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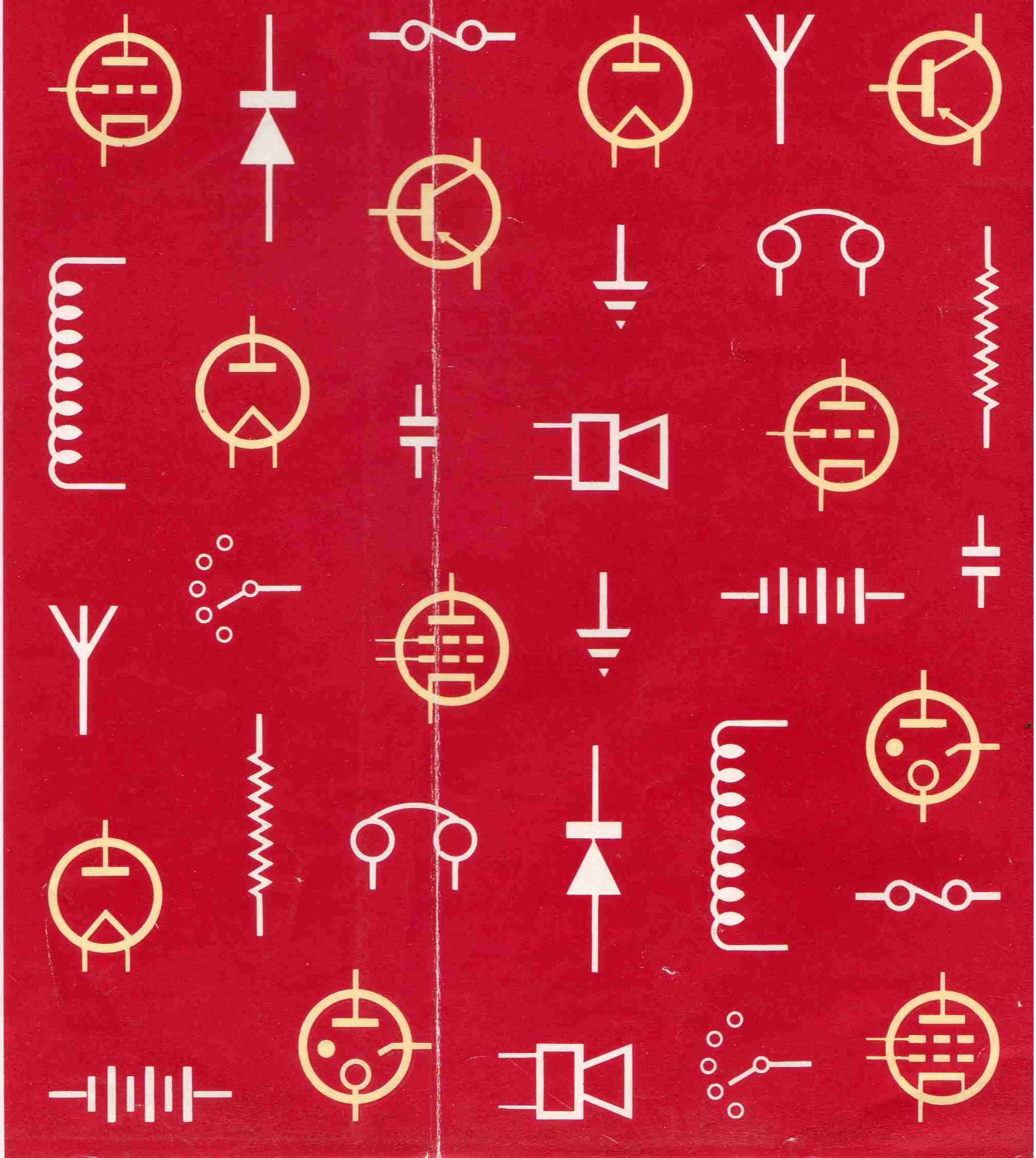

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IN THIS ISSUE

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A FIVE-TRANSISTOR AUTOMOBILE RECEIVER EMPLOYING DRIFT TRANSISTORS

by

Richard A. Santilli and C. Frank Wheatley

Introduction

This paper describes a five-transistor automobile receiver designed to use newly developed drift transistors. These drift transistors, which inherently have a high maximum available gain and low feedback capacitance, provide good performance with a minimum number of stages, thus contributing to low over-all circuit cost.

The lineup for the receiver is as follows: a 2N640 rf amplifier, a 2N642 converter, a 2N641 if amplifier, a 2N591 driver, and a 2N301 single-ended audio output stage. The receiver has a sensitivity of two microvolts for one watt of audio output. The audio circuit is capable of delivering an audio output of four watts at less than 10-percent distortion, and a maximum output of seven watts. This paper describes the circuit design considerations and the device capabilities.

Antenna Circuitry

The whip antenna used for automobile radios may be represented as a voltage generator having a capacitive internal impedance. Although there is also a resistive component of impedance present, the value is so low that it may be safely neglected when a 6-to-8 foot whip antenna is used over the broadcast band. Analysis shows that when the loaded Q is fixed for bandwidth purposes the maximum power transfer is obtained

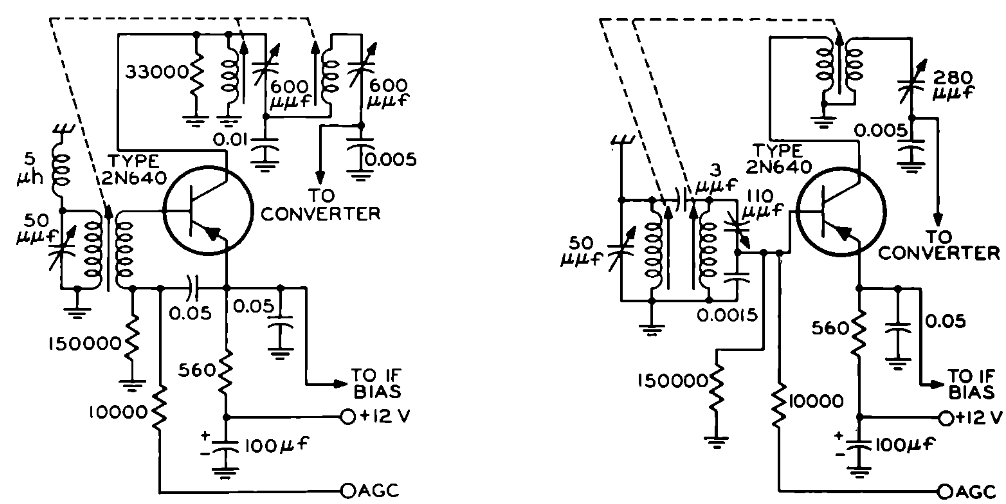


Fig. 1 — RF Amplifier Stage Using A 2N640 Drift Transistor. Stage (a) at left has single-tuned antenna circuit; Stage (b) at right has double-tuned antenna circuit.

when (a) the unloaded Q is as high as possibie, (b) trimmer and other shunt capacitors are kept to a minimum and (c) padders and other series capacitors are kept to a maximum. In addition, the power transferred decreases with decreasing frequency. As will be shown later, these factors must be taken into consideration to obtain a good signal-to-noise ratio.

The next problem was one of coupling the input impedance of the rf transistor to obtain the proper antenna loading. Although capacitive division eliminates the need for an antenna-coil secondary winding, this type of coupling requires the coil to be located between the antenna and the transistor base. This connection results in poor rejection of the high-field-strength, power-line interference occasionally encountered in auto radio applications. If the low end of the coil is grounded, as shown in Fig. 1 (a), selectivity "falloff" of 18 db/octave may be obtained. For reasons of power-line rejection and of signal-to-noise ratio, an antenna coil having a secondary winding was chosen.

Three rf tuned circuits were required to obtain an image and if rejection in excess of 70 db across the band. The antenna circuit could be single-tuned, as shown in Fig. 1 (a), or double-tuned, as shown in Fig. 1 (b). Although a double-tuned antenna has an obvious disadvantage as regards signal-to-noise ratio, the power-line rejection and the crosstalk performance make it very attractive. If a double-tuned antenna circuit were employed, top-side coupling using capacitive division for impedance matching would be the most advantageous. Overcoupling of about 25 or 30 per cent can be used without a sweep generator being required for alignment. Overcoupling reduces the antenna losses, and thus results in a signal-to-noise ratio which approaches that of the single-tuned antenna circuit.

RF Stage

RF gain of the order of 15 db is required to over ride converter noise adequately. Although neutralization may be used in a variable-frequency tuned amplifier, it was not considered necessary in this application.

Perhaps the most significant requirement for the rf transistor is the agc requirement. The control of rf input and output impedances and feedback capacitance is an obvious requirement. However, the range of signals encountered in an auto radio covers approximately 110 to 120 db. Consequently, the rf transistor must provide 80 to 90 db of agc. As a result, the dc beta (common-emitter current gain) and the I_{co} (cutoff collector current) must be controlled so that the agc system may supply sufficient power to utilize this cutoff range without amplification. This requirement becomes even more significant at higher ambient temperatures. The 2N640 drift transistor is designed specifically to meet these stringent requirements.

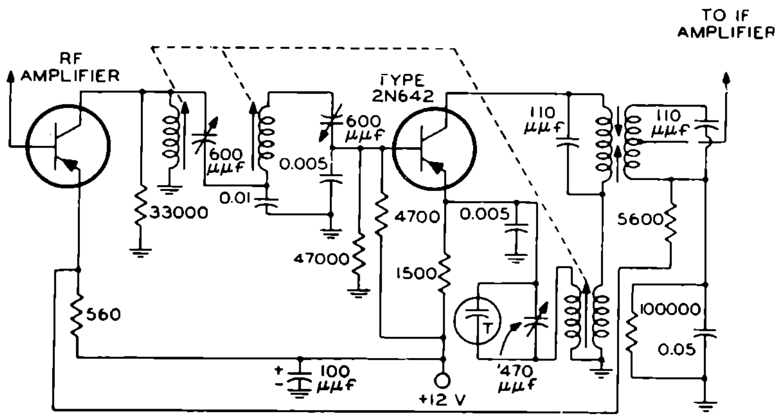


Fig. 2 — Converter Stage Using A 2N642 Drift Transistor.

When a double-tuned rf circuit is used, either top-side or bottom-side coupling may be used, together with capacitive division on the secondary winding for impedance matching to the base of the converter transistor. When a single-tuned rf circuit is used, capacitive division to the converter is also employed. The first rf coil was designed to be tuned with a 600-micromicrofarad capacitor to avoid a second winding or a tap. (It may be desirable to add a winding to the first coil illustrated in Fig. 1 (b) and tune with a smaller trimmer.) The value of 600 micromicrofarads was chosen to provide the correct collector loading as determined by dynamic stability considerations.

At maximum sensitivity, the rf stage is operated at a collector voltage of —12 volts and a collector current of 0.7 millampere, and produces a power gain of 27 db at the low-frequency end and 20 db at the high-frequency end of the band. This gain is sufficiently below the 2N640 maximum capabilities to assure excellent interchangeability and stability. At the low-frequency end of the band, the unloaded Q's of the first, second, and third tuned circuits are 65, 45, and 45, respectively, and the loaded Q's are 40, 30, and 30. At the high-frequency end the unloaded Q's are 65, 65, and 65 and the loaded Q's 48, 40, and 55. The coefficient of coupling of the double-tuned circuit, which is 1.3 times the critical coefficient provides a peak-to-valley ratio of approximately

0.2 db. This overcoupling is not sufficient to present alignment difficulties.

Converter Stage

The converter circuit shown in Fig. 2 is basically the autodyne type, in which emitter injection is obtained by capacitive division. This circuit uses a 2N642, also specially developed for this application. The rf signal is fed into the base, as mentioned previously. The if output from the collector is fed through a double-tuned transformer to the base of the transistor. The converter transistor operates at a collector voltage of —12 volts and a collector current of 0.6 milliampere, and produces a conversion gain of 37 db (262-kilocycle if). Again, this gain is below the 2N642 maximum capabilities and assures excellent interchangeability and stability.

The converter stage exhibits a strong tendency toward increased injection voltage with decreasing frequency. Conversion gain increases with injection up to a maximum point, and then starts decreasing with further increase of injection voltage. For the circuit shown in Fig. 2, the conversion gain is approximately independent of injection voltage between 40 and 150 millivolts and decreases with injection voltage outside this range. Because the sensitivity is greater at the low end of the band than at the high end, the injection voltage was controlled with frequency to decrease the variation of receiver gain across the band. The emitter is not as well bypassed for low-frequency rf signals as for high-frequency rf signals and thus provides additional flattening of the gain slope.

If the "tickler" winding is between the collector of the converter transistor and the if winding, the capacitance of the tickler winding to ground shunts the if primary and changes the coefficient of coupling. When this arrangement is used, this shunting capacitance must be considered in the design of the if transformer.

A capacitor having a negative temperature coefficient is used to shunt the converter tuning capacitor to provide oscillator stability with temperature.

The major problem in the design of the converter stage was that of blocking. When a very-high-level rf signal is applied to the base of the converter transistor the stage operates as a clamping circuit, thereby reverse-biasing the transistor and preventing oscillation. Without oscillation there is no if output, and, therefore, no agc to reduce the high-level rf input. If the incoming signal increases gradually, the agc has a chance to build up. However, it may come on abruptly if push-button tuning is used or if the radio is turned on in the presence of a strong signal. If the turn-on condition were the only one of concern, a suitable bypass capacitor could be incorporated so that the rf stage would gradually obtain bias (i.e., a "turn-on transition time" could be built into the circuit). However, this arrangement is no solution for push-button blocking, or blocking caused by tuning to a strong station.

The best method to handle blocking is to determine empirically the rf collector-signal level at which blocking occurs and then limit the collector signal below this level. The dc collector-to-emitter voltage can be chosen so that collector limiting will result. The signal level can further be reduced approximately 6 db by the use of a slightly-back-biased diode which shunts the collector load. A much greater degree of limiting may be obtained by the use of two diodes appropriately biased in a conventional limiting arrangement. Blocking is more severe when a single-tuned rf load is used. For both single-tuned and double-tuned rf loading, additional attenuation may be employed from the collector of the rf transistor to the base of the converter transistor to eliminate blocking, provided sufficient gain can be obtained elsewhere.

Blocking is less severe when the converter current or injection voltage is increased.

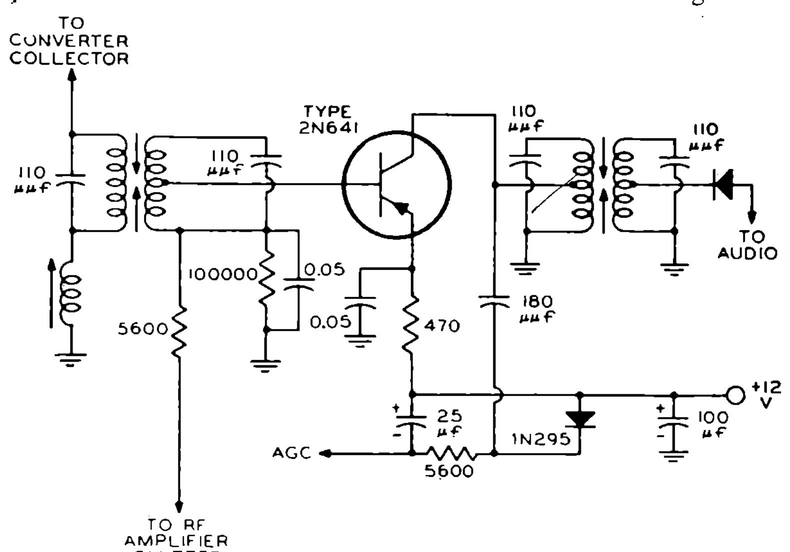


Fig. 3 — Unneutralized 262-Kilocycle IF Amplifier Stage Using A 2N641 Drift Transistor.

IF Stage

A 2N641 drift transistor is used in the unneutralized 262-kilocycle if amplifier, as shown in Fig. 3. Double-tuned input and output transformers are used to obtain the desired selectivity with coefficient of couplings set at 0.85 critical. Although more gain could be obtained with a higher collector load, and even more with neutralization, the if stage must deliver a high level of power.

Consequently, the collector load is determined by large-signal Class A power-amplifier criteria rather than dynamic stability alone. For the collector voltage of —12 volts and the collector current of 2 milliamperes, a collector load of 6000 ohms is used. The if stage contributes 32 db of gain.

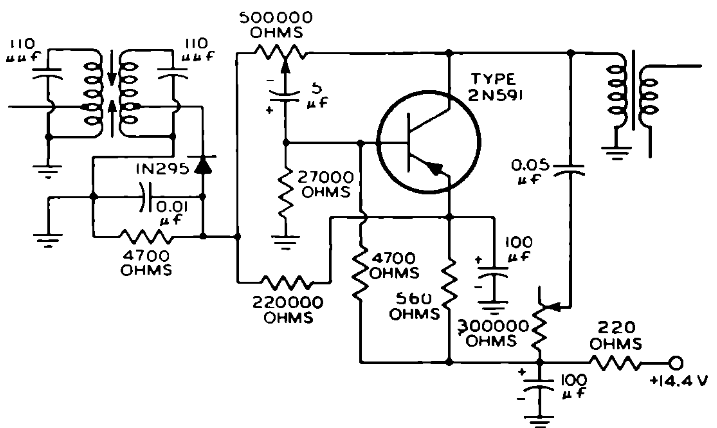


Fig. 4 — Audio Driver Stage Using A 2N591 Drift Transistor. Point-Contact Diode Used as Audio Detector is Shown at left.

The audio detector is fed from a tap on the secondary winding of the if output transformer. The agc detector is fed by a capacitor from the collector of the if transistor. This arrangement provides a slightly wider bandwidth for the agc than for the audio, and also permits a high level of agc voltage.

Approximately 5 or 6 db of agc is obtained from the if stage. This agc bias is obtained from the rf emitter.

Automatic Gain Control

The 110-to-120-db signal-handling requirement of this receiver makes agc a difficult problem. The germanium diode used for the agc detector develops approximately 2 volts of agc. A tendency toward distortion at very high levels was corrected by the use of a 2-micromicrofarad capacitor between the base and the collector of the rf transistor. This capacitor apparently extends the agc to some extent by introducing a feedthrough current which subtracts from the normal collector signal current. These currents are normally out of phase.

Another problem encountered in the agc system was that of spurious responses at very high levels. The agc bandwidth is fixed by the if bandwidth and is relatively narrow. When a strong signal is present and the receiver is tuned on carrier, the performance is as expected. As the receiver is tuned off carrier, however, the agc is rapidly removed. This change permits much higher levels in all stages prior to the output of the if stage, and shows up first as distortion of the envelope (and detected audio distortion). Further tuning off carrier removes the agc almost completely and permits a very high rf signal on the converter. In fact, if rf limiting is not employed, oscillator blocking may result. The wide agc system described previously reduces this effect.

It would also be possible to obtain freedom from blocking if additional agc voltage derived from the collector of the rf transistor were added to the normal agc bias. This arrangement would not alter the receiver performance appreciably in normal reception areas. If the receiver were in the presence of an extremely strong carrier, however, a quiet zone would be encountered in the vicinity of the carrier frequency. This arrangement was not incorporated in the receiver.

Audio Detection

A point-contact germanium diode is used as the audio detector, as shown in Fig. 4. For detection of small signals, maximum sensitivity is obtained by passing a certain value of forward dc current through the diode. As the current is reduced from this value, a slight loss in sensitivity is encountered. For signals 5 or 10 db lower, however, a considerable loss of sensitivity is observed. A high degree of quieting in the absence of a signal can be obtained with only a slight reduction in normal sensitivity by utilization of this nonlinearity of the detector. This arrangement does not actually improve the signal-to-noise ratio at maximum sensitivity, but it results in a subjective advantage similar to noise limiting. The compromise bias employed in this receiver causes a sharp curvature in the agc curve at sensitivity level. Admittedly, this curvature introduces detector distortion, as will be discussed later.

Maximum sensitivity is obtained when the input impedance of the audio transistor represents most of the load on the detector (i.e., when the

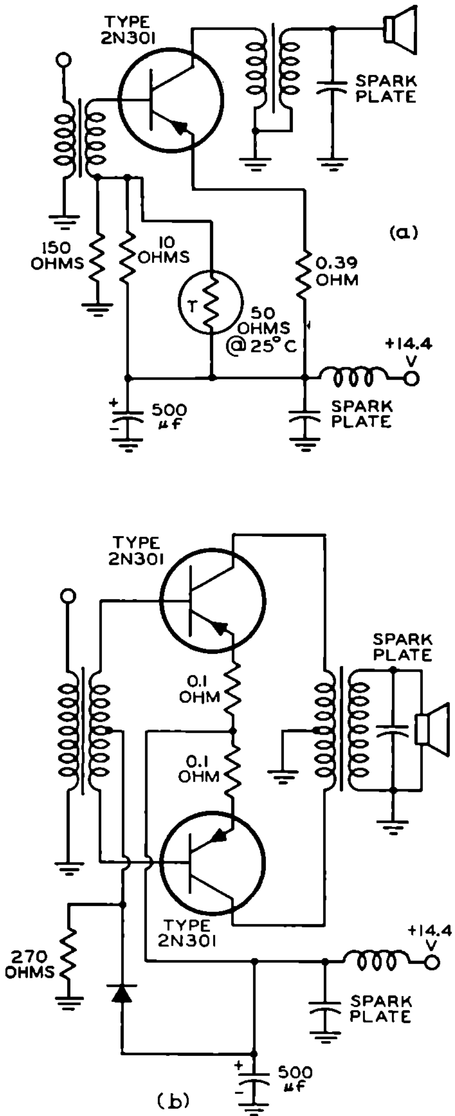


Fig. 5 (a) — Audio Output Stage Using a Single 2N301 Germanium Power Transistor to Provide a Power Output of 4 Watts. (b) Class B Push-Pull Output Stage Using Two 2N301 Transistors to Provide a 10-Watt Output.

dc load is small compared to the ac load). Again, a high degree of distortion is introduced in this manner, particularly at high modulation levels.

For operation at maximum sensitivity, however, the noise is sufficiently high to override the distortion. A signal-to-noise ratio of 20 db contributes as much undesired power as 10 per cent distortion. At signal-to-noise ratios below 20 to 25 db, therefore, distortion means little.

When the signal level is high enough to obtain a signal-to-noise ratio of 20 to 25 db (about 5 to 8 microvolts), the detector is fairly linear. In addition, either the output is clipped rather badly, or the volume control is turned down. The first 10 to 20 db of volume reduction inserts series resistance between the detector and transistor, thereby unloading the detector to approach an ac-to-dc ratio of unity.

Audio Driver

The 2N591 germanium-alloy audio driver transistor was specially designed for the high-temperature and high-voltage requirements of this application. The driver operates at a collector voltage of 12 volts and a collector current of 3 milliamperes and provides a power gain of approximately 44 db. Although the driver may contribute 3 or 4 per cent distortion at very low rf-signal levels, where the noise level is high, the volume control greatly reduces this distortion at normal rf-signal levels.

Volume Control

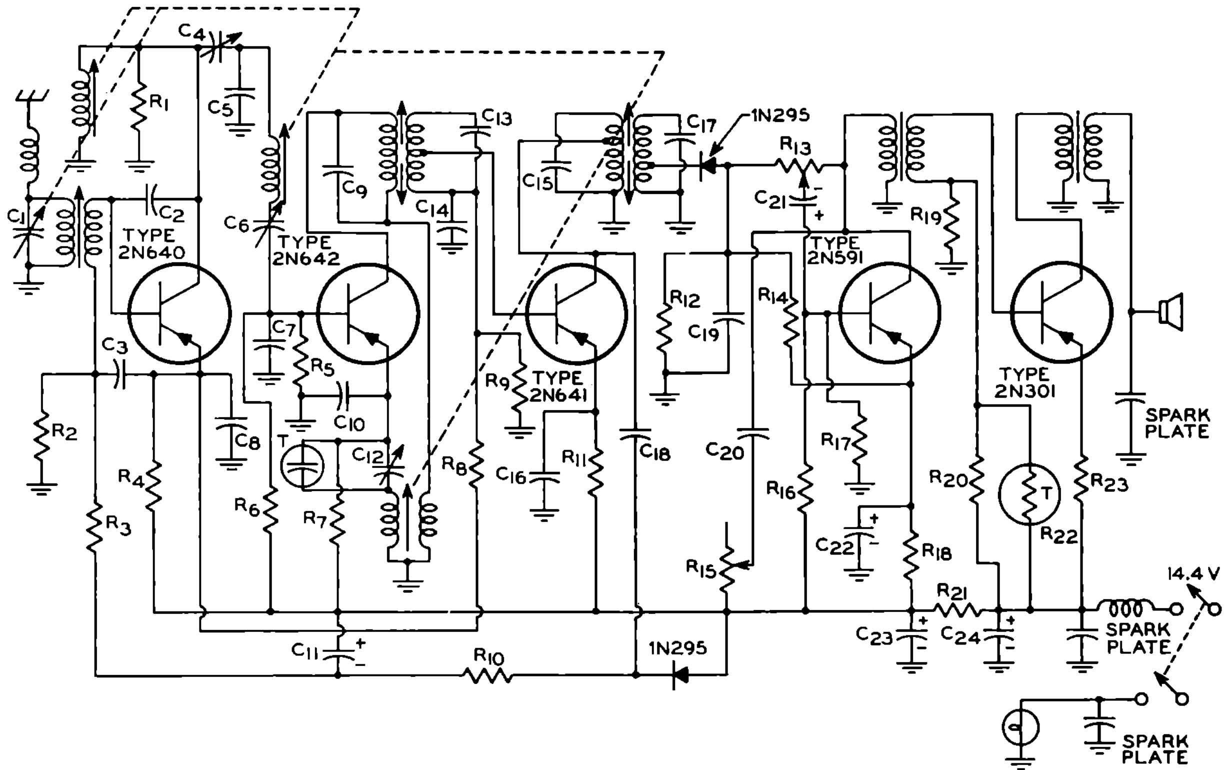
The volume control is a high-resistance potentiometer connected between the audio detector and the collector of the driver transistor. When the variable arm (connected to the base of the driver transistor) is placed at the detector end of the potentiometer, maximum sensitivity is obtained. The resistance of the potentiometer is high enough so that appreciable collector-to-base feedback is not encountered. As the arm is moved from the detector closer to the transistor collector, series attenuation results, thereby unloading the audio detector. Further reduction of volume causes collector-to-base feedback, which not only reduces the driver distortion (if significant), but also lowers the output impedance of the driver. When the output stage is driven by a high-impedance source, the presence of a unbypassed emitter resistor is not degenerative. Consequently, there is no loss of sensitivity (although there is a greater dynamic range required). When the output impedance of the driver is lowered by the feedback of the volume control, however, an output-stage emitter resistor provides a significant amount of loop feedback. Further increase of the volume-control setting results in additional feedback and, finally, attenuation produced by loading of the driver collector. Volume adjustment of about 100 db can be obtained by this method. For best effect, the volume control should have an S-taper with about 5-per-cent resistance at 35-per-cent rotation and 95-per-cent resistance at 65-per-cent rotation.

Tone Control

A treble-cut tone-control circuit is incorporated between the collector of the driver transistor and ground. The capacitor in this circuit is chosen so that the 3-db-down frequency for full cut is about 400 cycles at high volume settings.

Audio Output

The audio output stage for the 4-watt receiver consists of a single 2N301 germanium power



C_1: 50 μμf
C_2: 2 μμf
C_3 C_8 C_{14} C_{16} C_{20}: 0.05 μf
C_4 C_6: 600 μμf
C_5 C_{19}: 0.01 μf
C_7 C_{10}: 0.005 μμf
C_9 C_{13} C_{15} C_{17}: 110 μμf
C_{11}: 25 μf, electrolytic
C_{12}: 470 μμf
C_{18}: 180 μμf
C_{21}: 5 μf, electrolytic
C_{22} C_{23}: 100 μf, electrolytic
C_{24}: 500 μf, electrolytic

R_1: 33000 ohms
R_2: 150,000 ohms
R_3: 10000 ohms
R_4 R_{18}: 560 ohms
R_5: 47000 ohms
R_6 R_{12} R_{16}: 4700 ohms
R_7: 1500 ohms
R_8 R_{10}: 5600 ohms
R_9: 100,000 ohms
R_{10}: 470 ohms
R_{13}: 500,000 ohms potentiometer
R_{14}: 220,000 ohms

R_{15}: 300,000 ohms, potentiometer
R_{17}: 27000 ohms
R_{19}: 150 ohms
R_{20}: 10 ohms
R_{21}: 220 ohms
R_{22}: Thermistor, 50 ohms at 25° C
R_{23}: 0.39 ohm

Fig. 6 — Schematic Diagram for Complete 5-Transistor, 4-Watt Receiver With Single-Tuned Antenna, Double-Tuned RF Load, and Class A Output.

transistor operated under Class A conditions, as shown in Fig. 5 (a). When this transistor is driven without regard to distortion, the power-output level is 7 watts. A pair of 2N301 transistors in Class B operation can be used to provide a 10-watt output, as shown in Fig. 5 (b). When this circuit is overdriven, the power-output level is 17 watts. In either circuit, the power transistor must be provided with an adequate heat sink to avoid exceeding a maximum junction temperature of 85 degrees centigrade.

In the 10-watt circuit, the bias for the audio output stage is established by a developmental "compensating" diode which maintains essentially a constant collector idling current despite temperature or supply-voltage changes.

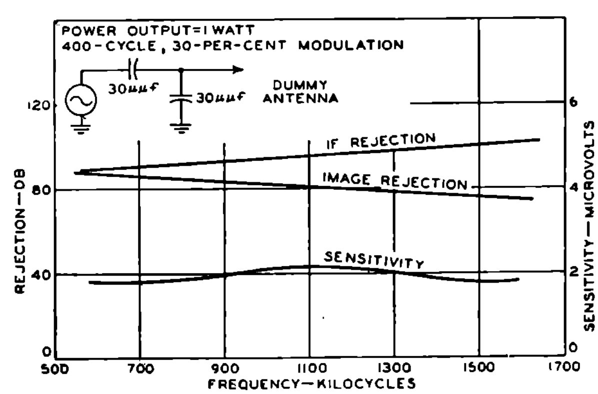


Fig. 7 — Sensitivity and Rejection Characteristics of 5-Transistor Receiver.

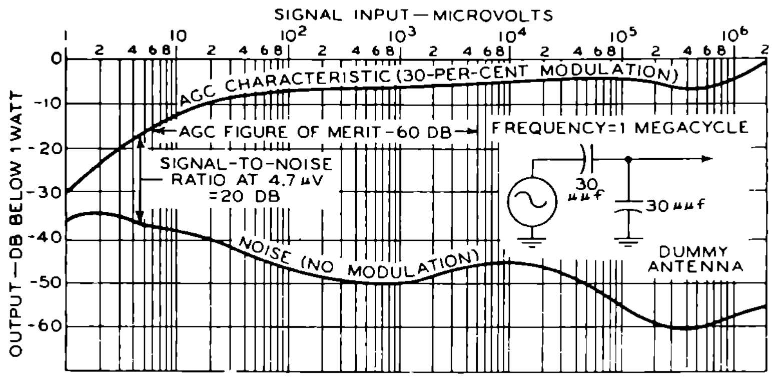


Fig. 8 — AGC and Noise Characteristics of 5-Transistor Receiver.

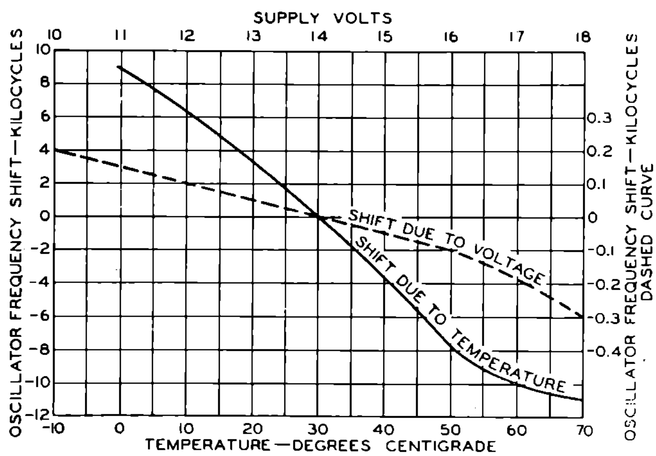


Fig. 9 — Frequency Shift of Uncompensated Oscillator With Frequency and Voltage.

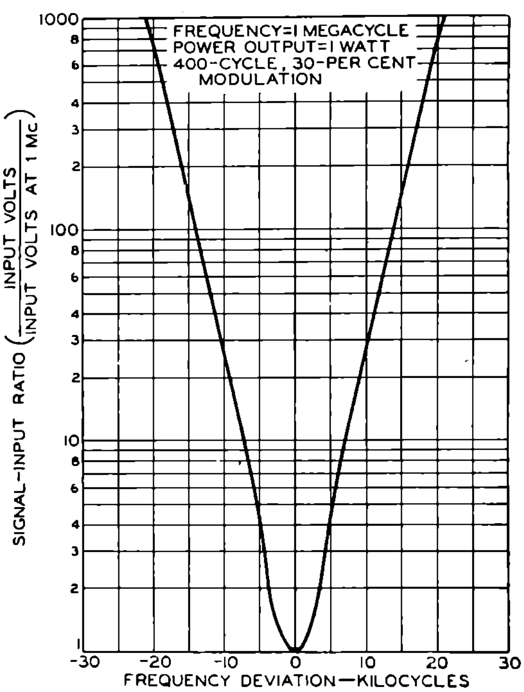


Fig. 10 — Over-All bandwidth of 5-Transistor Receiver at a Signal Frequency of 1 Megacycle.

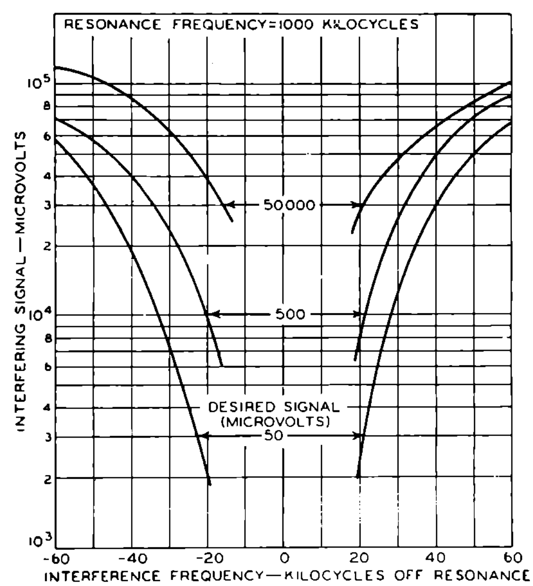


Fig. 11 — Crosstalk Characteristics of 5-Transistor Receiver for Signal Levels of 50, 500, and 50000 Microvolts.

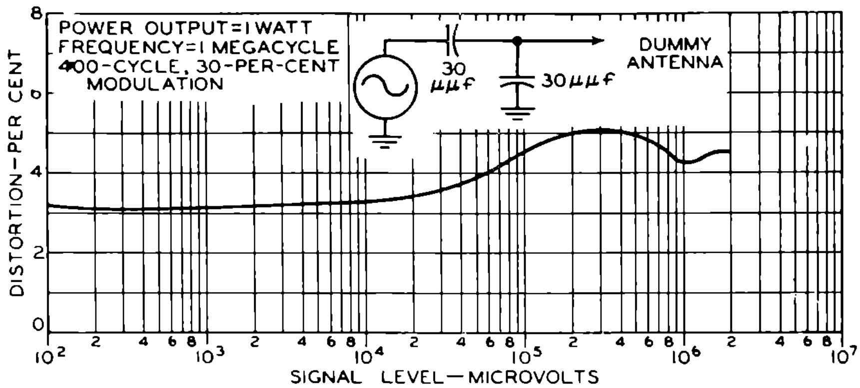


Fig. 12 — Distortion of the 5-Transistor Receiver as a Function of Signal Level.

Receiver Performance

Fig. 6 shows the circuit diagram for the complete five-transistor, 4-watt, Class A output receiver. (The characteristics given here apply to the single-tuned antenna and double-tuned rf load) Fig. 7 shows the tracked sensitivity, image rejection ratio, and if-rejection ratio as functions of frequency. With the dummy antenna shown in the insert, the sensitivity is 2 microvolts across the band. The image-rejection ratio varies from 85 db at low end of the band to 78 db at the high end. The if-rejection ratio varies from 89 db at the low end of the band to 102 db at the high end.

The agc and noise characteristics of the receiver are shown in Fig. 8. The receiver has a 60-db agc figure of merit (using a 5000-microvolt reference). A signal-to-noise ratio of 20 db occurs at less than 5 microvolts. No oscillator blocking was experienced with signal levels up to 2 volts.

Fig. 9 shows the frequency shift of the uncompensated oscillator with frequency and voltage. The frequency change is approximately 280 cycles per degree centigrade and 50 cycles per volt. Fig. 10 shows the over-all bandwidth of the receiver at a signal frequency of 1 megacycle. The 2x-down bandwidth is 7 kilocycles and the 1000x-down bandwidth is 40 kilocycles. The cross-talk characteristic in Fig. 11 is given for signal levels of 50, 500, and 50,000 microvolts. Fig. 12 shows the distortion as a function of signal level.

(With acknowledgements to RCA)

PICTURE TUBE DATA AND INTERCHANGEABILITY CHART

This chart lists no less than 36 picture tubes used in Australian TV receivers, with mechanical and electrical data, including base connections. Changes and adjustments required in making substitutions are detailed on the chart. Have you got your copy? It is free and post free, the address is on the contents page of this issue, and you should ask for "TV-2".

Off The Beaten Track

A SERIES DESCRIBING SOME OF THE MORE UNCOMMON VALVES AND VALVE DESIGNS

No. 8 — FLYING SPOT SCANNERS

When a "still" advertisement or announcement appears on TV, it will in general be a transparency which is viewed and transmitted by the station. The method of viewing, with which we are concerned in this article of the series, is a choice of two alternatives. In the one case the transparency may be illuminated by a light and ground glass screen behind it, and viewed just like any other scene or object by a vidicon. In the second case a flying spot scanning tube may be used to view the transparency.

The vidicon method is simple, and is becoming increasingly popular as compared with the flying spot method. Furthermore, the camera can be a normal one and may be used for other work as well. However, the flying spot scanner is an interesting tube, and well worth examining. Its use is not restricted to transmitting advertisements and announcements; it can also be used as part of a pattern generator set up for TV testing and alignment in place of a monoscope, as for example in a large factory, where the signal could be "piped" to test bays and service benches. A system using a flying-spot scanner has high resolution capability, and unlike the monoscope system, the pattern or picture can be changed at will.

The flying spot scanning tube is essentially a cathode ray tube, usually five inches in diameter. It differs from, say, an oscilloscope tube in that it uses a special phosphor and has a large light output with short persistence. The large light output means a high-energy beam, and in fact these tubes usually operate with an ultor voltage approaching 30 kilovolts. The short persistence, together with a phosphor having relatively little grain, produces a picture having high resolution.

A typical five-inch flying spot scanner tube is shown in Fig. 1. It is the 5ZP16, intended for monochrome work. This tube uses a P16 phosphor, which is particularly suitable for this application. The phosphor has a spectral energy characteristic with a peak in the near ultraviolet

Fig. 1.—A Typical Five-Inch Flying-Spot Scanning Tube, the 5ZP16.

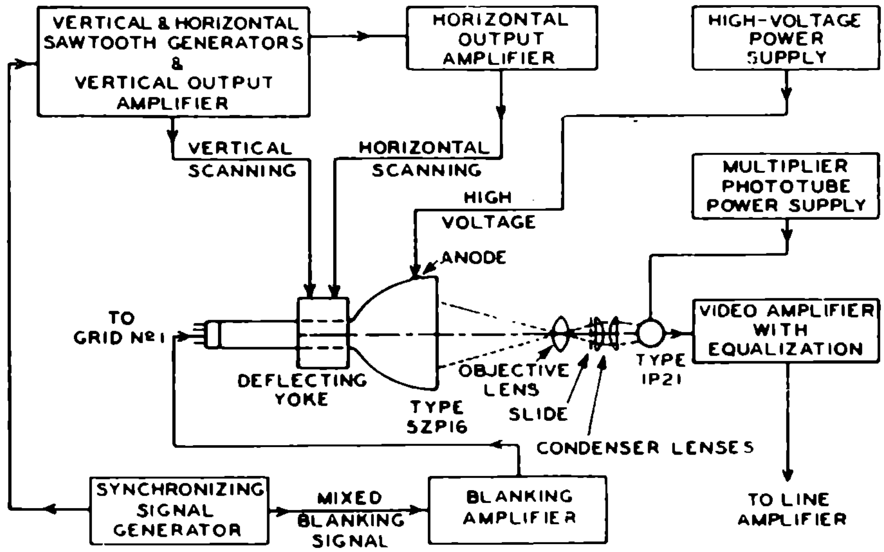


Fig. 2.—Block Diagram of a Flying-Spot Video-Signal Generator System for Slide Transparencies.

region. As previously mentioned, it has very short persistence, and has a stable, exponential decay characteristic. The tube face is made of a special, non-darkening glass, and has an optical quality and flatness which will not limit the performance of a high-quality objective lens needed to provide maximum signal resolution.

But let us turn for a moment to a typical flying-spot video-signal generator, and see how this tube works. A typical set-up is shown in Fig. 2.

A flying-spot video-signal generator consists essentially of (1) a flying-spot cathode-ray tube with associated power supplies, deflecting yoke, and scanning circuits to provide a small, rapidly moving source of radiant energy, (2) an optical system arranged to project the raster on the subject to be scanned, (3) the subject which may be a slide transparency, motion picture film, or an opaque object, (4) a multiplier phototube with associated power to convert the radiation transmitted or reflected by a subject into a video signal, and (5) an amplifier to increase the strength of the video signal.

A block diagram of such a system arranged for use with a slide transparency as the subject is shown in Fig. 2. For best results, the objective lens should be a high-quality enlarger type designed for low magnification and preferably corrected for use with ultraviolet radiation. The diameter of the objective lens should be adequate to cover the slide to be scanned.

Trailing results from the lag in buildup and decay of output from the screen. As the flying spot moves across a boundary from a light to a dark area of the subject being scanned, the persistence of energy output from the screen results in continued input to the phototube from the light area during the time the dark area is being scanned. Thus, the light area trails into the dark area in the reproduced picture. Similarly, as the flying spot moves from a dark area to a light area, the lag in buildup of the screen output causes the dark area to trail over into the light area. As a result of these effects, the reproduced picture has an appearance similar to that produced by a signal deficient in high frequencies. It is, therefore, necessary to enhance the high-frequency response of the video amplifier by introducing equalizing networks of the resistance-capacitance type with suitable time constants. Sufficient equalization should be provided to give the desired square-wave response.

The decay (persistence) characteristics of most standard phosphors are such as to require considerable equalization provided by networks with different time constants in several stages of the video amplifier. Their relatively long decay generally results in appreciable reduction of the useful signal-to-noise ratio. The persistence of the P16 phosphor, however, is extremely short to provide good signal-to-noise ratio. The P16 phosphor also has a stable, exponential decay characteristic for which equalization can be easily supplied. The small amount of equalization needed can be supplied by only one network. As a result, circuits and adjustments are simplified.

It will now be apparent that the flying-spot scanning tube is in fact a controlled light source used to illuminate the slide which is being transmitted. Compare this process with that used when a slide is viewed by a vidicon, as described recently in this series. The comparison shows a completely different approach to the problem, but remember that the vidicon can also view moving scenes.

The deflection equipment used with the tube is similar to that used in a normal TV set, with the addition of the synchronising signal generator and blanking amplifier. The blanking amplifier "blacks out" the tube during flyback, whilst the sync generator provides a signal which not only co-ordinates the other deflection circuits, but also provides sync for the equipment which is to transmit and display the pattern or picture. A typical deflection angle for this type of tube is 40°; this small angle minimises deflection defocusing and provides high corner resolution.

One problem in using these tubes is the very high ultor voltage. The tube is provided with an external conductive coating on the neck; this coating is grounded, and prevents corona between the deflection yoke and the neck. In addition, the bulb cone has an external moisture-repellant insulating coating to minimise sparking over the surface of the bulb under conditions of high humidity.

The high voltage applied to this type of tube increases the rate at which dust is precipitated on the surface of the tube. The rate of precipitation is further accelerated in the presence of corona. Such dust not only decreases the insulation of the bulb coating but also reduces the amount of radiation transmitted through the bulb face. The dust usually consists of fibrous materials and may contain soluble salts. The fibres absorb and retain moisture; the soluble salts provide electrical leakage paths that increase in conductivity as the humidity increases. A film of dust can nullify the protection provided by the insulating coating on the bulb, and the tube must be kept clean.

Any high-voltage system may be subject to corona, especially when the humidity is high, unless suitable precautions are taken. Corona, which is an electrical discharge appearing on the surface of a conductor when the voltage gradient exceeds the breakdown value of air, causes deterioration of organic insulating materials, induces arc-over at points and sharp edges, and forms ozone, a gas which is deleterious to many insulating materials. Sharp points or other irregularities on any part of the high-voltage system may increase the possibility of corona and should be avoided. Instead, rounded contours and surfaces must be used.

Another problem associated with the operation of such high-voltage tubes is X-ray radiation. X-ray radiation is produced at the face of the tube whilst it is operating, and although they are only "soft" rays, they constitute a danger to the health of operating personnel unless the tube is adequately shielded. The term "soft" in this context is used to describe X-rays of low penetrating power, or in other words, low-voltage X-rays. The radiation is produced in this instance by a voltage, which, whilst physically quite high, is still low in relation to the voltages used in deep-penetration radiological equipment. As pointed out, they can nevertheless be a health hazard.

It may be interesting at this point to look at a flying-spot scanner tube intended for colour work.

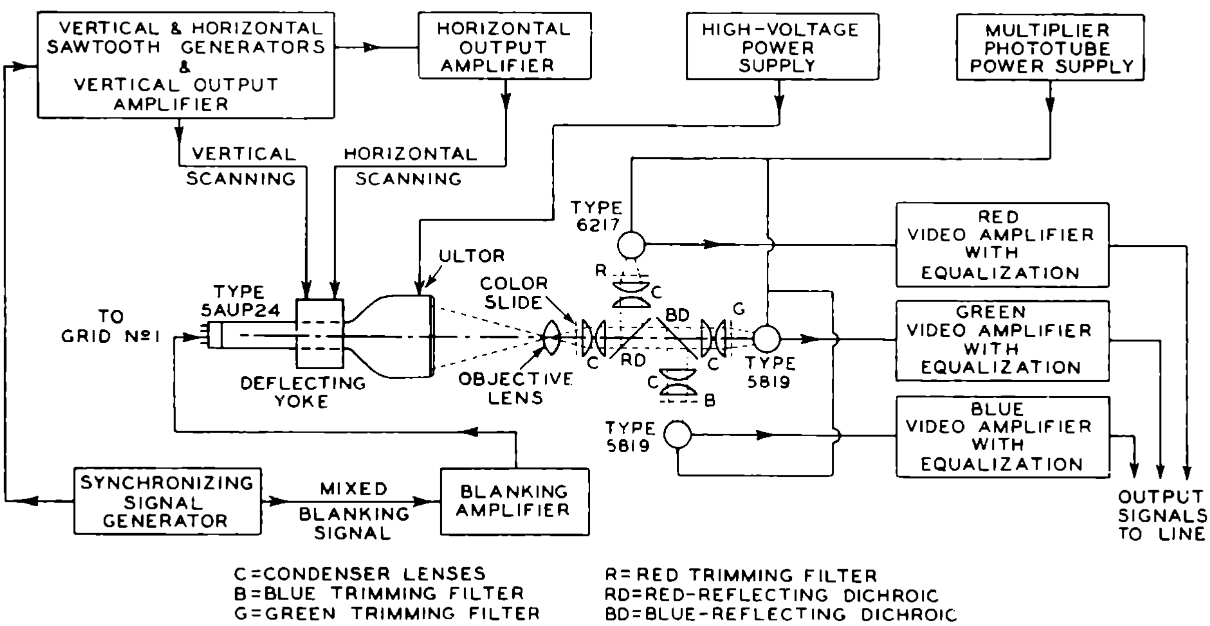


Fig. 3.—Block Diagram of a Colour Flying-Spot Video-Signal Generator System for Colour Slide Transparencies.

Essentially the tube will be the same as that just described, except that the phosphor used will probably be P24. This phosphor, like the P16, has very short persistence. It has a peak spectral response in the blue-green region, but has a sufficient light output over the visible part of the spectrum to render it suitable for colour work.

A typical colour flying-spot video-signal generator is shown in Fig. 3. The generator is similar to that previously described, except that the colours are separated into three primaries by red, green and blue filters, and each colour has its own multiplier photocell to produce an output signal for each colour. In order that each photocell and filter system may view the transparency, a system of two dichroic mirrors is used. These mirrors are perhaps best described as "half-silvered". They have the property of both passing and reflecting a beam of light, although both the direct and reflected rays undergo attenuation.

NEW RELEASES

1N2326

The 1N2326 is a new germanium alloy-junction diode with temperature-and voltage-compensating characteristics matched to transistor types 2N217, 2N270, 2N408 and similar types. In class B push-pull audio amplifiers, the 1N2326 can maintain a virtually constant quiescent operating point for the output stage over supply-voltage variations up to ± 40 per cent and ambient-temperature variations from -20°C to + 71°C.

When used in suitable circuits, the 1N2326 offers these two important advantages:

1. Substantial reduction in af distortion which would normally result from temperature variations.
2. Longer useful dry-cell life because standard dry cells may be used to substantially lower end-point voltages before af distortion becomes objectionable.

2N1224, 2N1225, 2N1226

In recognition and support of the trend toward industry standardization of transistor dimensional outlines, three germanium p-n-p drift transistors conforming to the TO-33 outlines have been introduced. These transistors have welded cases and are intended for use in military and commercial applications.

The 2N1224 is like the widely accepted 2N274 except for its dimensional outline. The 2N1224 is especially useful in oscillator applications up to 50 Mc, radio-frequency applications up to 20 Mc, and in if, mixer, converter and low-level video amplifier circuits.

The 2N1225 is like the widely accepted 2N384 except for its dimensional outline. The 2N1225 is especially useful as an oscillator at frequencies up to 125 Mc, rf amplifier at frequencies up to 50 Mc, and in pulse-amplifier, and high-speed non-saturating switching circuits.

The 2N1226 is like the 2N274 except for its dimensional outline and a maximum collector-to-base voltage rating of -60 volts. It is intended for those military and commercial applications requiring such a high voltage.

RADIOTRON 7198

The 7198 is a camera tube of the image-orthicon type designed primarily to provide reliable performance in television cameras for industrial or military service under adverse environmental conditions. It is capable of withstanding operating conditions involving severe shock and vibration, altitude up to 60,000 feet, wide-temperature range, and high humidity. Resolution capability of the 7198 is in excess of 600 TV lines.

The 7198 has high sensitivity combined with spectral response approaching that of the eye, and very good resolution capability. The response of the 7198 covers the range from about 3200 angstroms to 6950 angstroms. The target in the 7198 is designed to have low capacitance. Because of this feature, the 7198 has negligible microphonics due to movement of target and mesh with respect to each other and is capable of reproducing motion in low-light-intensity scenes with a minimum of smearing. When used with proper, low-noise amplifiers, the 7198 can produce signal information with illumination of the photocathode as low as 0.00001 footcandle.

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THE JUNCTION TRANSISTOR

By R. W. HURST

PART 3 — HIGH FREQUENCY OPERATION, PULSE OPERATION AND TEMPERATURE CONSIDERATIONS

Previous articles in this series have presented the three basic transistor configurations — the common-emitter, the common-base, and the common-collector — and have briefly described their characteristics. In all these descriptions, it was assumed that the signal frequencies were in the audio region, the signal amplitudes small, and the ambient temperature around 70°F. If any or all of these assumptions are not true, the transistor's behaviour cannot be predicted from the simplified descriptions of the first two articles. This part of the article will extend those descriptions to show transistor behaviour at high frequencies, at high temperatures, and in highly non-linear operation.

Present-day transistors are not capable of providing high-frequency operation and wide bandwidths with the same ease that valves can provide such performance. However, proper choice of transistor, transistor configuration and associated circuitry can result in very good video amplifiers and tuned high-frequency amplifiers, with performance equal to that of conventional valve circuits. The circuits and configurations necessary to obtain these results will now be considered.

High Frequency Operation

The high-frequency performance of a transistor circuit is strongly influenced by the type of transistor employed. A given circuit will provide widely different high-frequency responses with different transistors. For example, a simple current-driven common-emitter amplifier shown in Fig. 33 will provide a bandwidth of about 16 Kc for a 2N104, about 120 Kc for a 2N219, and about 1700 Kc for a 2N384. (These bandwidths ignore stray capacities shunting the load resistor.) Over most of the bandwidth shown, the transistors give current gains of beta — 45, 50 and 60, respectively — which fall off to gains of about 70 per cent of beta at the specified frequency. This frequency is known as the beta cutoff frequency, $f\beta$, and is an important parameter for describing a particular transistor's potentialities as a high-frequency device.

Much better high-frequency performance can be obtained from a transistor in the common-base configuration, as shown in Fig. 34. In this configuration, the 2N104's bandwidth is extended from 16 Kc to 700 Kc; the 2N219's bandwidth from 120 Kc to 6 Mc; and the 2N384's bandwidth from 1.7 Mc to 102 Mc.

Although these are impressive bandwidths, note that in each case, the current gain has dropped from beta (around 50) to alpha (less than 1). That is, going from common-emitter operation to common-base operation extends the bandwidth by a factor of beta, but at the same time reduces the current gain by the same factor: this is illustrated in Fig. 35.

In the common-base case, the frequency at which the current gain is down 70 per cent (6 Mc for the 2N219) is called the alpha cutoff frequency, $f\alpha$. It is the frequency most commonly given in transistor data sheets, and is

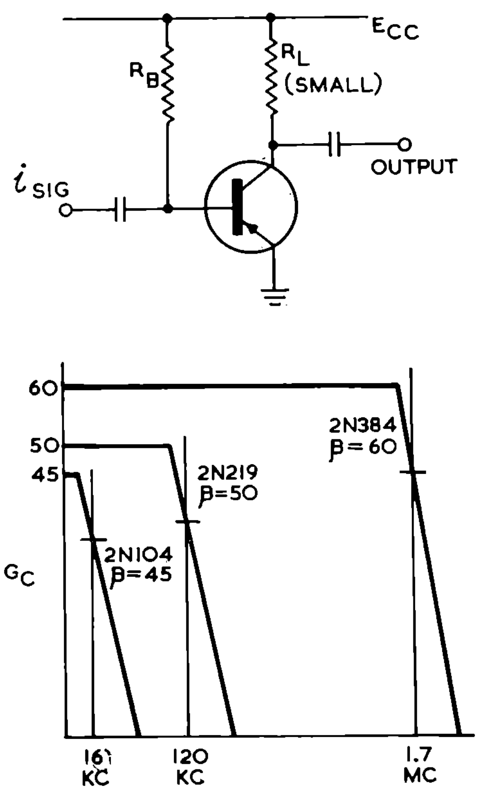


Fig. 33

related to fβ by the expression:

$$f_{\alpha} = \beta f_{\beta}$$

For maximum gain, it is necessary to operate a transistor in the common-emitter configuration; yet this configuration gives poor frequency response. There is a need, therefore, to arrive at some configuration or circuit arrangement which will yield better frequency response without sacrificing all of the common-emitter's gain capabilities.

A simple circuit arrangement which gives better frequency response is the common-emitter amplifier driven from a voltage (low-impedance) source. This arrangement, shown in Fig. 36, extends the frequency response of the 2N219 from 120 Kc to 2.9 Mc.

At this point, it is interesting to compare this common-emitter amplifier with a valve amplifier. We may do so easily, since the common-emitter amplifier is a voltage amplifier, and valves are also normally considered to be voltage amplifiers.

A typical pentode valve in the same circuit would have a bandwidth several times greater than the transistor bandwidth, but its gain would be only one-half to one-third as great. It would be possible to make a valve circuit which would behave almost like a 2N219, but to do so would require a pentode with a high transconductance — more than 22,000 μmhos — and would also

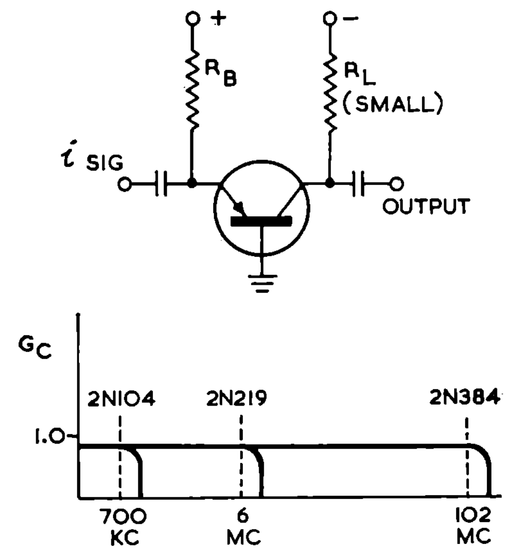


Fig. 34

require the use of a low-pass filter in the grid circuit, to simulate the poor frequency response of the transistor. The circuit of Fig. 37, which shows the pentode amplifier, also shows the low-pass filter.

The 70 per cent response point of the low-pass filter is at a frequency

$$f = \frac{1}{2\pi R_1 C_1} = \frac{1}{2\pi(85\Omega)(650\mu\mu f)} = 2.9 \text{ Mc}$$

which is the cutoff frequency for the 2N219 used in the voltage-driven common-emitter circuit of Fig. 36.

This comparison circuit may be extended to the case of the current-driven amplifier, as shown

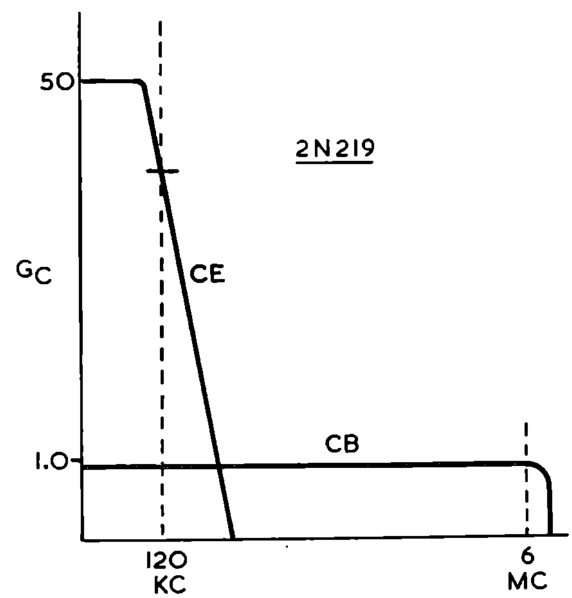


Fig. 35

in Fig. 38. Compare this arrangement with that of Fig. 36; they are identical except for the input arrangement. Now in order to compare this arrangement, the valve circuit is shown in Fig. 39; this circuit can be compared with Fig. 37. It is seen that the addition of R_2 is all that is necessary to change the circuit to the current-driven case.

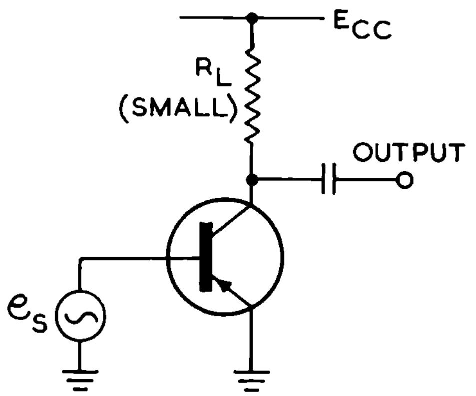


Fig. 36

In this case, the low-pass filter cuts off at

$$f\beta = \frac{1}{2\pi R_2 C_1} = \frac{1}{2\pi(2100\Omega)(650\mu\mu f)} = 120 \text{ Kc}$$

which is the frequency given above as $f\beta$ for the 2N219.

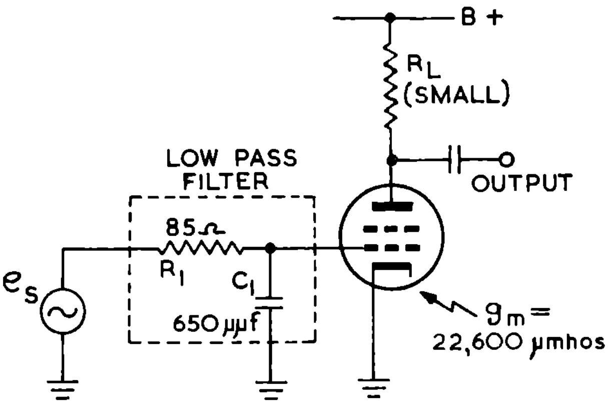


Fig. 37

In practical cases, transistors are not driven from perfect voltage sources (zero internal impedance) or from perfect current sources (infinite internal impedance), but from sources of some definite impedance. For example, a possible source for a 2N219 could have an internal impedance of 500 ohms. In this case, it can be shown that the original $f\beta$ bandwidth is improved by a factor of 4.6:

$$f^1 = f\beta\ (1 + \frac{R_2}{R_T})$$

$$= 120\text{Kc} \left[1 + \frac{2100}{500 + 85}\right]$$

$$= 120\text{Kc}\ (4.6)$$

$$= 550\text{Kc.}$$

Another simple circuit which extends bandwidth considerably is the circuit which includes an unbypassed emitter resistor. See Fig. 40.

In this case, the beta cutoff frequency is extended by an even larger factor, as shown in this expression:

$$f^1 = f\beta \left[1 + \frac{R_2}{R_T} + g_m R_e \frac{R_2}{R_T}\right]$$

Note that this expression is the same as that used above, except for the addition of one more term in the parentheses. Moreover, adding

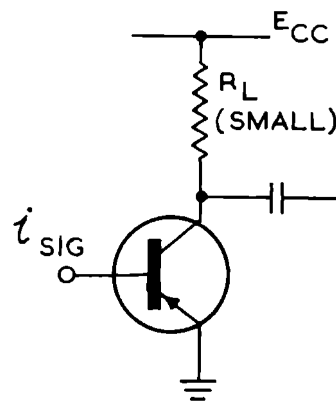


Fig. 38

R_e in the emitter causes a "reflection" of R_e to appear in series with R_1 in the low-pass filter, which means that a resistance equal to R_e must now be included in R_T, the total series resistance. If the value of R_e is, for example, 300 ohms, the new bandwidth is:

$$f^1 = 120\text{Kc} \left[1 + \frac{2100}{500 + 300 + 85} + (0.0226)(300)\frac{2100}{500 + 300 + 85}\right]$$

$$= 120\text{Kc}\ (1 + 2.36 + 16.0)$$

$$= 2.33 \text{ Mc.}$$

In all the foregoing discussion, the load resistor, R_L, was clearly tagged "small", in order to avoid, temporarily, the additional complications which arise when the load resistor is not small. The first complication is familiar to anyone who has used a valve in a wideband amplifier — stray capacity across R_L enters as an important bandwidth-limiting factor; just as in

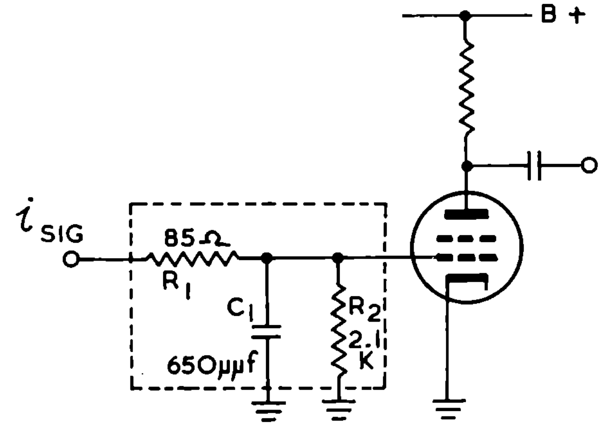


Fig. 39

valve practice, the effects of this capacity may be counteracted by an appropriate peaking network.

The other complication is also familiar to valve users — the so-called "Miller Effect". In a valve, a small feedback capacity — typically about $2\mu\mu f$ — appears between grid and plate. If the load resistor R_L is large enough to give this valve a gain of 7, for example, the small feedback capacity is multiplied by 1 + the gain, and appears as an 8-times larger capacitor shunting the input. In transistors, a similar effect takes place. Reverting to Fig. 40, if the load resistor is large enough to give a useful gain, of say 7 times, and we assume the feedback capacity (plate-grid) to be $9.5\mu\mu f$, then the feedback capacity is magnified by a factor of 1 + the gain, and appears across C_1. The magnified capacity appearing across C_1 will be

$$(1 + 7) \times 9.5\mu\mu f = 76.0\mu\mu f$$

This will, of course, affect the bandwidth of the circuit in the same way as would an increase of C_1 from $650\mu\mu f$ to $726\mu\mu f$.

The preceding examples used the 2N219. By present-day standards, this is a transistor of moderate frequency response, designed principally to be used as the mixer in transistorized broad-

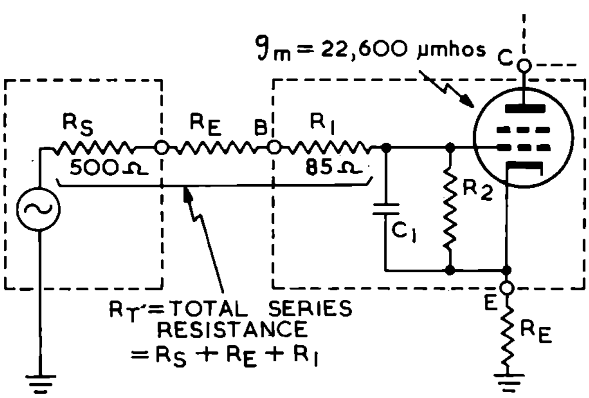


Fig. 40

cast-band receivers. The examples may be converted to show the behaviour of a 2N104 or a 2N384 by substituting the appropriate values in Table 2. The reader is invited to make the substitutions to extend his familiarity with the frequency response of various transistors.

TABLE 2

EQUIVALENT CIRCUIT PARAMETER	2N104	2N219	2N384	
R_1	290	85	50	Ω
R_2	1380	2100	1040	Ω
C_1	6900	650	90	μμf
C_f	40	9.5	1.3	μμf
g_m	32,000	22,600	56,800	μmhos

Pulse Operation

When a pulse is to be amplified, linearity is not a requirement of the amplifier. Consequently, the valve or transistor which is used as a pulse amplifier can be used as a switch, by driving it from cutoff to saturation: see Fig. 41. The output pulse is very large under these circumstances. Its peak valve is very nearly equal to the power supply voltage.

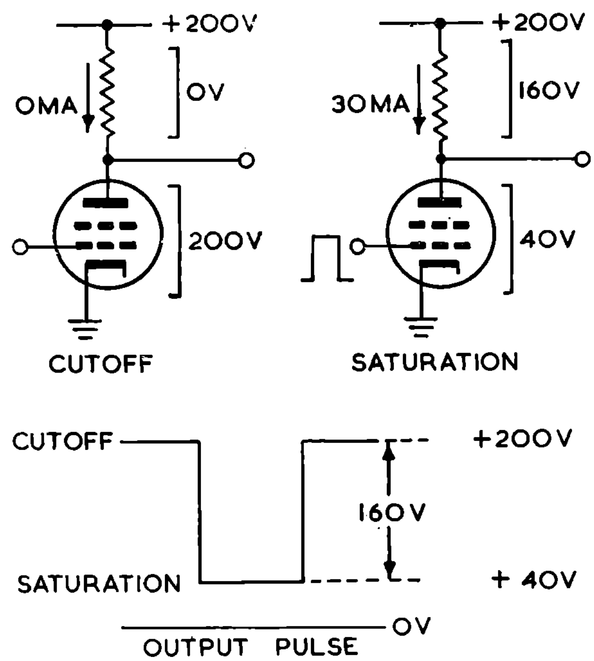


Fig. 41

Note that even in saturation, the valve still has a 40-volt drop across it. Since it draws 30ma in this condition, the power dissipated by the valve during the pulse is

P_p = IE = (30 ma) (40 volts) = 1.2 watts

and the power switched (peak pulse power) is

P_s = (30 ma) (160v) = 4.8 watts

which is 4 times as large as the valve dissipation.

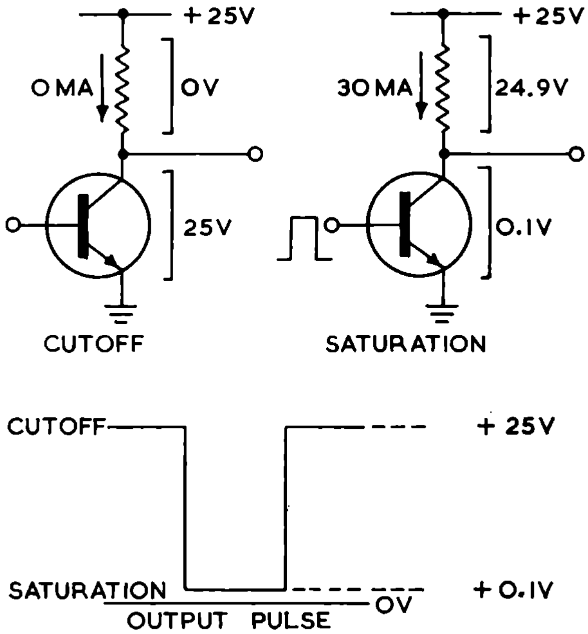


Fig. 42

A transistor can be used in much the same manner, but with two major differences; the output voltage is limited to (typically) 25 volts, and the drop across the saturated transistor is typically 0.1 volt: a typical case is shown in Fig. 42.

In this case, the transistor dissipates (during the pulse) only

$P_c = IE = (30 \text{ ma})\ (0.1 \text{ v}) = 3.0$ milliwatt

while the peak pulse power (power switched) is

$P_s = IE = (30 \text{ ma})\ (24.9 \text{ v}) = 747.0$ mw

While the valve switched only 4 times as much power as it dissipated, the transistor switched 249 times as much power as it dissipated. The transistor is obviously a very efficient switching device.

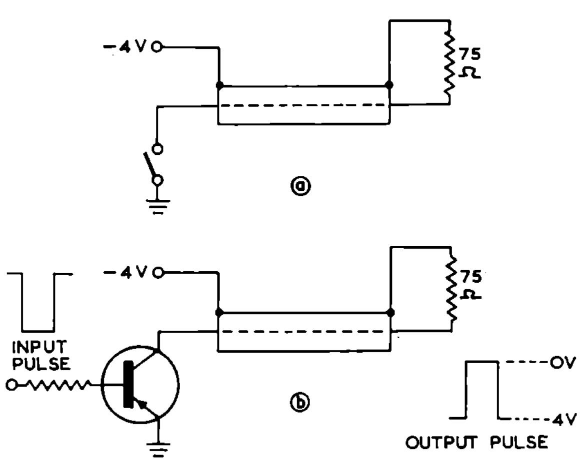


Fig. 43

The excellent switching characteristics of transistors make them very useful in such applications as driving a 4-volt pulse into a 75-ohm coaxial cable in video pulse distribution systems. In this application, the transistor behaves in the same manner as the switch in the circuit of Fig. 43a.

Every time the switch is closed and then opened, a 4-volt pulse appears across the 75-ohm resistor. If the switch is replaced by a transistor which is driven into saturation by an input pulse, as shown in Fig. 43b, the transistor will open and close like a switch, alternately disconnecting and connecting the battery and the coaxial cable, and thereby causing a pulse to apear across the 75-ohm terminating resistor.

One of the most attractive features of this circuit is that the transistor's excellent switching efficiency makes it possible to drive 4 volts into a 75-ohm line with a tiny, low-power transistor. The simple circuit shown has a number of drawbacks, such as failing to provide sending-end termination for the coaxial cable. However, this fault and others are easily remedied by more sophisticated circuitry, so circuits similar to it will no doubt find wide use in video pulse distribution systems.

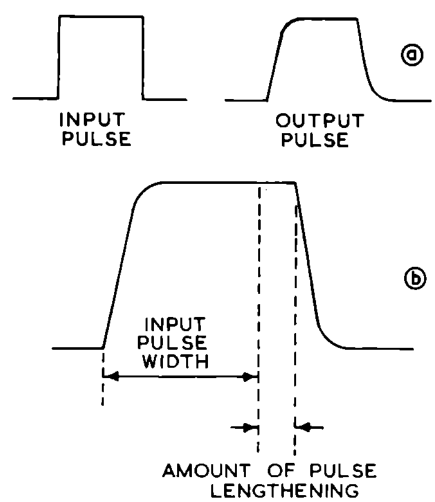


Fig. 44

Whenever a pulse is amplified — whether by valve or transistor — there is a tendency for the amplifier to degrade the pulse somewhat. In a valve amplifier, the stray capacities tend to degrade the rise and fall times of the pulse, in a manner shown exaggeratedly in Fig. 44a.

In a transistor amplifier, this action is worsenend by the various internal capacities of the transistor. These capacities were discussed in the preceding section of high-frequency amplifiers, although the exact values given there are not valid for large signal swings such as are commonly found in pulse amplifiers.

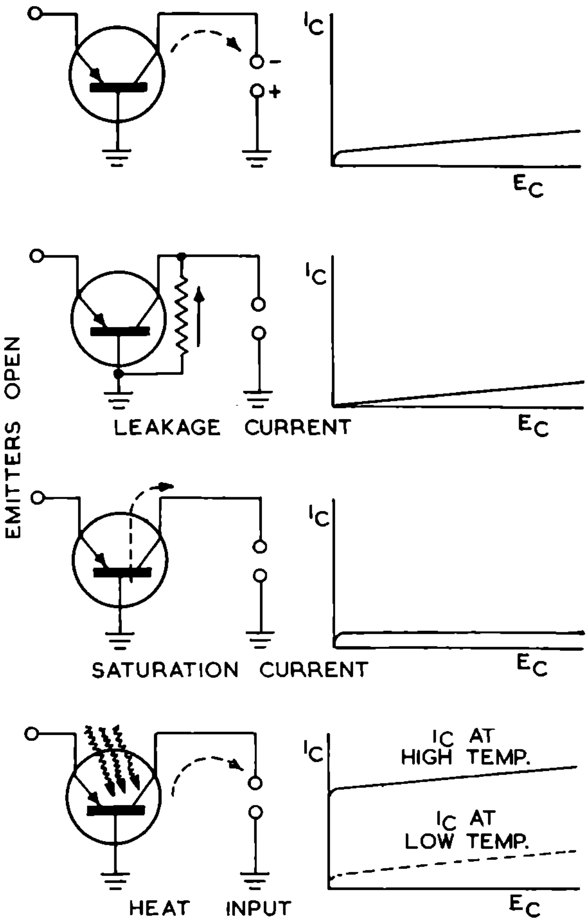


Fig. 45

In a saturated transistor pulse amplifier, another degradation enters in the form of pulse lengthening, which distorts the input pulse in the manner shown in Fig. 44b.

The degree of lengthening depends upon how hard the transistor is saturated. If the pulse width is not critical, a far amount of lengthening can be tolerated. However, if the input pulse must be faithfully reproduced at the output, careful engineering is needed to minimize the lengthening. The engineering problem is eased considerably by the availability of switching transistors which are constructed to minimize this effect.

Temperature Effects

In the first article of this series, the transistor was introduced as an extension of a junction diode. It was shown that the reverse-biased collector junction passed a small current, just as would be expected of any junction diode. See Fig. 45.

This current has two components. The first component results from simple resistive leakage paths, and the second component, called saturation current, results from a semiconductor action. Although the saturation current is little effected by changes in collector voltage, it is extremely sensitive to changes in temperature. As junction temperature is increased, the saturation current increases. The increase is very rapid; the current doubles about every 9°C. However, even at the highest operating temperatures, this relatively small current does not often cause trouble as long as the base is grounded.

However, the transistor is often operated with the emitter grounded. In this configuration, the tiny reverse* current must pass through the base-emitter junction as well as the base-collector junction, and in so doing, becomes amplified by the current-amplifying mechanism of the transistor. Consequently, the reverse current for the common-emitter configuration is beta times greater than the reverse current for the common-base configuration. With this larger current being doubled for every 9°C temperature rise, the high-temperature reverse current becomes a factor deserving serious consideration.

* Since the temperature problems are tied to only the saturation component of the total reverse current, the resistive (leakage) component will be ignored in this discussion, for the sake of simplification.

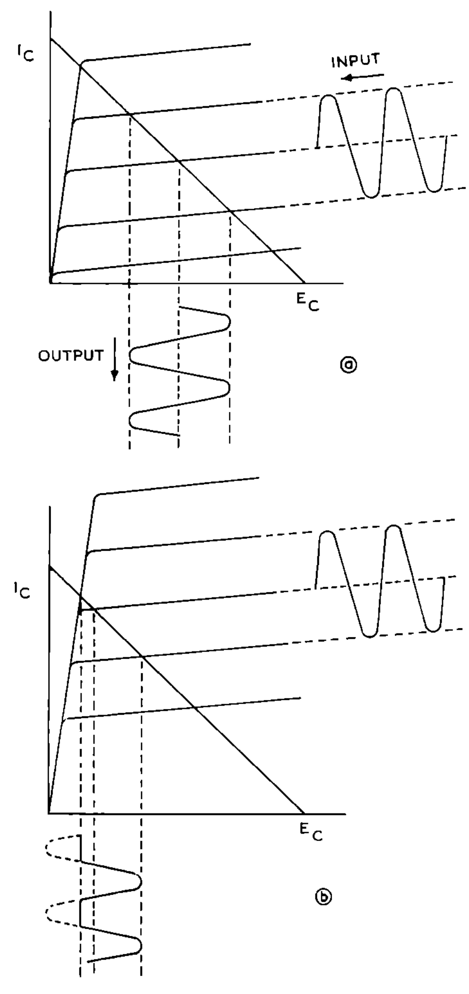
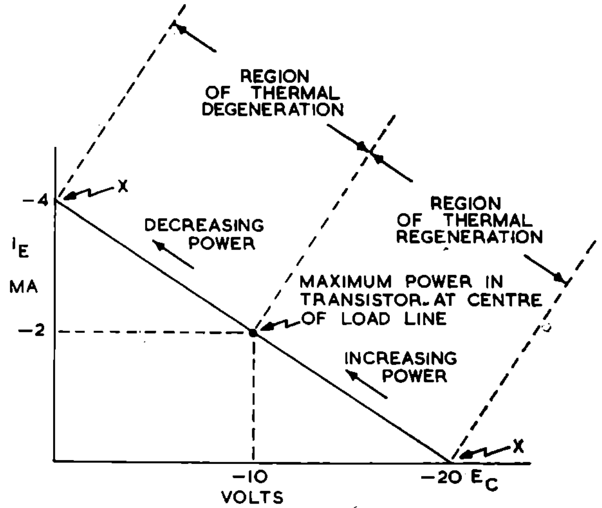


Fig. 46



AT TOP AND BOTTOM OF LOAD LINE, POINTS "X", POWER IN TRANSISTOR O.

Fig. 47

The family of curves is not distorted by the high-temperature reverse current, but merely displaced, intact, to another area of the graph. This displacement can seriously disturb the functioning of an amplifier. Consider the effect that elevated temperatures would have on a simple common-emitter amplifier similar to that of Fig. 33. At normal operating temperatures, a sine wave input gives a relatively distortion-free output, as shown in Fig. 46a. If the temperature is raised, however, the curves shift upwards, and the amplifier starts to clip, as shown in Fig 46b.

If the temperature is raised further, the signal is completely clipped, and the transistor is said to be saturated. Since no amplification can be obtained when the transistor is saturated, this situation must be carefully avoided, either by avoiding high temperatures, or by improved biasing circuitry, or both.

Note that the temperature referred to in all this discussion is not the temperature of the transistor's enviroment, but is rather the temperature of the collector junction itself. If the transistor is not drawing any current (as might be the case if it were in dead circuit or in storage) the junction temperature T_j and the ambient (environmental) temperature are the same.

However, when there is a power input, the heat dissipated at the junction will raise the junction temperature above the ambient temperature. For example, a 10-milliwatt power input might raise T_j by 5°C, and a 20-milliwatt power input by 10°C.

It is correct to deduce from these two cases that every 1 milliwatt of power input will cause a 0.5°C rise in junction temperature for this particular transistor. This characteristic of a transistor is called its thermal resistance, and would be listed in the data sheet of this particular transistor as having a value of 0.5°C/mw.

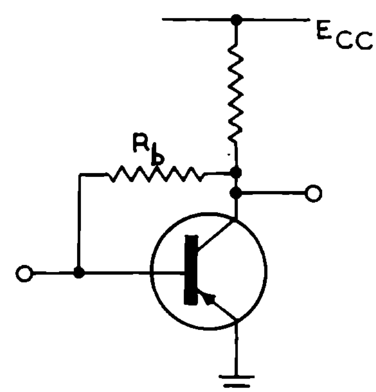


Fig. 49

The thermal resistance of a transistor is one of its more important characteristics. It enables the transistor user to compute the temperature of the junction itself, and, using this value of T_j, to compute the temperature-induced reverse current. For example, if a transistor has a reverse current of 6μa in a simple common-base circuit, it will have a reverse current of $\beta \times 6\mu a = 300\mu a$ in a simple common-emitter circuit.

If the simple single-resistor biasing technique is used (to exaggerate, for this example, the effects of temperature), and the circuit is adjusted for 54 mw dissipation, then a thermal resistance of 0.5°C/mw tells us that the junction temperature is higher than the ambient by

$$0.5°C/mw \times 54\ mw = 27°C;$$

and the actual T_j is $27° + 20° = 47°C$. Since the reverse current doubles every 9°C, it will be eight times larger in this 27° change, causing the

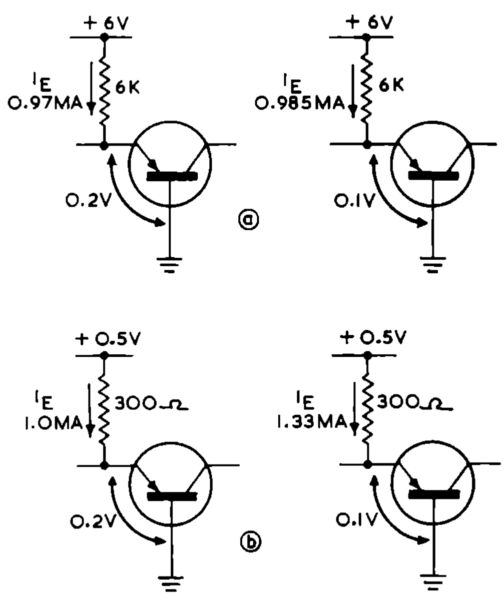


Fig. 48

300μa reverse current to become 2400μa or 2.4 ma.

The presence of this additional 2.4 ma could increase the dissipation to more than 54 milliwatts, thereby causing more heating of the junction, which would result in even more reverse current, thereby causing even more heating, and so on. This mechanism, which is called thermal regeneration, can have a number of undesirable effects. In its milder forms, it can cause the operating point to be unduly sensitive to ambient-temperature changes. More serious cases can drive the operating point into the high-current region, with the consequent possibility of clipping and distorting the signal. The most serious result of thermal regeneration, however, is thermal runaway, which is a thermal regeneration resulting in the transistor's self-destruction. Proper design can prevent both of the serious effects, and can also make even the mild effect completely negligible.

Thermal regeneration is not inevitable in transistor circuits. It occurs only when an increase in collector current causes an increase in collector power. This is true only for operating points falling on the lower half of the load line; See Fig. 47.

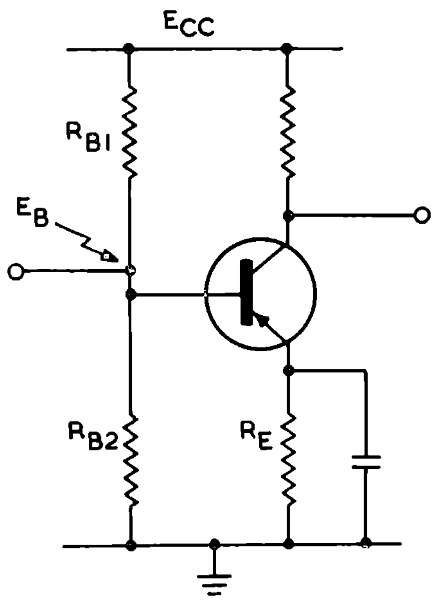


Fig. 50

When the operating point is on the lower half of the load line, an increase in current causes an increase in power; this is thermal regeneration. When the operating point is on the upper half of the load line, an increase in current causes a decrease in power; this is thermal degeneration.

It might seem that transistor circuits should always be biased into the degenerative re-ion as a safety precaution. However, this is usually not necessary. Proper circuit techniques make the regenerative region so stable that it is commonly

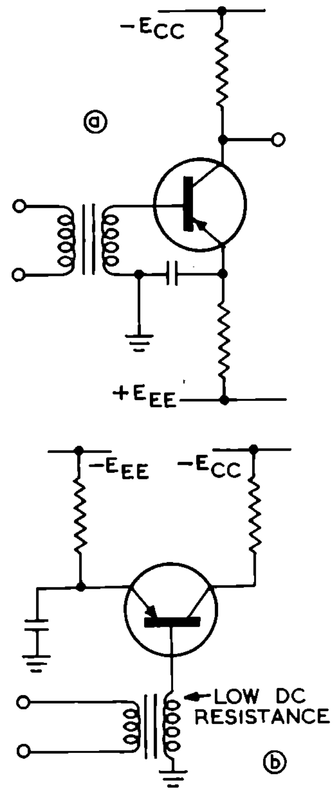


Fig. 51

used, particularly since it is the low-current region of the load line and therefore minimizes the current draw by the equipment.

The thermal resistance given in a transistor's data sheet may be used to compute the maximum power dissipation allowable at a given ambient temperature. A typical small germanium transistor usually has a maximum allowable junction temperature of 85°C If its thermal resistance is 0.5°C/mw, and the ambient temperature is 20°C, it can tolerate only enough power to raise the junction from 20 to 85°C, a change of 65°C.

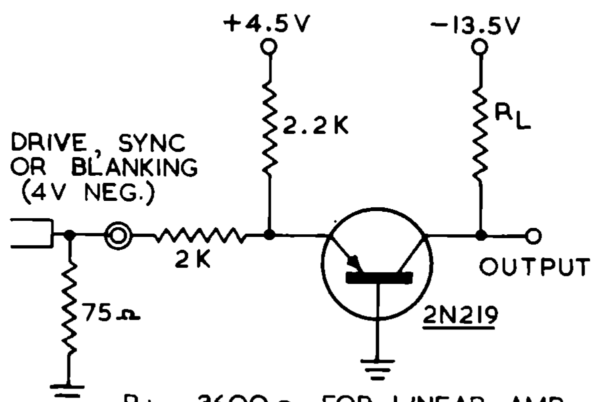


Fig. 52

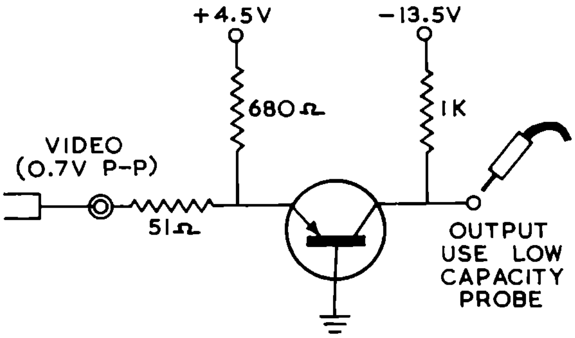


Fig. 53

Since each milliwatt contributes 0.5°C, 130 mw will give a rise of 65°C.

At this ambient temperature, 130 mw is the maximum allowable power input. At higher ambient temperatures, the maximum allowable power input is less than 130 mw, or, to put it another way if this transistor has a 50-mw power input (for example), the ambient temperature must not rise above 60°C, or the junction temperature will exceed 85°C. Moreover, at an ambient temperature of 85°C, the transistor cannot be operated at all since it cannot tolerate any power input.

The reverse current is a temperature effect related to the reverse-biased collector junction. There is another temperature effect related to the forward-biased emitter junction. The junction shown in Fig. 48a, which at normal temperatures has a voltage drop of about 0.2 volts across it, at higher temperatures has a smaller drop. This effect is of no consequence when the emitter is current-biased, but as voltage-biased conditions are approached, Fig. 48b, the bias currents can change appreciably with temperature. In general, voltage bias of the emitter should be avoided, unless the circuit can accommodate the changes in operating currents without giving improper operation.

The temperature dependence of transistors can be minimized by the proper circuitry. For example, the performance of the simple common-emitter amplifier can be improved by moving the biasing resistor from the power supply to the collector; see Fig. 49.

In this arrangement, any increase in collector current causes the voltage at the collector to decrease, thereby decreasing the bias current flowing through R_b into the base. When base current decreases, the collector current also decreases, thereby tending to restore the original operating condition.

A transistor may also be stabilized by the use of the circuit of Fig. 50. Here the fundamental requirement is that the bleeder current flowing through R_{b1} and R_{b2} be so much larger than the base current that changes in base current cannot influence the voltage at the midpoint of the R_{b1}/R_{b2} divider. Under this circumstance, the base voltage cannot change, and the circuit behaves like a biased-up common-base amplifier, insofar as the bias currents are concerned. Resistors R_{b1} and R_{b2} are effectively replaced by a battery connected between base and ground.

The emitter current (and therefore, the collector current) is determined principally by the "battery" voltage and the emitter resistor:

$$I_E = \frac{E_B}{R_E}$$

(This expression assumes that the "battery" voltage is much greater than 0.2 volts across the emitter junction.) The operating point is therefore almost independent of temperature, since neither E_B nor R_E is influenced by temperature. The two foregoing circuits are sometimes found combined in a single circuit for increased stability.

The common-base amplifier with current bias exhibits the best obtainable single-stage temperature independence. However, it lacks the gain capabilities of the common-emitter configuration. Some circuits can be arranged to make use of the good features of both the common-emitter and common-base configurations. This is accomplished by making the amplifier a common-emitter configuration for signal currents, and a common-base configuration for bias currents. This point is illustrated in Fig. 51, where two drawings of the same circuit are shown.

Conclusion

The information given in this series of three articles can be considered as nothing more than a narrow introduction to a broad subject. The material in these articles, properly assimilated, will facilitate a more detailed study of transistors, using one of the many excellent texts currently available.

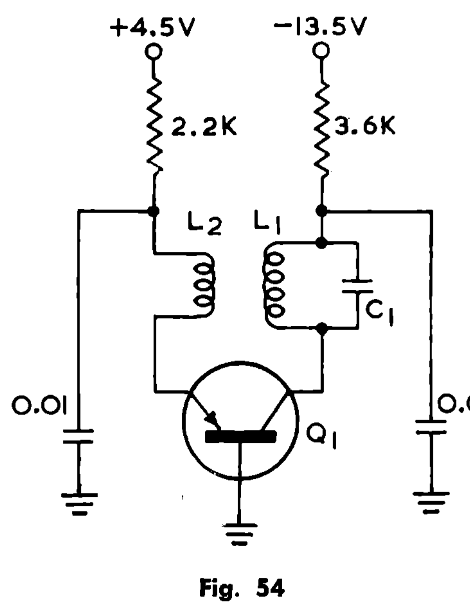


Fig. 54

The appendix which follows presents a few easily-constructed circuits to illustrate some of the principles discussed in the articles. Some of these circuits assume that the standard television pulse waveforms — sync, drive, blanking, and video — are available to the constructor. The other circuits require less elaborate facilities.

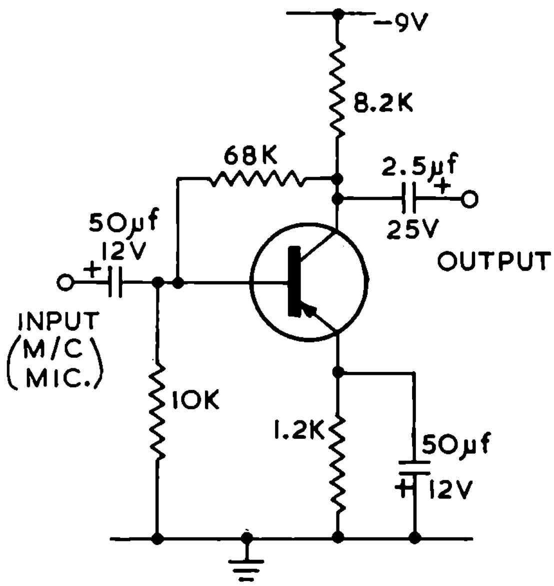


Fig. 55

APPENDIX

The following circuits have been designed to illustrate some of the principles of transistors. A 2N219, which is relatively inexpensive, will give very good results in any of the circuits except where noted otherwise. The potentials were chosen as those obtainable from inexpensive batteries.

A Common-Base Amplifier

(Fig. 52)

The 2N219 shown here can be replaced by almost any p-n-p transistor, although audio-type transistors (such as the 2N109) will give poorer pulse performance. To use an n-p-n transistor, change the $-13\frac{1}{2}$ volts to $+13\frac{1}{2}$ volts, and the $+4\frac{1}{2}$ volts to $-4\frac{1}{2}$ volts. For saturated operation with an n-p-n transistor, omit the 2200-ohm resistor and the $4\frac{1}{2}$-volt bias supply.

This amplifier has a gain of 1.8 as a linear amplifier (Gv — 3600/2000). As a saturated pulse-amplifier, its output is $13\frac{1}{2}$ volts of pulse.

A Common-Base Video Amplifier

(Fig. 53)

The transistor used here should have an alpha cutoff frequency somewhat greater than the desired bandwidth. As shown, a low-capacity probe should be used to view the waveform. An n-p-n transistor may be used by reversing the polarity of the power supply voltages.

A Common-Base Oscillator

(Fig. 54)

This simple oscillator will operate over a wide range of frequencies. For example, it will oscillate on the broadcast band if L_1 is 105-200 mh, C_1 about 180 $\mu\mu$f, L_2 about 20 turns of 26 wire wound between the pies of L_1, and Q_1 a 2N219. The circuit should oscillate to at least 12 Mc with the 2N219, with appropriate changes in the resonant tank L_1-C_1 and in the feedback winding L_2. (L_2 should have about one-sixth the number of turns of L_1.) With a 2N247, oscillations can be obtained to at least 60 Mc, and with a 2N384, to at least 150 Mc. If the circuit does not oscillate, try reversing the leads from L_2.

A Common-Emitter Microphone Preamplifier

(Fig. 55)

This circuit illustrates the temperature stabilization discussed in Part 3 of this article. The original circuit used a 2N109, but almost any p-n-p transistor should perform adequately. The circuit was designed to use a 2⅛″ speaker as a microphone. In such operation, one of the holes in the back of the speaker should be covered with cardboard having 1/32″ hole drilled in it. The remaining holes in the back of the speaker should be covered with felt, and the speaker mounted in a bafle or case.

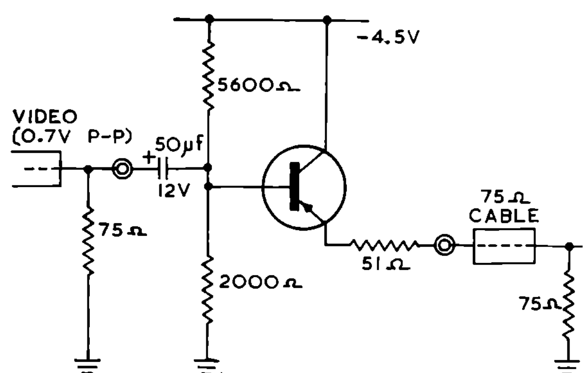


Fig. 56

A Common-Collector Line Driver

(Fig. 56)

This line-driver provides proper terminations for a 75-ohm line. It has a gain of one-half. Its bandwidth can be calculated from the expressions discussed in Part 3 of this article.

(With acknowledgements to RCA)

BOOK REVIEWS

"PRINCIPLES OF FREQUENCY MODULATION: Applications in Radio Transmitters and Receivers, and Radar." B. S. Camies. Iliffe & Sons Limited. Size 8¾" x 5½", 147 pages, 87 diagrams.

This book has a wide field of application. In concise form and logical sequence it gives a comprehensive account of the fundamentals of FM and its applications. Unlike most books on frequency modulation, which tend to concentrate on FM receivers, it also covers FM in transmitters, and the use of FM in microwave links, in radar, in telegraphy and in facsimile transmission.

The first part of the book deals with theory in some considerable detail, and is followed by a searching but very readable discussion of the relative advantages of AM and FM in the presence of interference. All this forms a very good basis for the material which follows, where the generation and detection of frequency-modulated waves receives a very full treatment.

Not the least interesting part of the book, and very valuable for the wider horizon it opens up, is the section dealing with the "non-broadcasting" applications of FM, such as radar and facsimile transmission. This would be a good book choice for students, technicians and engineers wanting to keep up with the fast-growing field of FM in broadcasting, television and mobile communication.

"TWO-WAY RADIO." A. H. Lytel. McGraw-Hill Book Coy. Inc. Size 9¼" x 6¼", 291 pages, 277 figures.

In the preface to this book the author says his purpose is to explain the growing field of two-way radio communications to the technicians and maintenance men entrusted with the installation, servicing and repair of this type of equipment. A perusal of the book shows that he has kept his objective well in view, and produced a valuable addition at a point where the literature of the radio art is rather thin.

The book deals with both AM and FM equipment, and in being written around commercial equipment has a realistic atmosphere which is most encouraging. It is also up to the minute in that recent techniques, such as transistorized power supplies, are covered. The treatment on antennas is good, whilst the chapter on selective calling systems is a must. Being intended for the technician, extensive chapters are provided on installation, servicing and test equipment, whilst the text throughout is profusely illustrated with photographs and diagrams.

"MICROWAVE DATA TABLES." A. E. Booth, M.I.R.E., Graduate I.E.E. Iliffe & Sons Limited. Size 10¼" x 7½". 61 pages.

This book comprises a collection of accurately computed tables which are primarily intended for use by those engaged in work involving waveguides and similar transmission lines. It contains 26 tables in all, and over 23,000 computed calculations. These were carried out on an electronic digital computer to at least one more figure than in the tables themselves, thus ensuring the accuracy of the tables. All the independent variables are measurable quantities and are given to the limit of measuring accuracy, while the dependent variables are given to one more figure than would be required for most practical purposes, and hence impose no accuracy limitations on practical design and development work.

The mathematical basis of all formulae used in the computations is stated, and comprehensive notes on the use of the tables are incorporated. As the tables are intended to be used as a desk companion, they are clearly printed on stout paper and are strongly bound to withstand constant usage in design office or laboratory. The decibel tables alone could render this book a worthwhile acquisition — they are the most extensive yet published.

AUSTRALIAN-MADE TRANSISTORS

We have had several requests lately for details of transistors made in Australia by AWV. Prospective transistor users naturally prefer to choose a local product as the choice offers several obvious advantages. No less than 35 transistors are in current local production at the time of going to press, including an impressive range of ten drift transistors. For the convenience of readers we have prepared a chart listing the types and tabulating their normal applications. This is followed on succeeding pages by a brief description of each of the 35 types.

TYPE	STRUCTURE	RADIO FREQUENCY					AUDIO FREQUENCY				SWITCHING			
							CLASS A		CLASSES A & B					
		HF AMP.	MIXER	OSCIL-LATOR	CONV-ERTER	IF AMP.	SMALL SIG.	DRIV-ER	LARGE SIG.	POWER	LOW SPEED	MED. SPEED	HIGH SPEED	HIGH VOLT.
2N109	PNP							■	■		■			
2N139	PNP			■		■						■		
2N140	PNP		■		■							■		
2N175	PNP						■							
2N176	PNP								■	■				
2N217	PNP							■	■		■			
2N218	PNP			■		■						■		
2N219	PNP		■		■							■		
2N220	PNP						■							
2N247	PNP	■	■	■	■	■							■	
2N270	PNP							■	■		■			
2N301	PNP								■	■				
2N301A	PNP								■	■				
2N351	PNP								■	■				
2N370	PNP	■												
2N371	PNP		■											
2N372	PNP			■										
2N373	PNP					■								
2N374	PNP				■									
2N376	PNP								■	■				
2N405	PNP							■						
2N406	PNP							■						
2N407	PNP							■	■					
2N408	PNP							■	■					
2N409	PNP			■		■								
2N410	PNP			■		■								
2N411	PNP		■		■									
2N412	PNP		■		■									
2N456	PNP								■	■	■			■
2N457	PNP								■	■	■			■
2N544	PNP	■												
2N591	PNP							■						
2N640	PNP	■												
2N641	PNP					■								
2N642	PNP				■									

AWV 2N109 is electrically identical with the AWV 2N217; it has a plug-in base.

AWV 2N139 is electrically identical with the 2N218; it has a plug-in base.

AWV 2N140 is electrically identical with the 2N219; it has a plug-in base.

AWV 2N175 is electrically identical with the 2N220; it has a plug-in base.

AWV 2N217 is used especially in class B push-pull power output stages of battery-operated portable radio receivers and audio amplifiers operating at power output levels of approximately 150 milliwatts. Intended especially for use in large-signal applications such as Class B audio service, and as a high-gain Class A driver device, the 2N217 has characteristics which permit the design of amplifiers requiring high sensitivity, low distortion, high power efficiency and low battery drain. These characteristics make this transistor particularly applicable to the design of battery-operated portable radio receivers having a transistorized output stage.

AWV 2N218 is designed especially for 455 Kc. intermediate frequency amplifier applications in transistorized portable radios and automobile radios operating from either a 6 or 12 volt supply. In a common emitter type of circuit, arranged to provide stability and interchangeability with some sacrifice in gain, the 2N218 features maximum power gain of 38 db.

AWV 2N219 has parameters controlled especially for converter and mixer-oscillator applications in standard AM broadcast-band transistorized portable radios and automobile radios operating from either a 6 or 12 volt supply. In a stabilised common emitter circuit the 2N219 features a conversion power gain of 30 db.

Moreover, the parameters of the transistor are controlled to provide satisfactory operation under low-voltage conditions.

AWV 2N220 is a low-noise transistor intended particularly for use in pre-amplifier or input stages of transistorized audio amplifiers operating from extremely small input signals. Free from microphonism and hum and having an extremely low noise factor, the 2N220 makes possible higher small-signal sensitivity of transistorized audio equipment such as hearing-aids, microphone preamplifiers, and recorders. In addition, the low noise factor and the low input impedance characteristics of the 2N220 permit the design of audio amplifiers in which the transistor is operated directly from low-impedance, low-level devices such as magnetic microphones and magnetic pick-ups without an input coupling transformer. In a common-emitter circuit, the 2N220 features an exceptionally low wide-band noise factor of 6 db maximum, a current amplification ratio of 65, and a matched-impedance power gain of approximately 43 db.

AWV 2N176 is designed for use in class A power output stages of audio-frequency amplifiers, particularly in automobile radio receivers. It may also be used in class B stages. In class A amplifier service at a mounting flange temperature of 80°C and with a dc supply voltage of —14.4 volts, the 2N176 can deliver a maximum-signal power output of approximately 2 watts with a power gain of 35.5 db.

AWV 2N247 is a drift transistor designed specifically for use as a radio-frequency amplifier in military and commercial equipment and in entertainment-type receivers operating at frequencies covering the AM broadcast band and up into the short-wave bands. In a unilateralized common emitter circuit with base input, this transistor can provide a power gain as high as 45 db.

AWV 2N270 is designed especially for use in large-signal audio-frequency applications such as single-ended or double-ended power output stages and high-gain class A driver stages of radio receivers and audio amplifiers. In push-pull class B amplifier service, two 2N270's can deliver a maximum-signal power output of approximately 500 milliwatts with a power gain of 32 db.

AWV 2N301 and 2N301-A are power transistors designed specifically for use in class A power output stages of audio frequency amplifiers, particularly in automobile radio receivers and military and commercial communications equipment. They are also useful in class B push-pull amplifier stages of such equipment. The 2N301-A differs from the 2N301 in that it has a higher maximum dc and peak collector-to-base voltage rating and is intended for those military and commercial applications requiring such high voltages.

AWV 2N351 is designed specifically for use in class A power output stages of audio frequency amplifiers, particularly in automobile radio receivers. It is also useful in class B push-pull amplifier stages of such equipment. In class A amplifier service at a mounting flange temperature of 80°C and with a dc supply voltage of —14.4 volts, the 2N351 can deliver a maximum-signal power output of approximately 4 watts with a power gain of 33.5 db.

AWV 2N370 is a drift transistor intended for rf amplifier service in military and commercial equipment and in portable domestic radio receivers operating at frequencies up to 23 Mc. The 2N370 is suitable for use in both AM broadcast receivers and short wave receivers. The 2N370 forms with the 2N371 and 2N372 a complement for high-gain rf tuners. The 2N370 in a common-emitter circuit without neutralization can provide a useful power gain of 31 db at 1.5 Mc, 17.6 db at 10 Mc, and 12.5 db at 20 Mc. Maximum power gain figures in a unilateralized circuit are 50.5 db, 26.2 db and 17 db respectively at the same frequencies.

AWV 2N371 is a drift transistor intended for rf local oscillator service in military and commercial equipment and in portable domestic radio receivers operating at frequencies up to 23 Mc. The 3N371 forms with the 2N370 and 2N372 a complement for high-gain rf tuners. The 2N371 can provide an oscillator-injection voltage to produce optimum mixing in such a tuner.

AWV 2N372 is a drift transistor intended for rf mixing service in military and commercial equipment and in portable domestic radio receivers operating at frequencies up to 23 Mc. The 2N372 in a common-emitter circuit without neutralization can provide a useful power gain of 31 db at 1.5 Mc, 17.6 db at 10 Mc, and 12.5 db at 20 Mc. Maximum power gain figures in a unilateralized circuit are 50.5 db, 26.2 db and 17 db respectively at the same frequencies.

AWV 2N373 is a drift transistor designed especially for 455 Kc if amplifier service in domestic battery-operated radio receivers. The 2N373 features exceptional stability, excellent uniformity of characteristics, and low feedback capacitance, which, together with closely-controlled small-signal parameters make this transistor especially useful in quantity-produced AM broadcast receivers. The 2N373 is capable of providing a useful power gain of 34 db in a common-emitter circuit without neutralization, with a maximum power gain of 57 db.

AWV 2N374 is a drift transistor especially designed for use as a converter (mixer-oscillator) in domestic battery-operated radio receivers operating in the AM broadcast band. The 2N374 features exceptional stability, excellent uniformity of characteristics, and low feedback capacitance, which, together with closely-controlled small-signal parameters make this transistor especially useful in quantity-produced AM receivers. The 2N374 is capable of providing a useful conversion power gain of 40 db in a common-emitter circuit with a dc collector/emitter voltage of —12 volts and an emitter current of 0.6 ma.

AWV 2N376 is designed specifically for use in class A power output stages of audio frequency amplifiers, particularly in automobile radio receivers. It is also useful in class B push-pull amplifier stages of such equipment. In class A amplifier service at a mounting flange temperature of 80°C and with a dc supply voltage of —14.4 volts, the 2N376 can deliver a maximum-signal power output of approximately 4 watts with a power gain of 35 db.

AWV 2N405 is electrically identical with the 2N406; it has a plug-in base.

AWV 2N406 is intended for use as a low-power class A af driver amplifier in battery-operated portable radio receivers. In a common-emitter circuit, the 2N406 features a typical small-signal current transfer ratio of 35, and a matched-impedance power gain of 43 db.

AWV 2N407 is electrically identical with the 2N408, it has a plug-in base.

AWV 2N408 is designed especially for use in class A and class B push-pull power output stages of battery-operated portable radio receivers, and in audio amplifiers operating at power output levels of approximately 150 milliwatts. In a common-emitter circuit, the 2N408 has a large-signal dc current transfer ratio (approximately linear to 50 ma) of 65, and a power gain (for two transistors in a class B push-pull circuit) of 33 db.

AWV 2N409 is electrically identical with the 2N410; it has a plug-in base.

AWV 2N410 is intended for 455 Kc if amplifier applications in transistorized portable radios. The 2N410 features a power gain of 31.2 db at 455 Kc in a common-emitter type of circuit employing fixed neutralization, and a maximum power gain of 38.8 db.

AWV 2N411 is electrically identical with the 2N412; it has a plug-in base.

AWV 2N412 has parameters closely controlled especially for converter and mixer-oscillator applications in standard AM broadcast band transistorized portable radios. In a common-emitter type of circuit employing fixed neutralization the 2N412 features a conversion power gain of 32 db at 1 Mc. Satisfactory operation under low-voltage conditions can be obtained in a suitably-designed circuit.

AWV 2N456 is a germanium p-n-p alloy type transistor intended for a wide variety of applications in military and industrial equipment. It is suitable for power switching, voltage regulators, multivibrator, dc-to-dc converter and power supply circuits, as well as relay-actuating devices. The 2N456 may also be used in low-frequency oscillator service and as a large-signal audio-frequency amplifier.

AWV 2N457 is a germanium p-n-p alloy type transistor intended for a wide variety of applications in military and industrial equipment. It is suitable for power switching, voltage regulators, multivibrator, dc-to-dc converter and power supply circuits, as well as relay-actuating devices. The 2N457 may also be used in low-frequency oscillator service and as a large-signal audio-frequency amplifier.

AWV 2N544 is a drift transistor designed for rf amplifier service in domestic battery-operated receivers and commercial communication receivers operating in the standard AM broadcast band. Characteristics are controlled to provide optimum performance in rf amplifiers designed for that band. In a typical neutralized rf amplifier circuit the 2N544 can provide a power gain of 30.4 db at 1.5 Mc, with a maximum power gain of 47.3 db. In addition, the very low value of collector/base capacitance of 1.65 $\mu\mu$f makes possible satisfactory gains the AM broadcast band without the use of neutralization.

AWV 2N591 is intended for use in large-signal class A audio frequency driver stages of automobile radio receivers. In class A af driver amplifier service at a dc supply voltage of —14.4 volts, the 2N591 can provide a power gain of 41 db with total harmonic distortion of only 3 per cent, measured at a power output of 5 milliwatts.

AWV2N640 is a germanium p-n-p alloy drift transistor intended for use as an rf amplifier in AM broadcast band automobile receivers. In an unneutralized circuit the 2N640 is capable of providing a maximum power gain of 47.5 db.

AWV 2N641 is a germanium p-n-p alloy drift transistor intended for use as an if amplifier at frequencies of 262.5 Kc or 455 Kc in automobile receivers. In a neutralized circuit the 2N641 is capable of providing a useful power gain of 41 db at 262.5 Kc and 40 db at 455 Kc, with a maximum power gain of 60 db.

AWV 2N642 is a germanium p-n-p alloy drift transistor intended for use as a frequency converter in AM broadcast band automobile receivers. The 2N642 provides a useful conversion power gain of 40 db at 1 Mc, and a maximum power gain of 50 db.

NEW RELEASES

(Continued from page 53)

RADIOTRON 7551, 7558

The 7551 and 7558 are two new beam power valves of the 9-pin miniature type. The 7551 is designed for use in mobile communications equipment operating from 6-cell storage-battery systems; the 7558 is designed primarily for use in fixed-station and other communications equipment using 6.3-volt heater supplies. In such communications equipment, these tubes are particularly useful in class C rf-amplifier, oscillator, and frequency-multiplier service, and can be operated at full ratings up to 175 Mc. They are also useful in modulator and af power-amplifier applications.

In rf power-amplifier and oscillator service, the 7551 and 7558 each have a maximum plate-input rating of 24 watts (ICAS). Each type can supply a useful power output of approximately 10 watts for a driver power of 1.5 watts. As a frequency doubler up to 175 Mc, each type can supply a useful power output of 4.5 watts for a driving power of approximately 0.6 watt. As a frequency tripler to 175 Mc, each type can supply approximately 2.3 watts for a driving power of 0.6 watt. The heater of the 7551 is designed to operate within a voltage range of 12 to 15 volts and will withstand momentary excursions from 11 to 17 volts.

RADIOTRON 7552, 7554

The 7552 and 7554 are high-mu triodes featuring ceramic-metal construction and space-saving pencil-tube design. They are especially suited for uhf service in portable field equipment, missile-guidance systems, and satellite-communication applications. The ceramic-metal construction permits a smaller, more sturdy tube which can operate at plate-seal temperatures up to 225°C., and shows evidence of greater endurance to nuclear radiation than glass-metal construction.

Additional design features of the 7552 and 7554 include: large cathode area as compared with comparable planar types, fast warm-up time —only 12 seconds to reach 90 per cent of dc operating plate current, and excellent thermal stability. Each type can be operated at altitudes up to 100,000 feet without pressurization and has a maximum plate-dissipation rating of 2.5 watts. The 7552 is designed to class A rf amplifier service up to 1000 Mc and above; the 7554 for class C operation as an oscillator, rf-amplifier, and frequency-multiplier tube up to 3000 Mc and above.

AWV