltage drop (V_{BE}).

When the diodes are out of the bias circuit, the current-temperature curves become dependent on the negative supply voltage. Therefore, the value of V_{EE} can be adjusted so that the transconductance decreases, increases, or remains constant with temperature. As shown in Fig. 4, the gain increases with temperature for a -3-volt V_{EE} supply, but decreases with increasing temperature for a -6-volt V_{EE} supply. At some intermediate value of V_{EE} (approximately -4.5 volts), the gain should be constant as a function of temperature. In any case, however, a constant ac gain with temperature is accompanied by a change in the collector operating point of transistor Q_4.

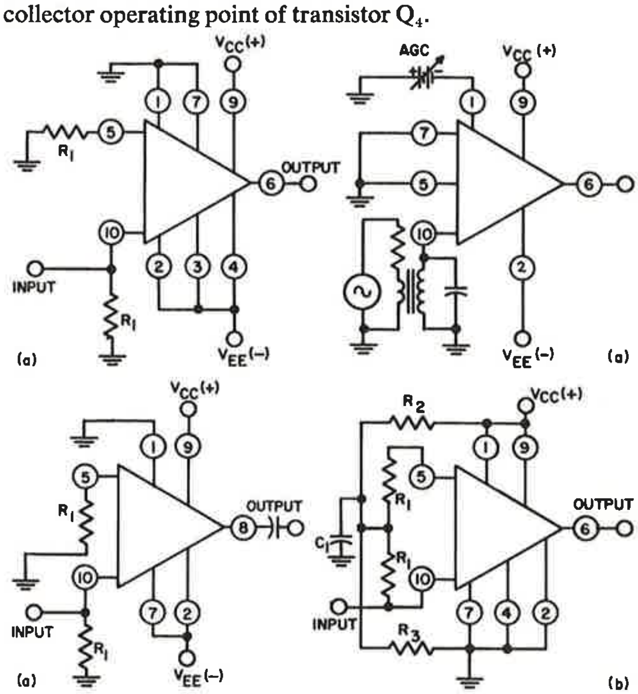


Fig. 2—Circuit configurations for the CA3002 with (a) dual voltage supplies, and (b) a single supply.

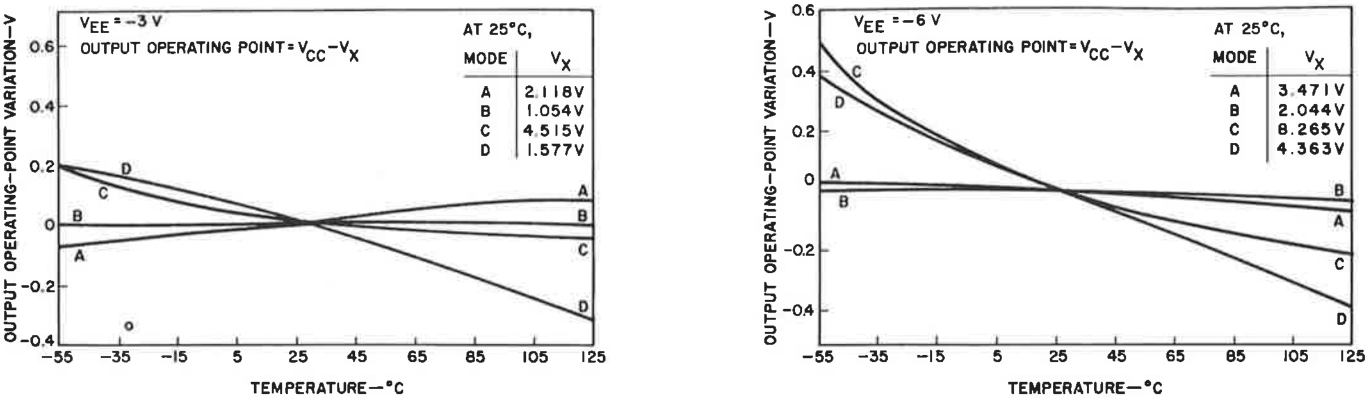


Fig. 3—Output operating-point variation of the CA3002 (normalized to the 25°C operating point) as a function of temperature with V_{EE} supply voltages of −3 and −6 volts.

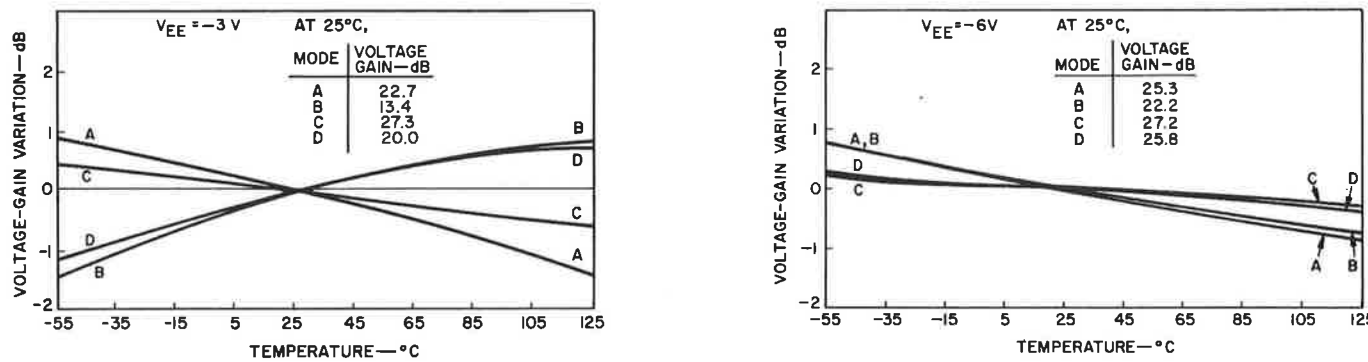


Fig. 4—Voltage-gain variation of the CA3002 (normalized to the 25°C voltage gain) as a function of temperature with V_{EE} supply voltages of −3 and −6 volts.

Table II lists typical design performance characteristics for the four operating modes of the CA3002. By use of the data in this table and in Figs. 3 and 4, it is possible to select the proper operating mode to provide the most transconductance per milliwatt of dissipation, the specified output-swing capability, and the desired temperature performance for a particular design requirement.

In operating mode C, a valid non-saturated operating point may be obtained by use of non-symmetrical voltage supplies. For example, when V_{EE} is −3 volts, the operating point will not be in saturation if a positive supply voltage of 4.5 volts or more is used (as indicated by Fig. 3). Resistor R_6 may then be returned to the negative supply instead of to ground to ensure the desired negative swings.

Table II—Typical Design Performance Characteristics for the Four Operating Modes of the CA3002 (Terminals 7 and 1 are grounded; temperature = 25°C)

Mode	*± Supply Volts*	*Output Operating Point (Volts) at Terminal 8 to Ground*	*Voltage Gain (dB) at 1 Mc/s*	*+ Supply Current (mA)*	*− Supply Current (mA)*	*Power Dissipation (mW)*
A	6	2.6	26.4	5.0	4.2	55.2
B	6	3.8	22.5	4.7	3.7	50.4
C	6	0		(transistor Q_4 saturated, transistor Q_6 cutoff)		
D	6	1.8	25.4	5.1	4.9	60
A	4.5	2.0	24.0	3.6	3.0	29.7
B	4.5	3.0	19.8	3.4	2.6	27.0
C	4.5	0		(Q_4 saturated, Q_6 cutoff)		
D	4.5	1.8	24.5	3.7	3.3	31.5
A	3	1.1	22	2.3	2.0	12.9
B	3	2.0	14.5	2.1	1.5	10.8
C	3	0		(Q_4 saturated, Q_6 cutoff)		
D	3	1.5	20	2.2	1.9	12.3

CHARACTERISTICS

Input Unbalance Current. The input unbalance current of the CA3002 is defined as the difference between the currents flowing into the base input terminals 10 and 5. Fig. 5 shows a curve of input unbalance current as a function of temperature. This unbalance current determines the maximum value of total effective external resistance that may be used in each base circuit (resistors R_1 in Fig. 2). A maximum value of 10,000 ohms is recommended for each base circuit. However, larger resistances may be accommodated if the resistors can be adjusted to maintain low input offset voltages, or if the operating points of Q_1 and Q_5 are not in the linear region (as in trigger circuits).

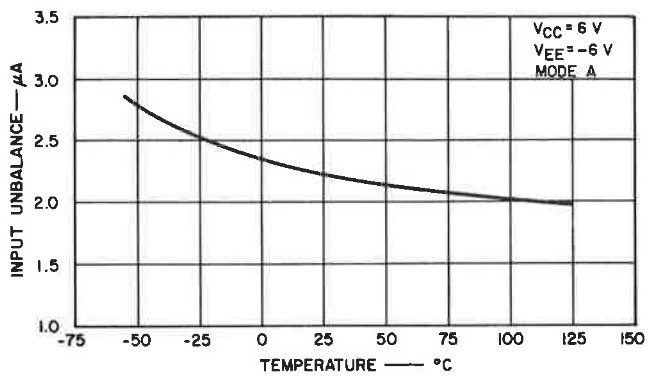


Fig. 5—Input unbalance current of the CA3002 as a function of temperature.

Input Impedance. The input impedance is essentially a characteristic of the input emitter-followers, Q_1 and Q_5. Because these transistors are lightly loaded, they have parallel input impedances that are approximately 0.1 megohm at low frequencies and rise to infinity and become negative at a few megacycles per second. In most cases, these impedances are negligible in comparison with the impedances of external base resistors or inductors. The input capacitance is 3 to 5 picofarads.

The input impedance decreases with decreasing operating temperature. A typical low-frequency value of parallel input resistance is 55,000 ohms at −55°C. If a resonant line or tuned circuit that has appreciable impedance in the vhf range is connected to either input terminal, a series parasitic resistor of 50 to 100 ohms should be placed in series with the input lead to prevent vhf oscillation.

Output Impedance. The output impedance is essentially that of the output emitter-follower Q_6, and is a function of the current in Q_6. The current, in turn, is determined by the operating mode, the supply voltages, and the connection of resistor R_6 to ground or to terminal 2. In operating mode D with R_6 returned to ground and ±6-volt supplies, the output resistance is approximately 80 ohms over most of the useful frequency range and rises to about 110 ohms (its highest value) at −55°C.

Frequency Response. The mid-frequency voltage gain of the CA3002 if amplifier is essentially independent of absolute resistor values, but depends on the resistor ratios. Fig. 6 shows a test circuit used to measure the response characteristics of an iterative-coupled amplifier that uses an input-coupling capacitor of 15 picofarads.

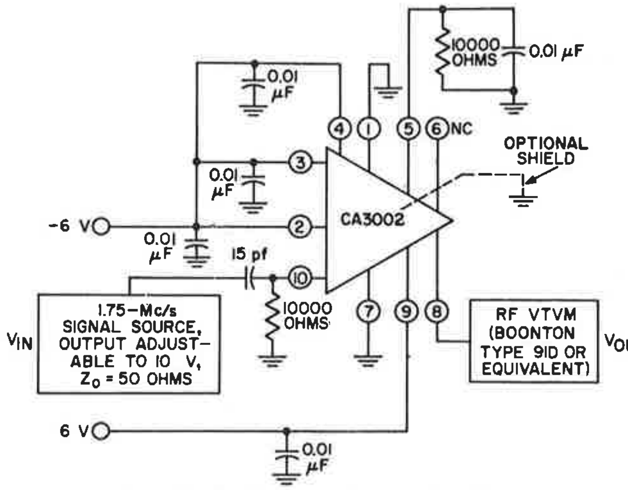


Fig. 6—Voltage-gain test circuit.

The response curves for several values of positive and negative supply voltage are shown in Fig. 7. The gain of the

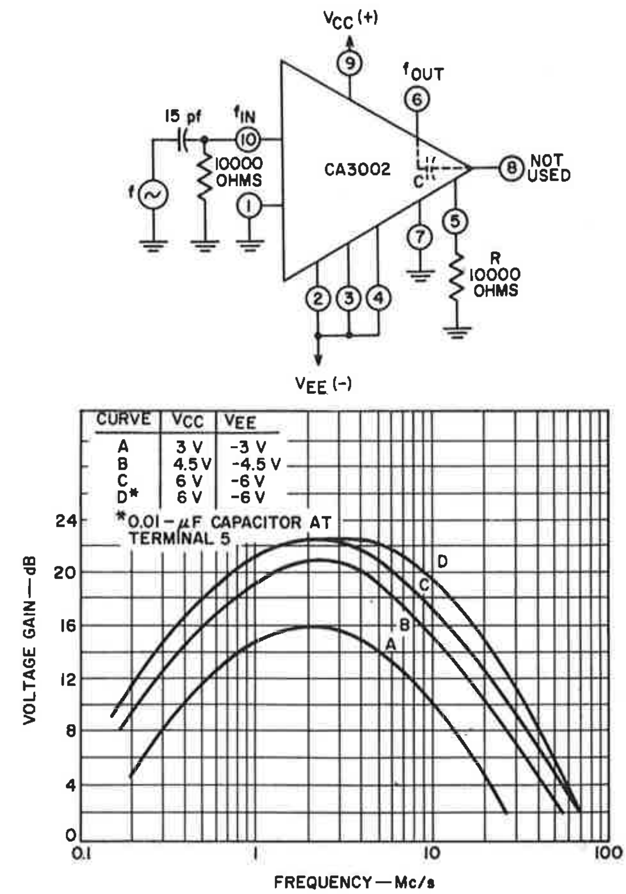
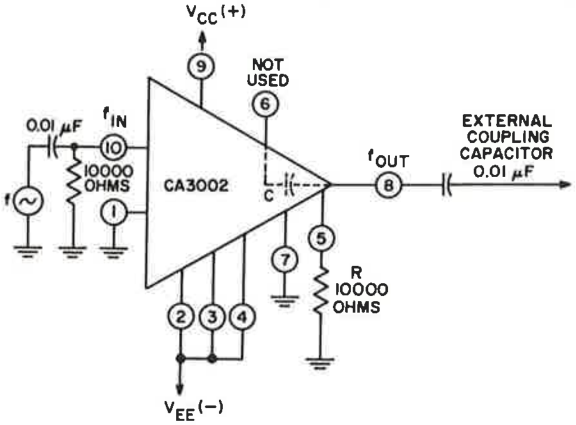


Fig. 7—Effective single-stage response characteristics of iterative-coupled if amplifier. Curve D represents operation with 0.01-microfarad capacitor connected at terminal 5 (not shown).

amplifier is reduced at low frequencies by the 15-picofarad input-coupling capacitor and at high frequencies by the RC roll-off within the circuit. The addition of a 0.01-microfarad bypass capacitor at terminal 5 improves both the high-frequency response and the mid-frequency gain by eliminating ac feedback from terminal 6 to terminal 5.

If a wideband video response is desired, the 15-picofarad internal silicon output-coupling capacitor of the CA3002 must be replaced with a larger external coupling capacitor connected to terminal 8. The response curves for an iterative-coupled amplifier that uses 0.01-microfarad input-coupling and output-coupling capacitors are shown in Fig. 8.



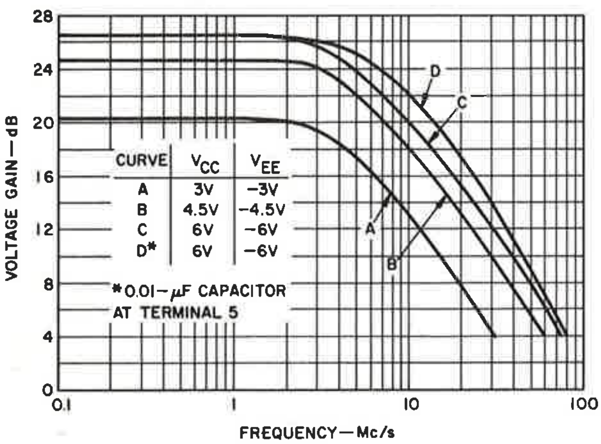


Fig. 8—Effective single-stage response characteristics for if amplifier using 0.01-microfarad coupling capacitors. Curve D represents operation with 0.01-microfarad capacitor connected at terminal 5 (not shown).

The response of the amplifier is substantially extended at the low frequencies. If 1-microfarad coupling capacitors are used, the low-frequency response can be extended below 100 cycles per second. Again, the addition of a 0.01-microfarad capacitor at terminal 5 improves the high-frequency performance. A shield separating the external leads at terminals 5 and 6 also reduces the feedback and extends the response.

AGC. The voltage gain of the CA3002 can be controlled over a wide range by adjustment of a negative dc voltage applied at terminal 1. Fig. 9 shows the voltage gain at 1.75 megacycles per second (measured in the test circuit of Fig. 6)

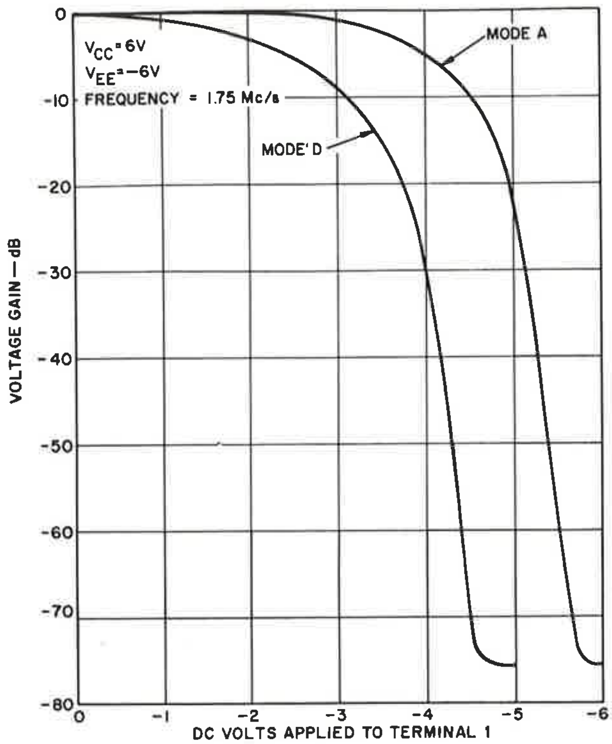


Fig. 9—Voltage gain of the CA3002 as a function of negative dc supply voltage applied at terminal 1 (normalized to a gain of 26 dB).

as a function of the dc voltage. When the gain is controlled in this manner, the CA3002 can be used as an if amplifier with a 75-dB agc range, or as a video gating, squelching, or blanking circuit with a similar range. The circuit function depends only on the manner in which the dc voltage applied to terminal 1 is controlled. The agc range is dependent on frequency, and decreases from 75 dB at 1 megacycle per second to 60 dB at 25 megacycles per second.

Third-Order Intermodulation Distortion. Fig. 10 shows the peak-to-peak input signal required to produce third-order intermodulation distortion of 3 per cent as a function of gain control for the CA3002 integrated circuit. The maximum tolerable signal input for 3-per-cent intermodulation dis-

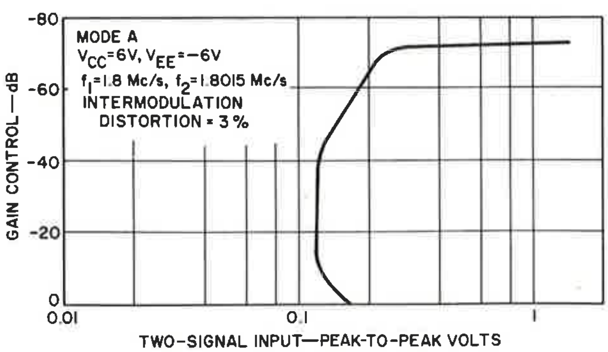


Fig. 10—Third-order intermodulation-distortion characteristic as a function of agc.

tortion is relatively constant over the entire agc range, but increases dramatically as cutoff is attained. When the CA3002 is operated in mode A with supplies of ±6 volts and an agc of −30 dB, a peak-to-peak input signal in excess of 100 millivolts is typically required for 3-per-cent distortion.

Noise Figure. Because noise figure is an important design parameter for both video and if-amplifier applications, it was evaluated for the CA3002 over the frequency range of 1 kilocycle per second to 10 megacycles per second. Fig. 11 shows noise performance as a function of frequency when a 1000-ohm source is used. The noise figure is 4 dB over a

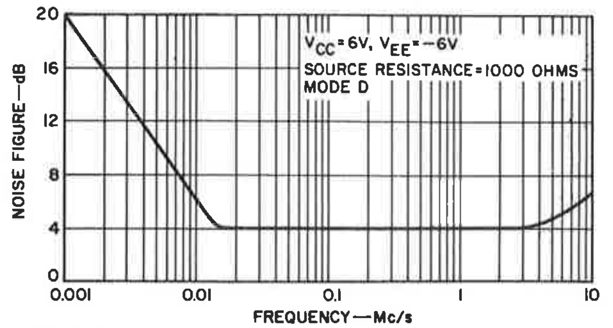


Fig. 11—Noise figure of the CA3002 as a function of frequency.

large portion of the usable range. The 1/f noise corner occurs at approximately 45 kilocycles per second, and the high-frequency noise rise begins at approximately 4 megacycles per second. Fig. 12 shows noise figure as a function of source resistance at 1.75 megacycles per second. The typical noise figure is less than 4 dB. It is reasonably flat for source resistances from 500 to 2500 ohms, but rises rapidly at values below 500 ohms.

When external base-bias resistors are used, terminal 5 should be bypassed by an external capacitor for any stage in which low noise figure is required. If the base-bias resistors are not bypassed, the noise figure increases. In a practical receiver, bypassing may be avoided if the input at terminal 10 is transformer driven (from a filter) and terminal 5 is grounded. In the later if stages, noise figure can usually be ignored.

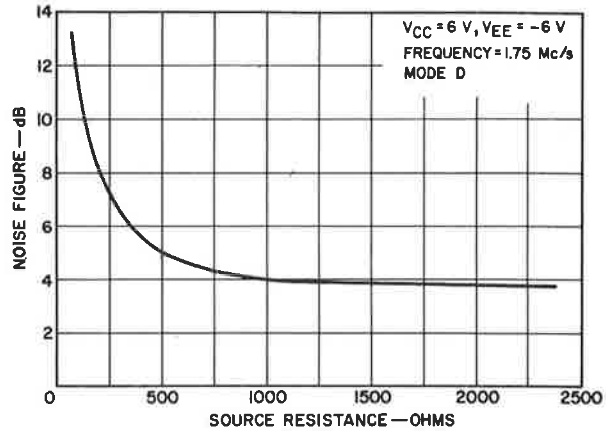


Fig. 12—Noise figure of the CA3002 as a function of source resistance.

APPLICATIONS

Four-Stage 1.75-Megacycle-per-Second IF Amplifier. Effective if design for AM circuits requires consideration of both the signal level at the input stage (as a function of agc range) and the acceptable signal-to-noise ratio at the output. The agc action must be initiated at the first stage at the proper voltage level to prevent excessive modulation distortion throughout the entire agc range. This input-signal voltage level is calculated to be approximately 25 millivolts rms for 100-per-cent modulation at an allowable distortion of 10 per cent. Before the applied signal reaches 25 millivolts, the first stage must be gain controlled and completely cut off before gain control is applied to subsequent stages.

Fig. 13 shows a four-stage 1.75-megacycle-per-second amplifier used to evaluate the performance of the CA3002

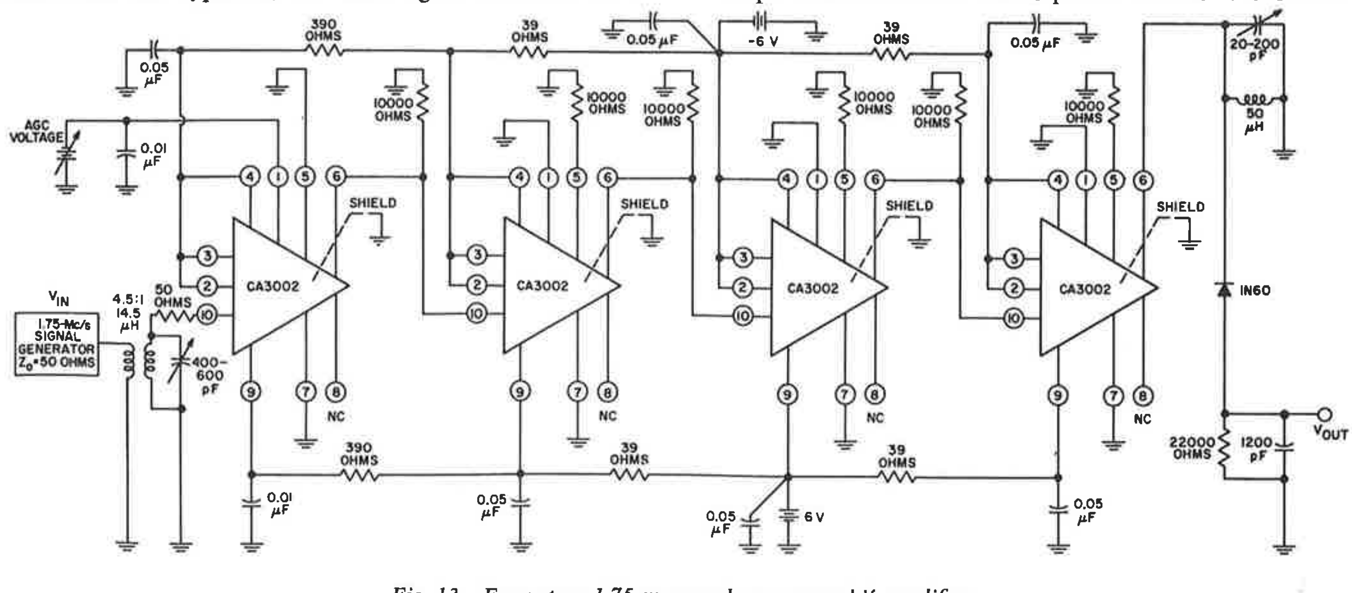


Fig. 13—Four-stage 1.75-megacycle-per-second if amplifier.

amplifier for AM applications. A tuned circuit and a diode detector are connected to terminal 6 of the output stage to evaluate detected output and signal-to-noise ratio. The audio bandwidth of the detector output filter is 3 dB down at 4.2 kilocycles per second. The tuned circuit at the input is driven by a 50-ohm generator and provides a 1000-ohm source to the circuit. The first stage is operated at reduced supply voltages (about ±3 volts) to reduce the required agc control voltage. This lower supply-voltage level ensures that a sufficient control voltage can be developed by a separate CA3002 unit used as an agc amplifier without introducing a separate supply voltage. An additional advantage of lower-voltage operation in the first stage is a reduction in noise figure.

If desired, the first-stage tuned circuit in Fig. 13 can be replaced by a crystal filter and a transformer. Because the CA3002 input impedance is high and does not vary appreciably with agc, no impedance variations are presented to the crystal filter.

The voltage gain realized from terminal 10 of the first stage to terminal 6 of the fourth stage is 85 dB, or approximately 21 dB per stage. From the 50-ohm input, the typical voltage gain is 98 dB. Because the maximum signal-handling capability of the output stage is slightly greater than 0.7 volt rms, gain control should begin when this value is measured at terminal 6. The modulation distortion is acceptable over the entire 60-dB agc range. For input signals greater than 8 millivolts, modulation distortion begins to increase because of fourth-stage overload. Overloading can be prevented by application of a delayed gain control to the second stage. The signal-to-noise performance as a function of input signal is shown in Fig. 14.

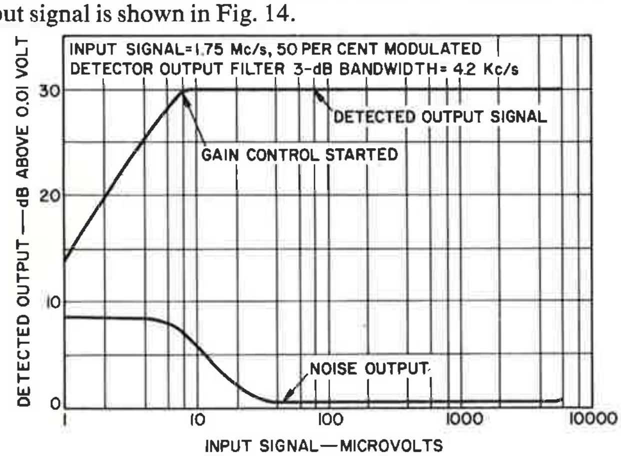


Fig. 14—Output signal and noise of the amplifier shown in Fig. 13.

Envelope Detector. The CA3002 integrated circuit can be operated as an envelope detector in either of two ways, as shown in Fig. 15: (1) the emitter of the output transistor Q_6 can be operated at zero voltage by connection of an external resistor in the bias loop of the constant-current transistor Q_3, or (2) the current in transistor Q_6 can be reduced by connection of a large resistor (12,000 to 18,000 ohms) in series with its emitter resistor.

In the circuit for method 1, the current in the differential pair (Q_2 and Q_4 in Fig. 1) is increased to the point at which the common-collector output transistor Q_6 is biased almost to cutoff. For this current increase, the constant-current transistor Q_3 is operated with terminal 4 open, and the emitter resistor R_9 is shunt loaded by the external resistor at terminal 3. Envelope detection can be accomplished only in mode A with method 1.

Although the output transistor is nearly cut off, all the other active devices are operating in their linear regions.

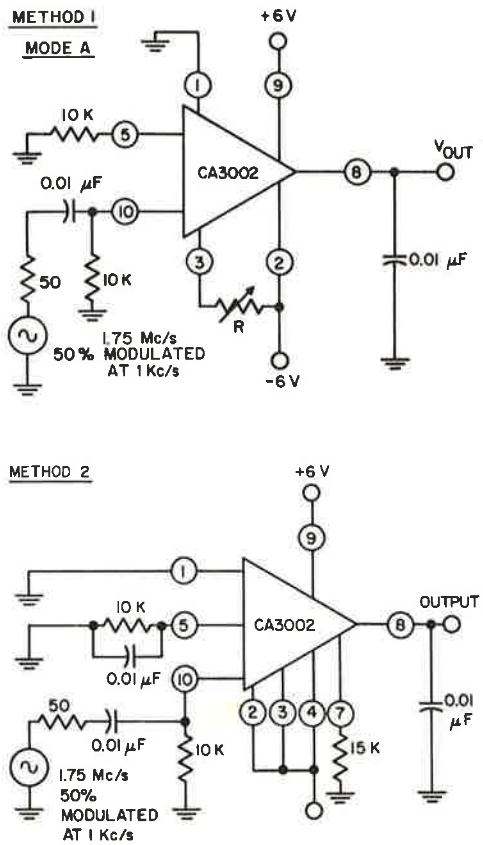


Fig. 15—Envelope detectors using CA3002 integrated circuits.

For small ac signals, therefore, the circuit provides linear operation except for Q_6, which is turned on only by a positive signal. The maximum acceptable input signal depends on the linear range of the differential amplifier. An external filter capacitor is connected between terminal 8 and ground to remove the rf signal from the detected audio output.

In the circuit diagram for method 2 shown in Fig. 15, a fixed value of resistance (15,000 ohms) is used to reduce the emitter current in the output transistor (Q_6) to approximately 100 microamperes. This operating point provides the nonlinearity for detection in transistor Q_6. Again, the remainder of the circuit produces gain because it is operating linearly. As in the case of method 1, an external filter capacitor is connected between terminal 8 and ground to remove the rf signal from the detected audio output.

Fig. 16 shows the input-output characteristics of the envelope-detector circuits shown in Fig. 15. The usable range of input signals for distortion below 3 per cent is 10 to 100 millivolts (20-dB range) for method 1 and 12 to 60 millivolts (14-dB range) for method 2. Automatic gain control of the if amplifier must maintain the input signals to the detector within this range.

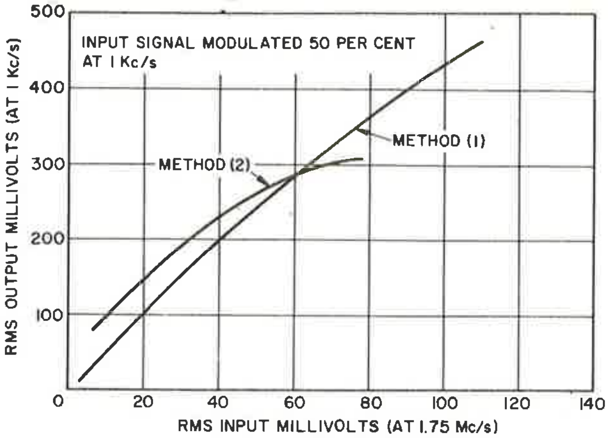


Fig. 16—Input-output characteristics of the envelope detectors shown in Fig. 15.

Product Detector. A differential pair driven by a constant-current transistor can be used as a product detector if a suppressed-carrier signal is applied to the differential pair and the regenerated carrier is applied to the constant-current transistor. There are two requirements for linearity: (1) the circuit must be operated in a linear region, and (2) the current from the constant-current transistor must be linear with respect to the reinserted carrier voltage.

The CA3002 satisfies these requirements and can be used as a product detector in the circuit shown in Fig. 17. A double-sideband suppressed-carrier signal is applied at terminal 10, and the 1.7-megacycle-per-second carrier is applied to terminal 1. Because of the single-ended output, a high-frequency bypass capacitor (0.01 microfarad) is connected between terminal 8 and ground to provide filtering for the high-frequency components of the oscillator signal at the output.

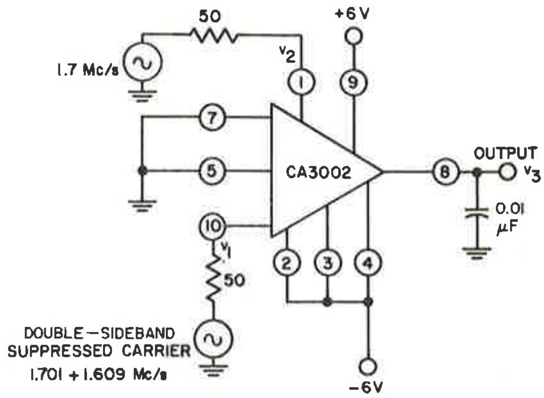


Fig. 17—Product detector circuit.

When the amplitude of the suppressed-carrier signal and of the oscillator signal are varied, the gain and distortion characteristics shown in Table III are obtained. The conversion voltage gain is constant at input signals up to 16 millivolts and would be 6 dB less for a single-sideband signal than for the double-sideband signal. The distortion increases with increasing input signal; for distortion of less than 1 per cent, the input drive level does not exceed 8 millivolts. The gain maximizes for oscillator voltages of 1 to 2 volts, and the distortion characteristic is also best in this region. Distortion increases both at low oscillator drive levels (0.25 volt) and at high levels (3 volts).

Schmitt Trigger. Fig. 18 shows the use of the CA3002 as a Schmitt trigger. In this application, the input is applied to terminal 5 and both the output and the feedback are taken from the output emitter-follower at terminal 8. The emitter-follower output isolates the feedback loop from the differ-

Table III—Performance Data for CA3002 as Product Detector

V_1 Double-Sideband Voltage (mV)	V_2 Oscillator Voltage at Terminal 1 (V)	V_3 Output at Terminal 8 at 1 kc/s (mV)	Conversion Voltage Gain (dB)	dB down from Fundamental of Harmonics *			
				Harmonic			
				2nd	3rd	4th	5th
1	1.7	12.5	21.9	60			
4	1.7	50	21.9	51	61		
8	1.7	100	21.9	46	56		
16	1.7	200	21.9	37	46		
32	1.7	310	19.8	32	30	51	64
4	0.25	22	15.6	15	42	44	
4	0.5	42	20.3	32	52		
4	1.0	60	23.5	45	60		
4	1.3	60	23.5	49	61		
4	1.7	50	21.9	51	61		
4	2.0	48	21.6	52	62		
4	2.5	31	17.8	49	60		
4	3.0	15	11.4	42	60		

*Harmonic Distortion Greater than 65 dB Down If Omitted

AWV