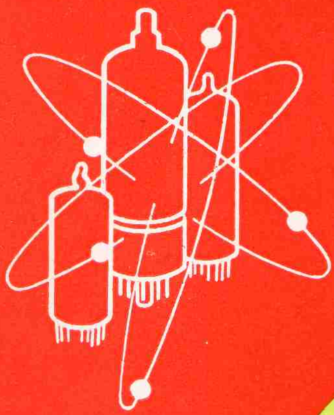

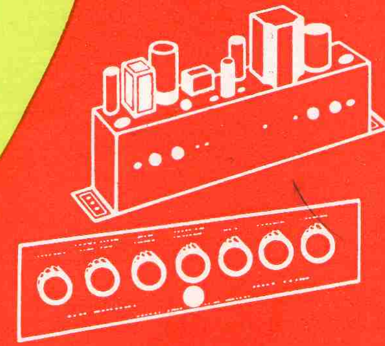

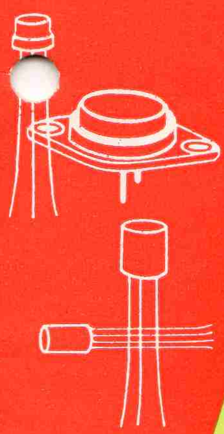

RADIOTRONICS

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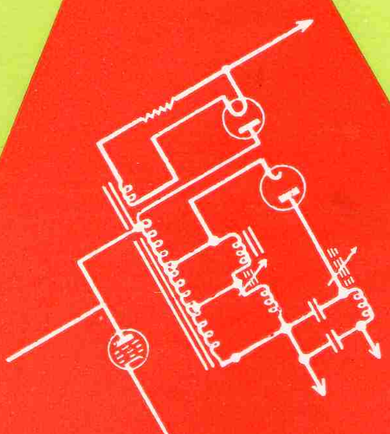

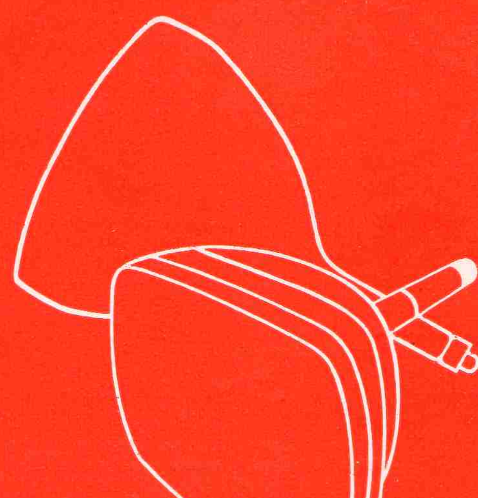

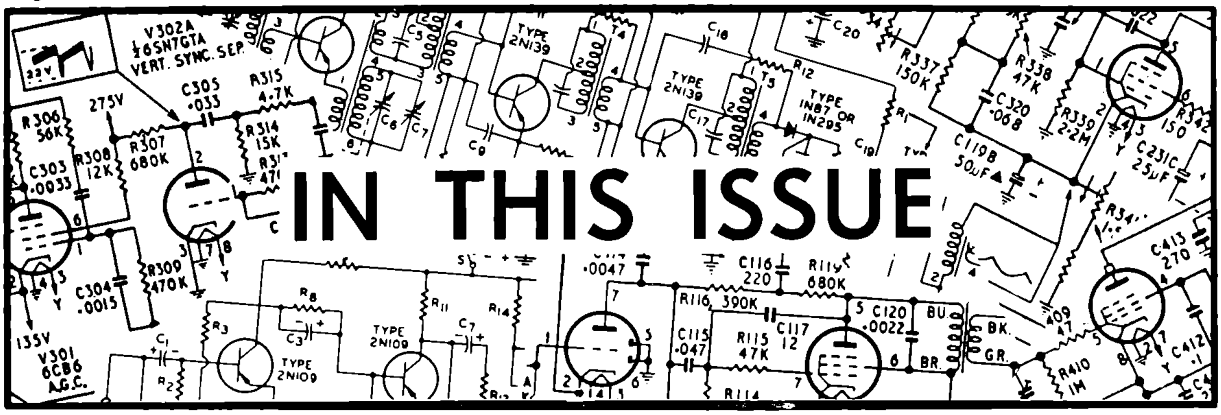

IN THIS ISSUE

Sound Reproducing Systems ---

• Monaural • Binaural • Monophonic • Stereophonic

by HARRY F. OLSON

A clarification of the definitions of the various types of systems encountered currently, as presented by a recognized authority in both acoustics and sound reproduction.

The reproduction of sound is the process of picking up sound at one point and reproducing it either at the same point or some other point, either at the same time or some subsequent time. There are many different types of systems employed for the reproduction of sound. In this connection, sound reproducing systems in use today may be classified as follows: monaural, binaural, monophonic, and stereophonic. There appears to be considerable confusion in the proper use of these terms in designating the four fundamental types of sound reproducing systems. The result is an almost indiscriminate application of the terms to unrelated systems. For this reason it appears desirable to define and describe the use of the four terms. Therefore, it is the purpose of this paper to define and describe the characteristics of monaural, binaural, monophonic, and stereophonic sound systems.

Monaural

A monaural sound reproducing system is a closed circuit type of sound reproducing system in which one or more microphones are connected to a single transducing channel which in turn is coupled to one or two telephone receivers worn by the listener, as in Fig. 1.

Binaural

A binaural sound reproducing system is a closed circuit type of sound reproducing system in which two microphones, used to pick up the original sound, are each connected to two independent corresponding transducing channels which in turn are coupled to two independent corresponding telephone receivers worn by the listener, as in Fig. 2.

Monophonic

A monophonic sound reproducing system is a field type sound reproducing system in which one or more microphones, used to pick up the original sound, are coupled to a single transducing channel which in turn is coupled to one or more loudspeakers in reproduction, as in Fig. 3.

Stereophonic

A stereophonic sound reproducing system is a field type sound reproducing system in which two or more microphones, used to pick up the original sound, are each coupled to a corresponding number of independent transducing channels which in turn are each coupled to a corresponding number of loudspeakers arranged in substantial geometrical correspondence to that of the microphones, as in Fig. 4.

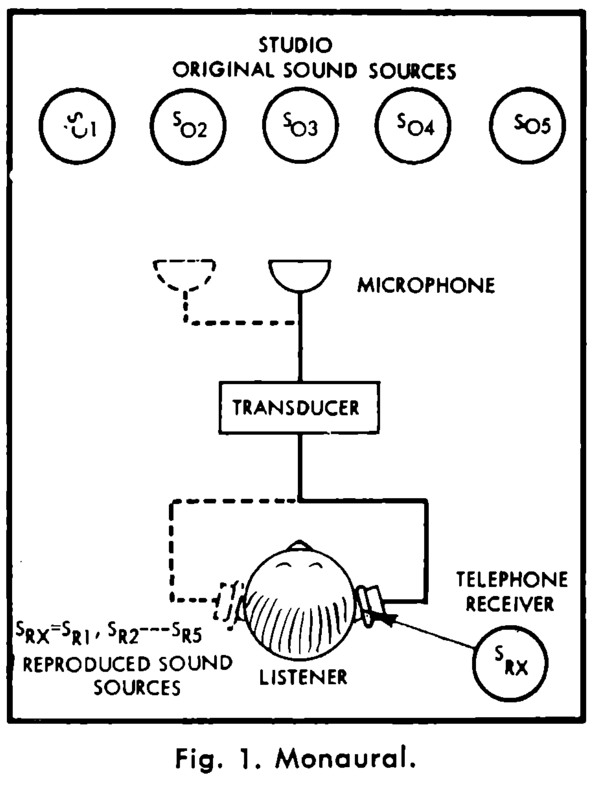

Fig. 1. Monaural.

Description of Systems

Following the definitions[1] of monaural, binaural, monophonic, and stereophonic sound the next consideration will be a description of some of the characteristics of the four systems.

To achieve realism in a sound reproducing system, four fundamental conditions must be satisfied, as follows:

1. The frequency range must be such as to include without frequency discrimination all of the audible components of the various sounds to be reproduced.

2. The volume range must be such as to permit noiseless and distortionless reproduction of the entire range of intensity associated with the sounds.

3. The reverberation characteristics of the original sound should be approximated in the reproduced sound.

4. The spatial sound pattern of the original sound should be preserved in the reproduced sound.

A diagram of a monaural sound reproducing system is shown in Fig. 1. The most common example of a monaural sound reproducing system is the telephone in which there is, in general, a single source of sound, one microphone, a transducer, and one telephone receiver coupled to one ear of the listener. In most local applications, the carbon microphone is coupled directly to the telephone receiver. In long distance telephony, valve and transistor amplifiers may be used between the microphone and loudspeaker. For other more limited applications, as for example, monitoring purposes, the transducer may be a radio transmitter and receiver, a television sound transmitter and receiver, a disc phonograph recorder and reproducer, a sound motion picture recorder and a reproducer and/or a magnetic tape recorder and reproducer. In some applications, there may be more than one sound source. One or more microphones may be used. In some applications two telephone receivers may be used transmitting

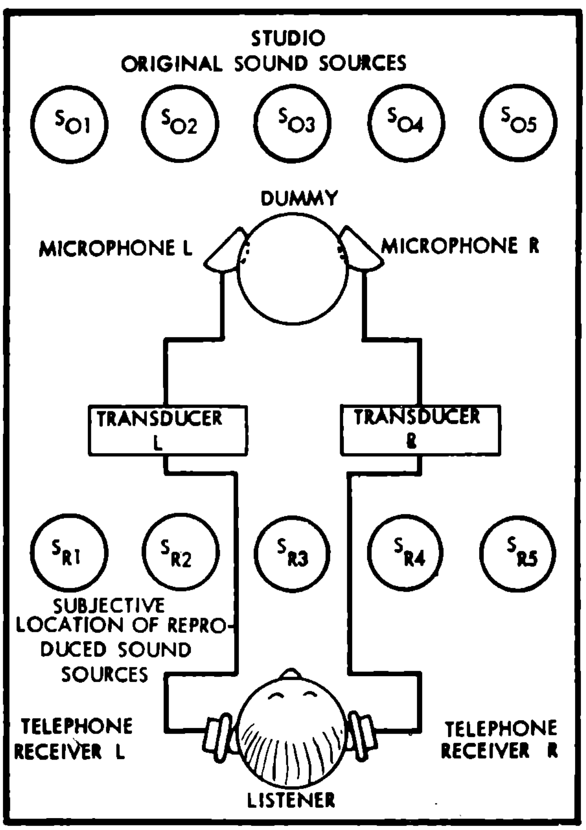

Fig. 2. Binaural.

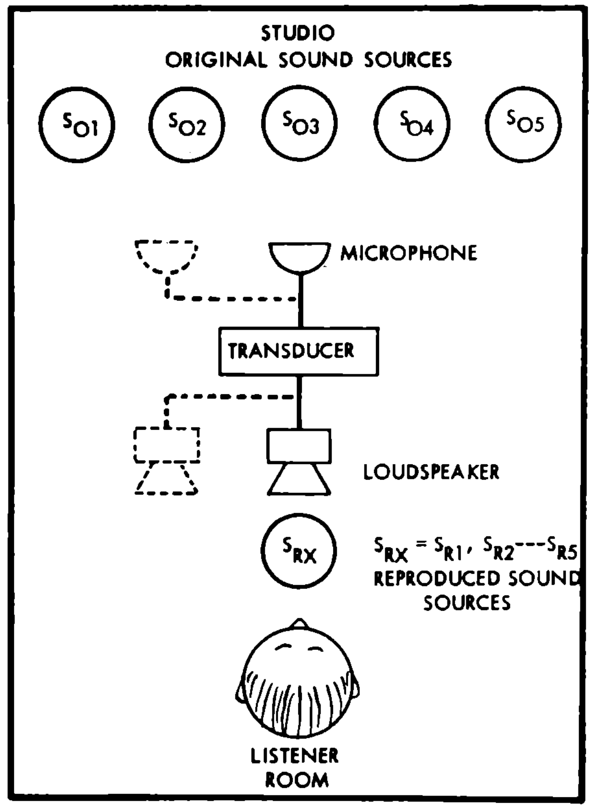

Fig. 3. Monophonic.

[1] The definitions of the terms monaural, binaural, monophonic and stereophonic, agree substantially with those of modern dictionaries. In addition, the terms binaural and stereophonic as defined in this paper have been standardized. As a result, the incorrect usage of binaural to designate a stereophonic system is disappearing. Monaural is still incorrectly used to designate a single-channel field-type sound reproducing system. Monophonic is a relatively new term, which has been introduced to supply a void in terms to describe the four fundamental sound systems. Monophonic and stereophonic are harmonious and congruent terms which complement each other and have a common relationship in describing field-type sound systems. Monaural and binaural are also harmonious and congruent terms which complement each other and have a common relationship in describing closed-circuit sound systems.

the same program to each of the ears of the listener. The monaural sound reproduction system is of the closed-circuit type in which the ear of the listener is transferred to a microphone location by means of the microphone, transducer, and telephone receiver combination. The acoustics of a single room are involved in the reproduction of the sound, namely, the studio in which the microphone is located. The monaural sound reproducing system may be constructed so as to satisfy conditions 1, 2, and 3 on realism of sound reproduction. It cannot, under any conditions, satisfy condition 4.

A diagram of a binaural sound reproducing system is shown in Fig. 2. There is no widespread use of the binaural sound reproducing system. The use is limited to specific applications. The binaural sound reproducing system consists of two separate channels. Each channel consists of a microphone, transducer, and telephone receiver. The microphones are mounted in a dummy simulating the human head in shape and dimensions and at the locations corresponding to the ears of the human head. The transducer may be an amplifier, a radio transmitter and receiver, a phonograph recorder and reproducer, a motion picture recorder and reproducer, or a magnetic tape recorder and reproducer. The binaural sound reproducing system is of the closed-circuit type. The listener is transferred to the location of the dummy by means of a two-channel sound reproducing system. The binaural sound reproducing system may be constructed so as to satisfy all four conditions on realism of sound reproduction.

A diagram of a monophonic sound reproducing system is shown in Fig. 3. It is the most widely employed of all sound reproducing systems. Examples are the disc phonograph, radio, sound motion picture, television, magnetic tape reproducer and sound systems. The monophonic sound reproducing system is of the field type, in which the sound is picked up by a microphone and reproduced by means of a loudspeaker into a field. The sound at the microphone is reproduced at the loudspeaker. The transducer may be an amplifier, radio transmitter and receiver, a phonograph recorder and reproducer, a sound motion picture recorder and reproducer, a television transmitter and receiver, a magnetic tape recorder and reproducer. The monophonic sound reproducer may be constructed to satisfy conditions 1, 2 and 3 on realism of sound reproduction. It cannot under any conditions satisfy condition 4.

A diagram of a stereophonic sound reproducing system is shown in Fig. 4. The stereophonic sound reproducing system is of the field type, in which the sound is picked up by two or more microphones which are coupled to a corresponding number of independent transducing channels which in turn are coupled to corresponding number of loudspeakers arranged in substantial geometrical correspondence to that of the microphones. The transducer may be an amplifier, radio transmitter and receiver, a phonograph recorder and reproducer, a sound motion picture recorder and reproducer, a television transmitter and receiver, or a magnetic tape recorder and reproducer. Two channels are used in the disc phonograph and radio. Two and three channels are used in the magnetic tape reproducer. Two, three and more channels are used in motion picture reproducers. The stereophonic sound reproducer may be constructed to satisfy conditions 1, 2 and 3 on realism of sound reproduction. It can be constructed to provide auditory perspective of the reproduced sound and in this sense the stereophonic sound reproducer satisfies condition 4 on realism of sound reproduction. Stereophonic sound is being rapidly commercialized. The first wide scale use was in sound motion pictures. This was followed by the magnetic tape reproducer. The stereophonic disc phonograph is being commercialized this year. Experiments are now being conducted in the transmission and reproduction of stereophonic sound by means of a radio system. In one arrangement, the two channels are transmitted on two separate radio links, one by a frequency modulation system and the other by an amplitude modulation system. In another arrangement, the two channels are transmitted

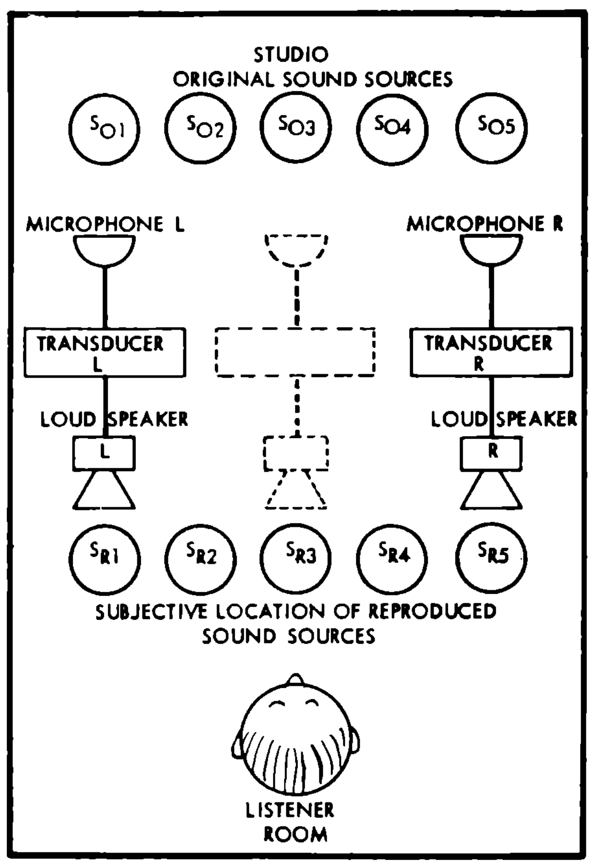

Fig. 4. Stereophonic.

and reproduced by means of a multiplex frequency modulation system.

Summary

The four fundamental types of sound reproducing systems—namely, monaural, binaural, monophonic, and stereophonic—have been defined and described in this paper. The terms monaural and binaural are used to designate closed circuit sound reproducing systems. The terms monophonic and stereophonic are terms used to designate field-type sound reproducing systems. Monaural and binaural (or monophonic and stereophonic) are mutually harmonious and congruent terms which complement each other in describing closed-circuit type (or field type) sound reproducing systems. The definitions as presented in this paper agree substantially with modern dictionaries. The terms binaural and stereophonic have been standardized. In view of this and the logic presented in this paper it is only a question of time until all four terms, monaural, binaural, monophonic, and stereophonic are standardized.

Acknowledgement

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FEATURES OF FLUORESCENT SCREENS

The fluorescent screens of cathode-ray tubes are identified according to phosphor number, e.g., P1, P2, P4, P5, P7, etc.

Phosphor P1 produces a brilliant spot having yellowish-green fluorescence and medium persistence. Types having this phosphor are particularly useful for general oscillographic applications in which recurrent-wave phenomena are to be observed visually.

Phosphor P2 is a medium-persistence screen which exhibits yellowish-green fluorescence and phosphorescence. The phosphorescence may persist for over a minute under conditions of adequate excitation and low-ambient light. Types utilizing this phosphor are particularly useful for observing either low- or medium-speed non-recurring phenomena.

Phosphor P4 is a highly efficient screen having white fluorescence and medium-short persistence. Types having this phosphor are of particular interest for television picture tubes, e.g., 21CEP4.

Phosphor P5 produces a highly actinic spot having blue fluorescence and medium-short persistence. Types having this phosphor are especially useful in photographic applications, involving film moving at very high speeds.

Phosphor P7 is a very long-persistence, cascade (two-layer) screen. During excitation by the electron beam, this phosphor produces a purplish-blue fluorescence. After excitation, the screen exhibits a yellowish-green phosphorescence which persists for several minutes. Types having this phosphor are particularly useful where either extremely low-speed recurrent phenomena or medium-speed non-recurrent phenomena are to be observed.

Phosphor P11 produces a brilliant actinic spot of blue fluorescence and medium-short persistence to permit its use in all photographic applications except those in which film moves at high speed. P11 screens, because of their unusually high brightness characteristic, may also be used for visual observation of phenomena.

Phosphor P12 is a long-persistence phosphor which exhibits both yellowish-orange fluorescence and phosphorescence. Types utilizing this phosphor are particularly useful for observing low- and medium-speed recurring phenomena.

Phosphor P14 is a long-persistence cascade (two-layer) screen. During excitation by the electron beam, this phosphor exhibits purplish-blue fluorescence. After excitation, it exhibits a yellowish-orange phosphorescence which persists for a little over a minute. Types utilizing this phosphor are particularly useful for observing either low- and medium-speed non-recurring phenomena or high-speed recurring phenomena.

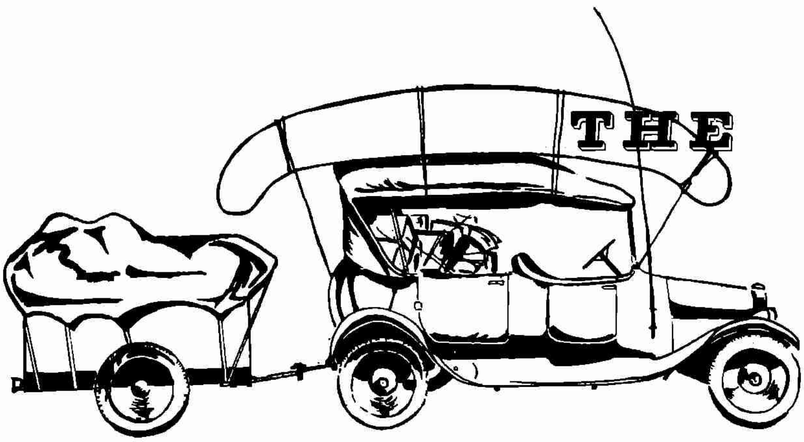
Phosphor P15 has radiation in the visible green region and in the invisible near-ultraviolet region. The ultraviolet radiation has short persistence which is appreciably shorter than that of the visible radiation. This phosphor finds application in flying-spot cathode-ray tubes.

Phosphor P16 has violet as well as near-ultraviolet fluorescence and phosphorescence with very short persistence. This phosphor has a stable, exponential decay characteristic and is particularly useful for the high-speed scanning requirements of a flying-spot video-signal generator.

Phosphor P20 has high luminous efficiency, yellow-green fluorescence and medium-short persistence. The screen may be used in applications requiring relatively short persistence and good visual efficiency.

Phosphor P22 is the designation for three separate phosphors used in combination in a colour picture tube. The separate phosphors are blue, green, and red, respectively. The persistence of the group phosphorescence is classified as medium.

Phosphor P24 is a short-persistence phosphor with green fluorescence and phosphorescence. Its spectral-energy emission characteristic has sufficient range to provide useable energy over the visible spectrum required for generating colour signals from colour transparencies.



THE WEEKEND SPECIAL

A COMPLETE, PORTABLE 40-METRE CW STATION

by Lee Aurick, W2QEX

For many hams, a weekend jaunt or a vacation trip with the family means being off the air for the duration. The author, to whom such trips meant a sacrifice of practically the only time available for work at the home station, undertook to solve the problem by the design of a portable 40-metre cw station which would fit unobtrusively into the family luggage and yet provide a high degree of operating convenience and efficiency.

In planning the station, the author considered the following features essential:

1. The entire station should fit in a portable typewriter case.
2. The transmitter should have a vfo and provision for oscillator "spoting".
3. To assure freedom from objectionable frequency variations during operation under marginal conditions, the vfo should have a regulated plate-voltage supply.
4. The final should load properly when connected to a 72-ohm load (pre-cut 40-metre doublet with coaxial feed).
5. The transmitter should include a single tuning and keying monitor.
6. Changeover from "transmit" to "receive" should be a one-switch operation.
7. The receiver should provide good bandspread for the 40-metre band.
8. The receiver should deliver sufficient af output power to operate a small built-in speaker.
9. The entire station should use proved circuits and be of minimum cost.

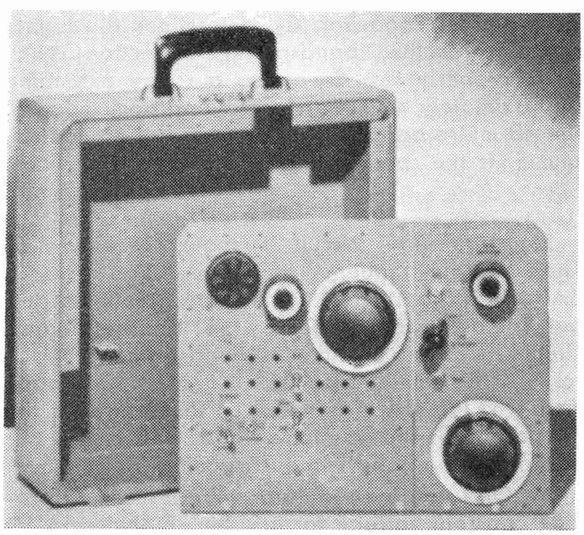
The rig shown in the accompanying photographs and circuit diagrams fulfills all these design requirements with one minor exception—the 66-foot doublet antenna and its 50-foot 72-ohm feeder, of course, do not fit easily into the portable typewriter case. They have to be carried elsewhere in the family luggage.

* 6AQ5-A has controlled heater warm-up time. It is otherwise identical with the 6AQ5.

Circuit Details

The limitations on the size and cost of the station dictated the use of a two-valve regenerative receiver. The one which seemed to offer the most advantages and best met the other requirements was the "Novice Special" described by Mix in QST for June, 1956. With minor modifications (a slight change in the method of tuning, and the use of a permanently mounted 40-metre coil instead of plug-in coils), this receiver was adopted. The power supply described by Mix for use with the receiver was also adopted, and used for the transmitter as well as the receiver.

As shown in Figure 1, 6AQ5-A's* are used in both the detector and af amplifier stages. The detector provides smooth and stable regeneration,



Four brackets made of $\frac{3}{4}$-inch aluminium angle are utilized. One supports the power-supply "deck". The others, mounted on three sides of a portable typewriter case, support the front panels of W2QEX's 40-metre cw station.

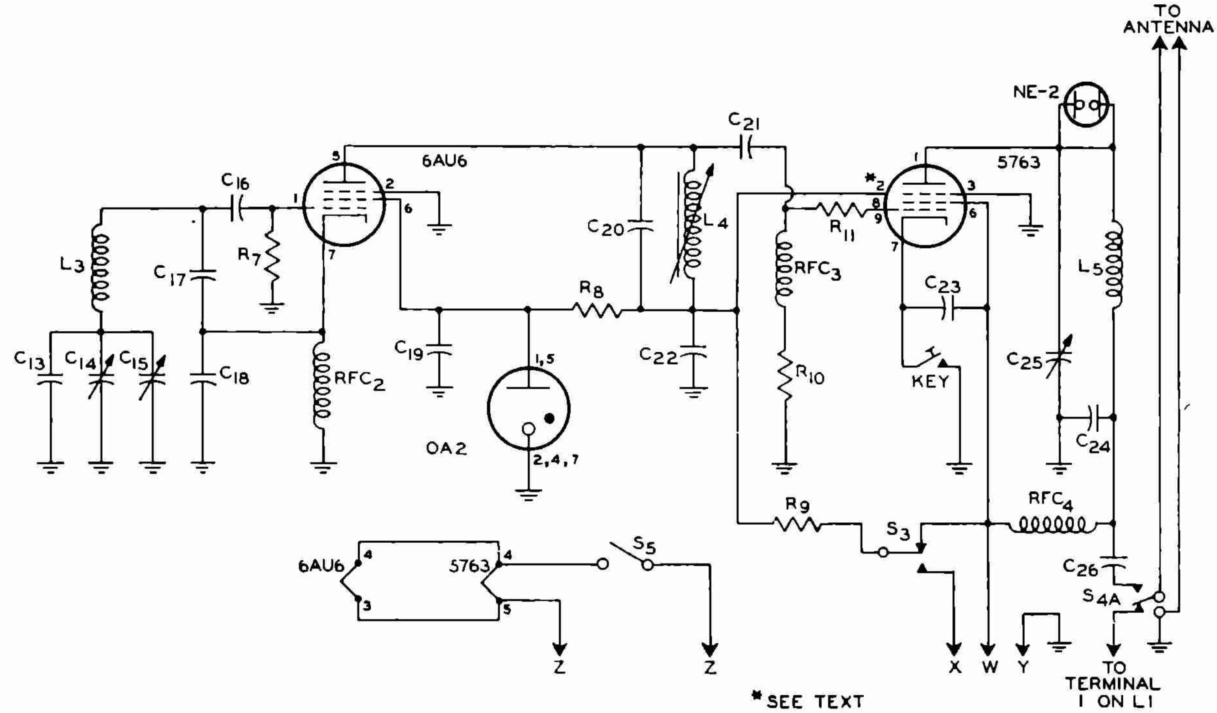


Figure 1: Circuit of the two-valve regenerative receiver.

and the tuning arrangement spreads the 40-metre band over 70 divisions (10 to 80) on the tuning dial. The af amplifier stage delivers sufficient output to operate the built-in speaker on practically every station that can be heard.

L_1 is a 9-turn inductor, air-spaced with one inch internal diameter, tapped at 2 turns (terminal 2), $4\frac{1}{4}$ turns (terminal 3), and 5 turns (terminal 4). C_3 is the "bandset" capacitor which, with the fixed mica padder capacitor, C_2, determines the tuning range. C_1 is the "bandspread" capacitor. When C_3 is properly set, C_1 covers a range extending approximately 40 kilocycles beyond each edge of the 40-metre band. R_2 is the regeneration control, and S_1 is the speaker-headphone selector switch.

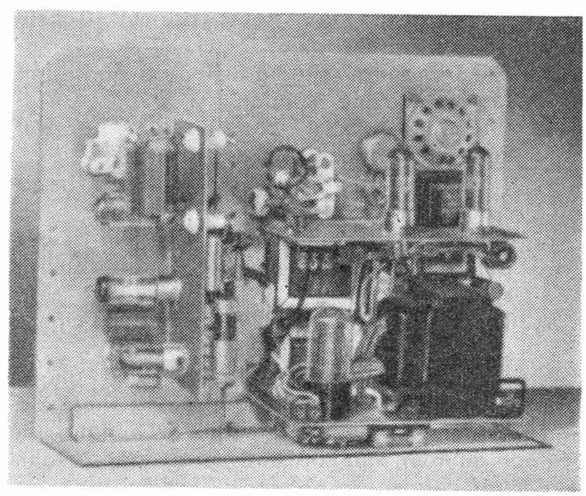
Transmitter

The transmitter circuit is shown in Figure 2. The variable-frequency-oscillator stage uses a 6AU6 in a Clapp circuit with electron-coupled output. The oscillator is tuned by the "bandset" capacitors C_{13} and C_{14}, and the "bandspread" capacitor C_{15}. The combination of C_{20} and L_4 in the plate circuit of the 6AU6 is tuned to the centre of the cw portion of the 40-metre band, and covers this portion of the band without retuning.

An OA2 voltage-regulator tube is used to provide constant voltage for grid No. 2 of the 6AU6, which is the "plate" of the oscillator.

Because the 5763 is a single-ended type and is operated as a "straight-through" rf amplifier, the output stage is neutralized to minimize any tendency to self-oscillation. The neutralizing circuit is extremely simple and requires no adjustments. All that is necessary is the connection shown in Figure 2 between the bottom of L_4 and pin 2 of the 5763 socket. The capacitance between pin 2, which has no internal connection, and the plate pin (pin 1) provides a feedback voltage of the proper phase and amplitude for neutralization.

The output-tank circuit of the amplifier (C_{24}, L_5, and C_{25}) is a simplified pi network designed to provide proper loading for the 5763 when connected to the 72-ohm feeder for the 40-metre doublet antenna.



The oscillator-grid and amplifier-plate coils are mounted at opposite ends of the transmitter "deck" with their axes at right angles. The oscillator-plate coil is mounted below the "deck".

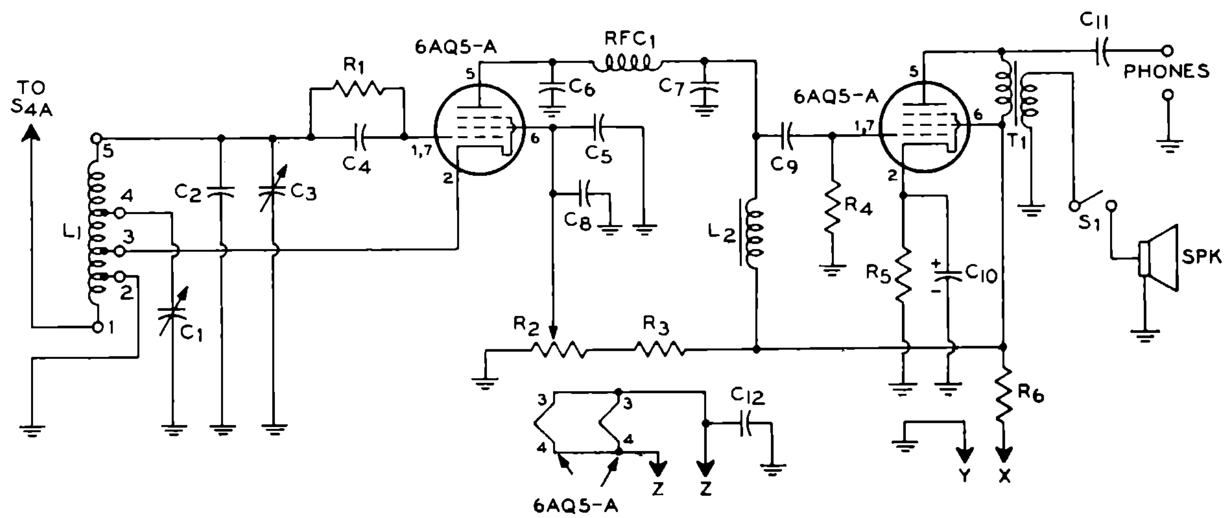


Figure 2: Transmitter circuit.

Switch S_3 is a momentary-contact push-button type which, when depressed, applies plate and screen-grid voltage to the 6AU6, permitting the oscillator to be "spotted" to the received frequency.

An NE-2, 1/25-watt neon lamp, is used as a tuning and keying monitor.* The leads of this lamp are soldered to the stator of C_{25}. The lamp is mounted so that its tip protrudes through a small hole in the front panel directly below the tuning knob for C_{25}.

The oscillator tank coil L_3 is a 37-turn inductor, air-spaced with one inch internal diameter. L_4 is 23 turns of No. 20 enameled wire wound on a $\frac{1}{2}$-inch diameter iron-core form. L_5 is 28 turns of No. 20 enameled wire, $1\frac{1}{4}$ inches in diameter and $1\frac{1}{2}$ inches long.

Switch S_4 is the transmit-receive switch, and applies high voltage and the antenna lead-in to either the transmitter or receiver. Switch S_5 is used to remove heater voltage from the transmitter valves during long stand-by periods.

* Almost any low-power neon lamp would do.

Assembling the Complete Station

The entire station was installed in a portable typewriter case approximately $12\frac{3}{4}$ inches square and $4\frac{1}{4}$ inches deep.

To simplify construction and maintenance, the transmitter, receiver, and common power supply were built on separate "decks" and the front panel was divided into two "operating areas", which can be individually removed. The left-hand area contains the power supply (the heaviest item) and the receiver; the right-hand area the transmitter. The panels and "decks" occupy a space about $9\frac{1}{4}$ inches high. The $3\frac{1}{4}$-inch-high compartment at the bottom of the case is used to store the line-cord, key, and station log.

The front panels are supported by brackets made of $\frac{3}{4}$-inch aluminium angle mounted on three sides of the case. These brackets are recessed about $\frac{1}{8}$ inch so that the front panels are flush with the edges of the case. For additional rigidity, the three "decks" were made $4\frac{1}{8}$ inches deep so that their rear edges rest against the rear of the case. The "decks" containing the receiver and power supply are $6\frac{3}{8}$ inches wide, and the trans-

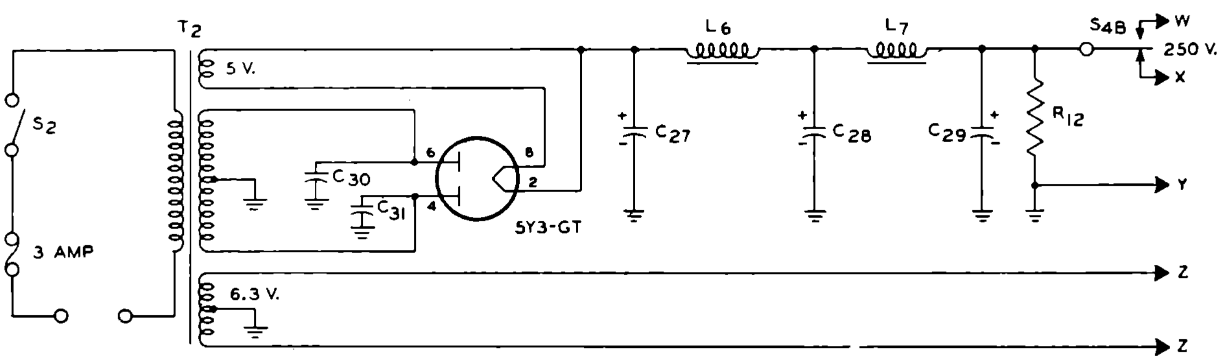


Figure 3: Power-supply circuit.

mitter "deck" is $6\frac{5}{8}$ inches wide. A small bracket supports the power-supply "deck".

To minimize coupling between the oscillator-grid and amplifier-plate coils, these coils are mounted at opposite ends of the transmitter "deck", with their axes at right angles. The oscillator-plate coil is mounted below the "deck".

To assure mechanical stability in the oscillator circuit, L_3 is rigidly mounted and connected by a very short lead to the oscillator-tuning capacitor C_{15}. In addition, the rotor of C_{15} is grounded through a rigid No. 10 copper-wire connection to a ground lug on the "deck" directly below the capacitor.

The only critical point in the receiver is the position of the feedback tap (terminal 2) on L_1. If the receiver does not regenerate smoothly, try moving this tap $\frac{1}{4}$ inch at a time. It will be found easier to solder connections to this coil if the turns on both sides of the tap points are first depressed.

The power supply, shown in Figure 3, requires no special mention.

Station Performance

This 40-metre portable station has met every one of the operating requirements initially established. The best DX achieved to date has been about 500 miles, using a low and hastily erected doublet antenna, and with 225 volts on the plate of the 5763. Thorough workouts, under a variety of operating conditions during the 1958 ARRL Field Day, summer vacations in the country, and several weekend trips, have convinced one ham family that amateur radio and family travel are not necessarily incompatible.

PARTS LIST

C1	100 $\mu\mu$f, variable.
C2	330 $\mu\mu$f, mica.
C3	7-45 $\mu\mu$f, ceramic trimmer.
C4	100 $\mu\mu$f, mica.
C5	0.001 μf, disc ceramic.
C6	0.001 μf, disc ceramic.
C7	0.001 μf, disc ceramic.
C8	1 μf, 400 v.w. paper.
C9	0.02 μf, 400 v.w. paper.
C10	10 μf, 25 v, electrolytic.
C11	1 μf, 400 v.w. paper.
C12	0.001 μf, disc ceramic.
C13	25 $\mu\mu$f, mica.
C14	4.5-25 $\mu\mu$f, ceramic trimmer.
C15	10 $\mu\mu$f, variable.
C16	100 $\mu\mu$f, mica.
C17	270 $\mu\mu$f, mica.
C18	270 $\mu\mu$f, mica.
C19	0.001 μf, ceramic.
C20	100 $\mu\mu$f, mica.
C21	100 $\mu\mu$f, mica.
C22	100 $\mu\mu$f, mica.
C23	0.001 μf, ceramic.
C24	150 $\mu\mu$f, mica.
C25	35 $\mu\mu$f, variable.
C26	0.03 μf, ceramic.
C27	16 μf, 450 v.w. electrolytic.
C28	16 μf, 450 v.w., electrolytic.
C29	16 μf, 450 v.w. electrolytic.
C30	0.001 μf, disc ceramic.
C31	0.001 μf, disc ceramic.
R1	6.8 megohms, 0.5 w.
R2	50,000 ohms, 4 w., w. w.
R3	150,000 ohms, 1 w.
R4	50,000 ohms, 0.5 w.
R5	330 ohms, 1 w.
R6	1,000 ohms, 5 w.
R7	22,000 ohms, 0.5 w.
R8	5,600 ohms, 2 w.
R9	6,800 ohms, 5 w., w. w.
R10	18,000 ohms, 0.5 w.
R11	200 ohms, 0.5 w.
R12	50,000 ohms, 5 w.
L1	See text.
L2	35 h at 15 ma.
L3	See text.
L4	See text.
L5	See text.
L6	16 h at 50 ma.
L7	16 h at 50 ma.
RFC1	2.5 mh.
RFC2	0.5 mh.
RFC3	2.5 mh.
RFC4	2.5 mh.
S1	SPST toggle.
S2	SPST toggle.
S3	SPDT push button.
S4	DPDT toggle.
S5	SPST rotary.
T1	Primary, 5,000 ohms, Secondary to suit speaker.
T2	Secondaries 500 v CT at 70 ma, 5 v at 2 a, 6.3 v at 2.5 a.

With acknowledgements to RCA

RF AMPLIFIERS USING 2N544

This article discusses considerations involved in the design of rf amplifier stages using the 2N544 transistor. The 2N544 is a drift transistor of the germanium p-n-p alloy type designed specifically for use as an rf amplifier at standard broadcast frequencies. The article includes circuits and performance data for the 2N544 in a tuned rf amplifier stage with and without neutralization, and in an unneutralized "broadband" rf amplifier stage. It also gives detailed graphical data on the characteristics of the 2N544 which determine its performance in common-emitter rf amplifier circuits.

General Considerations

The 2N544 performs best as an rf amplifier when used in common-emitter-type circuits. Fig. 1 shows the maximum available gain of the 2N544 in a common-emitter circuit as a function of frequency and dc supply voltage. These gain curves were calculated for an unilateralized circuit having conjugate matched impedances — that is, with source and load impedances equal to the corresponding terminal impedances of the device at all frequencies of interest.

The maximum usable gain of the 2N544 in a practical common-emitter rf amplifier stage is less than the maximum available gain by an amount which depends upon the variations in transistor characteristics and the dc supply voltage used. This reduction in gain is necessary to achieve stability, and its amount is dependent upon whether the stage is neutralized or unneutralized, and upon the maximum spread in transistor characteristics. It may be achieved by impedance mismatching at the input terminals, output terminals, or both, or by the use of resistive damping ("broadband" circuits). Figs. 2, 3, and 4 show

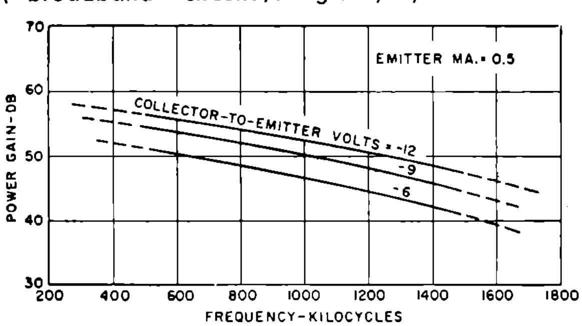


Fig. 1.—Maximum available gain as a function of frequency and collector-to-emitter potential.

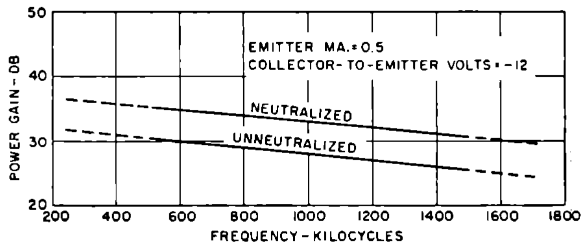


Fig. 2.—Maximum usable gain, neutralized and unneutralized, at a collector-to-emitter potential of −12 volts.

the maximum usable gain of the 2N544 in neutralized and unneutralized common-emitter circuits as a function of frequency and dc voltage.

Transistor Parameters

To design either neutralized or unneutralized common-emitter rf amplifier stages using the 2N544, it is necessary to know the manner in which the terminal parameters of the transistor vary with frequency and dc supply voltage. It is also necessary to know the variation of the intrinsic collector-to-base capacitance $C_{b'c}$ among

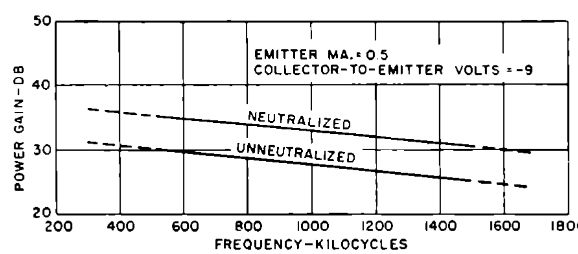


Fig. 3.—Maximum usable gain, neutralized and unneutralized, at a collector-to-emitter potential of −9 volts.

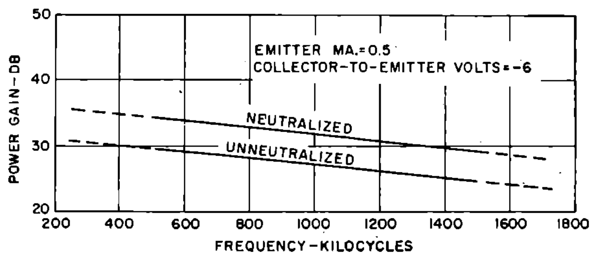


Fig. 4—Maximum usable gain, neutralized and unneutralized, at a collector-to-emitter potential of −6 volts.

individual units, as well as the manner in which this capacitance varies with frequency and dc supply voltage. Figs. 5, 6, and 7 show, respectively, the input parameters, output parameters, and intrinsic transconductance of the 2N544 as functions of frequency and dc supply voltage. Fig. 8 shows the range of values for $C_{b'c}$ and the variation of this capacitance with dc supply voltage over the frequency range 0.5 to 1.5 megacycles per second.

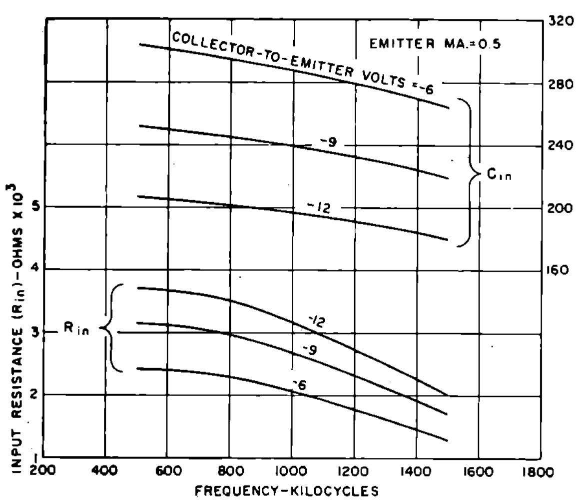


Fig. 5.—Input capacitance (C_{in}) and input resistance (R_{in}) as functions of frequency and collector-to-emitter potential.

Circuit-Design Considerations

Neutralized tuned rf amplifier stages, in comparison with unneutralized stages, provide higher power gains and signal-to-noise ratios, and in superheterodyne receivers, more satisfactory rejection of image and other spurious responses. In "broadband"-type circuits neutralization also provides better discrimination against adjacent-channel interference. Cost considerations, however, frequently dictate the use of unneutralized tuned or untuned ("broadband") rf amplifier

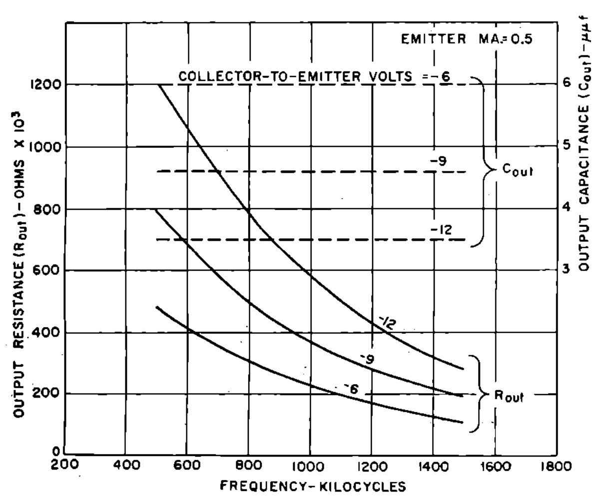


Fig. 6.—Output capacitance (C_{out}) and output resistance (R_{out}) as functions of frequency and collector-to-emitter potential.

stages, especially in portable receivers of the "personal" type. In addition to low cost, the principal advantage of "broadband" amplifiers is their small space requirements. A broadband amplifier can be made to have the desired bandwidth—for example over the broadcast band—either by the proper choice of shunt resistance or by incorporation of equivalent losses in the output transformer. Substantially uniform response over the entire pass-band can be obtained if the amplifier is peaked at the high-frequency end of the band to offset the reduction in gain resulting from the decreased input impedance of the following stages at the higher frequencies. In such cases the required peaking capacitance can be supplied entirely by the output capacitance of the rf amplifier stage and the input capacitance of the following stage.

Fig. 9 is the circuit of a tuned rf amplifier stage using the 2N544. Performance data for this circuit with and without neutralization are given in Table I. Fig. 10 is the circuit of an unneutralized "broadband" rf amplifier stage. Performance data for this circuit are given in Table II.

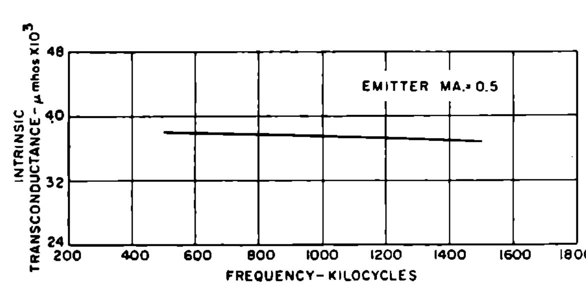


Fig. 7.—Intrinsic transconductance as a function of frequency.

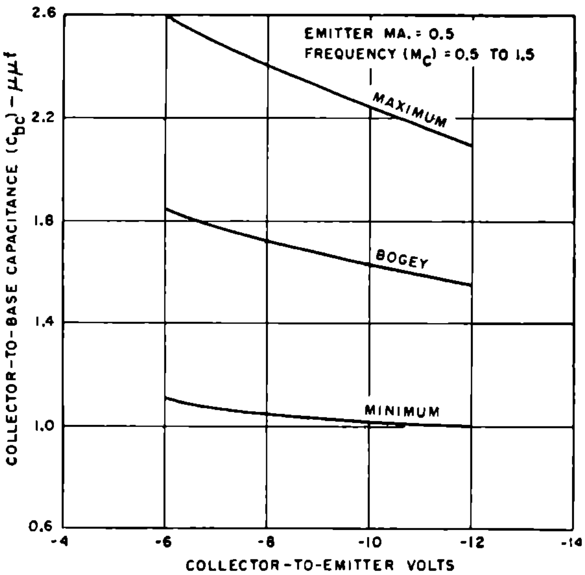


Fig. 8. — Range of intrinsic collector - to - base capacitance ($C_{b'c}$) and variation of $C_{b'c}$ with collector-to-emitter potential.

TABLE I

Frequency (Kc)	Power Gain (db) Neutralized	Unneutralized
600	30	26
1000	32	27
1500	30	26

Table I.—Gain characteristics of the tuned rf amplifier shown in Fig. 9.

TABLE II

Frequency (Kc)	Power Gain (db)
600	18.5
1000	22.0
1500	20.0

Table II.—Gain characteristics of the "broad-band" rf amplifier stage shown in Fig. 10.

PARTS LIST FOR THE CIRCUIT OF FIG. 9

C1, C6—Tuning capacitor, 22 to 185 μμf, air variable
C2, C7—Trimmer capacitor, 2 to 20 μμf, mica variable
C3, C4, C8—0.05 μf, 50 volts, ceramic
C5—Neutralizing capacitor, 7 μμf, 500 volts, mica
R1—39000 ohms, 0.5 watt
R2—3900 ohms, 0.5 watt
R3, R4—470 ohms, 0.5 watt
T1—Antenna transformer; wound on 3/4-inch by 1/8-inch ferrite core, 4 inches long; primary—86 turns No. 3/41 Litz wire; secondary—4 turns No. 3/41 Litz wire; Q at 1 Mc; loaded—100; unloaded—370.
T2—Interstage transformer

Characteristics	Neutralized	Unneutralized
Primary tuned resistance at 1 Mc—ohms	10200	3630
Primary input resistance at 1 Mc—ohms	9340	3590
Turns ratio—Terminals 1 and 3 to terminals 1 and 2	3.87	6.48
Turns ratio—Terminals 1 and 2 to terminals 4 and 5	3.32	2.12
Unloaded Q at 1 Mc (transformer mounted in chassis)	52	52
Loaded Q at 1 Mc (transformer mounted in chassis)	23.8	25.6

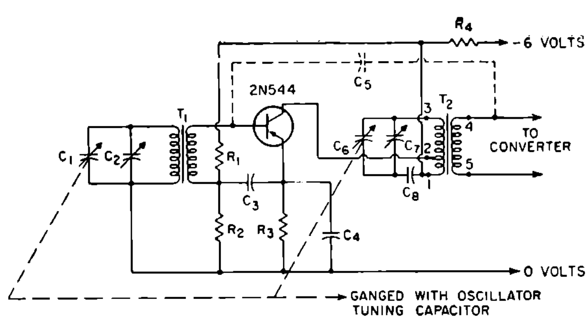


Fig. 9.—Circuit of a tuned rf amplifier stage with and without neutralization.

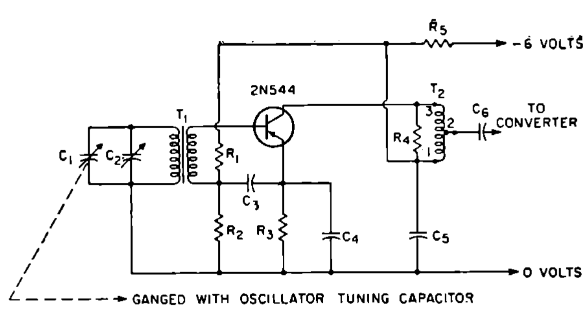


Fig. 10.—Circuit of an unneutralized "broad-band" rf amplifier stage.

PARTS LIST FOR THE CIRCUIT OF FIG. 10

C1—Tuning capacitor, 22 to 185 $\mu\mu f$

C2—Trimmer capacitor, 2 to 20 $\mu\mu f$, mica

C3, C4, C5, C6—0.05 μf, 50 volts, ceramic

R1—39000 ohms, 0.5 watt

R2—3900 ohms, 0.5 watt

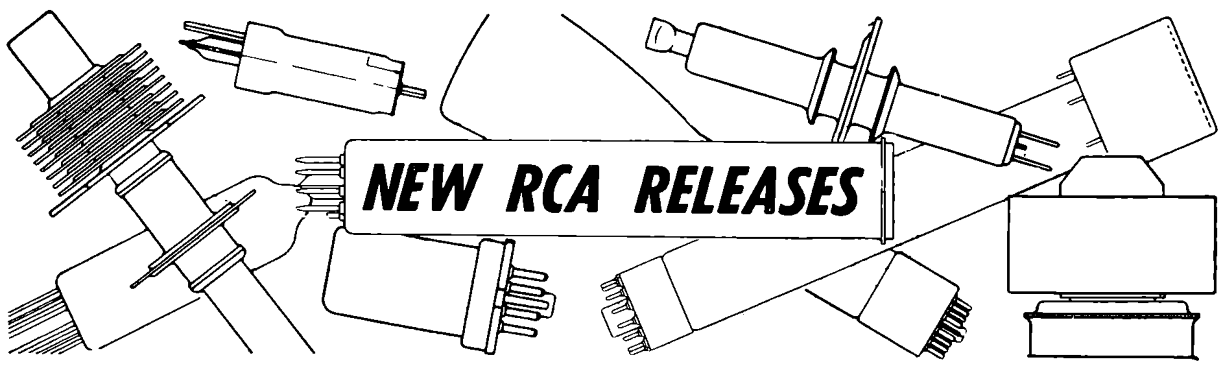
R3, R5—470 ohms, 0.5 watt

R4—2200 ohms, 0.5 watt

T1—Antenna transformer; wound on 3/4-inch by 1/8-inch ferrite core, 4 inches long; primary—110 turns No. 3/41 Litz wire; secondary—6 turns No. 3/41 Litz wire; Q at 1 Mc; loaded—170; unloaded—105.

T2—Interstage transformer; primary tuned resistance at 1 Mc, 37800 ohms; primary input resistance at 1 Mc (secondary terminated in input resistance of 800 ohms), 4200 ohms; turns ratio (terminals 1 and 3 to terminals 1 and 2), 2.12; Q at 1 Mc (transformer mounted in chassis); unloaded—18; loaded—2.

(With acknowledgements to RCA)



2N561, 2N1014

The 2N561 and 2N1014 are high-voltage, high-current power transistors intended for use in heavy-duty industrial and military applications including power-switching circuits, dc-to-dc converter circuits, voltage-regulator circuits, power-supply circuits, and relay-actuating circuits. They may also be used in af oscillator service, and in large-signal class A or in class B push-pull af amplifier service.

The 2N561 and 2N1014 are alloy-junction transistors of the germanium p-n-p type. They feature a maximum peak-collector-current rating of —10 amperes. The 2N561 has a maximum peak collector-to-base voltage rating of —80 volts, and a maximum peak collector-to-emitter voltage rating of —50 volts with base open. The 2N1014 is like the 2N561 except that it has higher maximum peak collector-to-base (—100 volts) and collector-to-emitter (—65 volts) voltage ratings for use particularly in those industrial and military applications requiring such high voltages.

In a typical common-emitter type of "on-off" power-switching circuit having a dc supply voltage of 28 volts and a driving power of on 42.6 milliwatts, these transistors can provide a power output of 54 watts with a power gain of 31.3 db. In a typical dc-to-dc converter circuit having a dc supply voltage of 28 volts and a dc supply current of 4.2 amperes, they can provide a dc output voltage of 420 volts at an efficiency of 88 per cent.

The design of the 2N561 and 2N1014 utilizes a special mount structure in which the collector is electrically and thermally connected to a mounting flange. This mounting arrangement provides for good electrical contact and excellent transfer of heat from the transistor junctions to the heat sink.

2N647

The 2N647 is a new alloy-junction transistor of the germanium n-p-n type. It is intended especially for use along with its p-n-p counterpart, the 2N217, in class B complementary-symmetry power output stages of compact, transformerless battery-operated portable radio receivers and phonographs. It is also suitable for audio amplifiers in which compactness, good frequency response, and relatively high power output at low cost are important design considerations. This transistor may also be used in conventional class B push-pull and in class A audio amplifier circuits.

In a typical class B complementary-symmetry circuit, a 2N647 (n-p-n type) and a 2N217 (p-n-p type) used together in the output stage and driven by a 2N217 as a class A driver are capable of providing a power output of approximately 100 milliwatts at a power gain of 54 db.

In a typical push-pull circuit, two 2N647's used in the output stage and driven by another 2N647 as a class A driver are capable of providing a power output of approximately 100 milliwatts at a power gain of 66 db.

The 2N647 has a large-signal dc-current transfer ratio essentially constant over the operating current range to insure circuit linearity, a collector cutoff current of only 14 microamperes to insure stable performance under varying ambient temperature, and excellent uniformity of characteristics to provide unit-to-unit interchangeability.

RADIOTRON 6DE4

The 6DE4 is a new, high-performance damper diode on an octal base, especially designed to provide improved performance and increased reliability in 110-degree TV deflection circuits. It is rated to withstand a maximum peak inverse plate voltage of 5000 volts and can supply a maximum peak plate current of 1100 milliamperes and a maximum dc plate current of 175 milliamperes.

Design features of the 6DE4 which make possible outstanding performance in damper circuits include:—

(a) Cathode having special low-impedance coating to withstand the high voltage pulses encountered in TV damper service, minimize sputtering, and assure uniform emission over the entire cathode surface.

(b) Specially designed micas together with a plate structure having rounded contours to reduce voltage gradients and insure against high-voltage breakdown.

(c) Slots in plate structure surrounding the cathode to minimize the danger of sustained arcs.

RADIOTRON 6DK6

The 6DK6 is a sharp-cutoff pentode of the 7-pin miniature type intended for use as an if amplifier in TV receivers. It features high transconductance (9800 micromhos) at low plate and grid-No. 2 voltages, and low interelectrode capacitances. These features contribute to high gain per stage and are of particular importance in the design of TV receivers in which only two if stages are used.

RADIOTRON 6EB8

The Radiotron 6EB8 is a high-mu triode—sharp-cutoff pentode of the 9-pin miniature type intended for use in video-output amplifier, sync-clipper and phase-inverter circuits of TV receivers.

The pentode unit of this valve features high transconductance (12500 micromhos) combined with a plate dissipation rating of 5 watts, and is particularly useful in video-output amplifier circuits. The triode unit has an amplification factor of 100, is especially useful in voltage-amplifier applications such as in sync-separator, sync-clipper and phase-inverter circuits.

RADIOTRON 7203/4CX250B AND 7204/4X250F

The new Radiotron 7203/4CX250B is a very small and compact forced-air-cooled beam power valve utilizing ceramic-metal-seal construction throughout. Capable of dissipating 250 watts in its plate, this new type is useful as an af power amplifier and modulator, as a wide-band amplifier in video applications, as a linear rf power amplifier in single - sideband suppressed - carrier equipment, and as a class C amplifier and oscillator. It can be operated with full ratings at frequencies up to 500 Mc.

The 7204/4X250F is the same as the 7203/4CX250B except for its heater rating. The 7204 has a 26.5-volt, 0.58-ampere heater as compared with a 6-volt, 2.6-ampere heater in the 7203. The ceramic-metal-seal construction employed throughout in these types can withstand operation at higher temperatures than a glass-seal construction and thus provides improved reliability. The specially designed, high-efficiency radiator which is brazed directly to the plate for better heat transfer, makes possible the maximum plate-dissipation rating of 250 watts with no sacrifice in valve reliability.

RADIOTRON 7213

The Radiotron 7213 is a new forced-air-cooled, uhf beam power valve with ceramic-metal seals, designed for use as a linear rf power amplifier and as a class C rf power amplifier in airborne and fixed-station equipment. Small in size for its power capability, the 7213 has a maximum plate-dissipation rating of 1500 watts and can be used with full ratings at frequencies up through the Aeronautical Radio-Navigation Band of 960 to 1215 Mc.

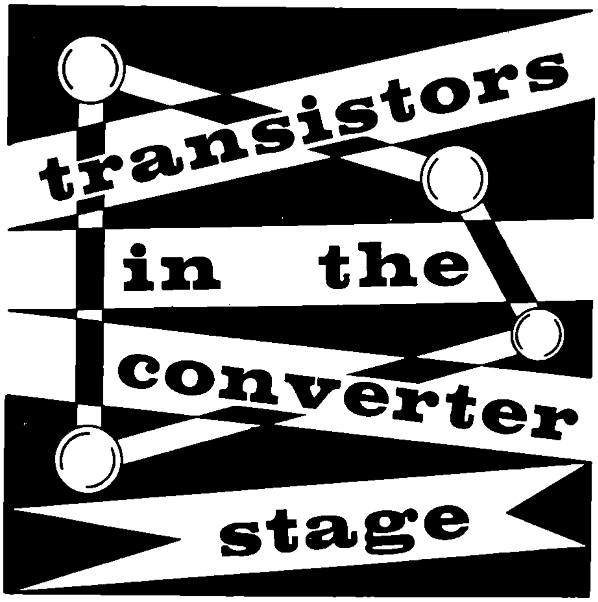
When used under CCS conditions as an rf power amplifier in class C telegraphy service, the 7213 has a maximum plate-voltage rating of 2500 volts and a maximum plate-input rating of 2500 watts. Under these conditions in a grid-drive circuit, the 7213 is capable of delivering useful power output of 1350 watts with a power gain of 20 at 600 Mc. As a linear rf power amplifier in single-sideband suppressed-carrier service, the 7213 is capable of providing a maximum-signal power output (CCS) of 1250 watts.

RADIOTRON 7262

Designed for use in small, compact, transistorized TV cameras, the vidicon camera tube type 7262 has an overall length of only $5\frac{1}{8}''$, and employs a low-power heater requiring only 0.6 watt. It can produce pictures of broadcast quality with as little as one footcandle of highlight illumination on its faceplate.

Newly developed photoconductor processes have been used in the 7262 to provide a high degree of uniformity of characteristics from tube to tube. The photoconductive surface has uniform thickness permitting uniform sensitivity and dark

(Continued on page 185)



This article describes a broadcast-band frequency-converter circuit which can use interchangeably transistors types 2N140, 2N219, 2N411, or 2N412. The circuit has an average gain of 30 db when operated from a 9-volt battery, and operates dependably at reduced battery voltage. Within the operating conditions of the circuit described, the performance of all four transistor types is the same. Types 2N140 and 2N411 have the linotetrar 3-pin base; types 2N219 and 2N412 have flexible leads, and are AWV Preferred Types.

Circuit Description

The basic circuit of the converter is shown in Fig. 1. The circuit uses a series-feed oscillator, with emitter injection to minimize interaction between the oscillator and antenna sections, and requires a five-terminal oscillator coil. Capacitor C_e is used to couple the signal from the oscillator coil to the emitter, and also as an rf bypass for the emitter-circuit resistor R_e. The considerations which determine the values for the various components, and the design of the oscillator coil and the antenna section are discussed below.

Bias Network

The resistors which comprise the bias network (R_1, R_2, and R_e) are chosen to provide a constant emitter current of 0.5 milliampere and to assure satisfactory operation at low battery voltages. Maximum operating-point stability and, therefore, the highest degree of transistor interchangeability are obtained with large values of R_e and small values of R_1 and R_2. R_e, however, reduces the available collector voltage, so that the use of extremely large values for this resistor can impair the operation of the circuit at low battery voltages; R_1 and R_2 shunt the input circuit and, therefore, should be large to minimize loss of input-signal energy. Consequently, the values of these resistors must be very carefully chosen. Values which have proved satisfactory are: 9-volt supply—R_1 = 27,000 ohms, R_2 = 6,800 ohms, R_e = 3,300 ohms; 6-volt supply— R_1 = 22,000 ohms, R_2 = 6,800 ohms, R_e = 2,700 ohms.

Capacitors

The major consideration in the selection of a tuning capacitor for a portable transistor receiver is usually size. The small two-section variable capacitors generally available for this application, however, create two serious problems for the set designer:

(1) The small maximum capacitances of the two sections require that the primary windings of the antenna and oscillator coil have very high inductance. When the antenna is small because of cabinet dimensions, it may be difficult to obtain the necessary inductance on the antenna core.

(2) When the rotors are fully open the stray capacitance between the two stators is comparable to the residual capacitances of the individual sections. This condition results in undesirable coupling between the oscillator and antenna circuits, and, therefore, in instability at the high end of the tuning range. It can, however, be avoided by the use of a grounded shield plate between the two stators.

The values of capacitors C_a and C_e are chosen to minimize "squegging". C_a should be no larger than necessary to provide adequate bypassing for R_2 at the intermediate frequency; C_e should have the smallest value that will provide adequate injection at the low-frequency end of the tuning range. Values of 0.02 microfarad for C_a and 0.01 microfarad for C_e have proved satisfactory from both standpoints in several experimental circuits.

Oscillator Coil

The oscillator coil is wound on a high-permeability, high-Q ferrite core. The tuned or primary winding should have the highest possible Q to minimize changes in oscillator frequency with changes in temperature and battery voltage.

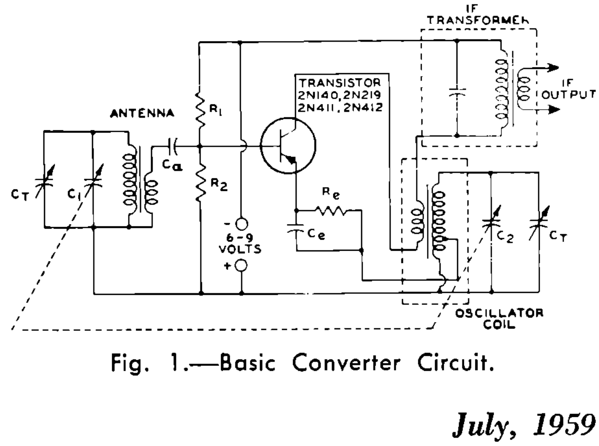


Fig. 1.—Basic Converter Circuit.

The collector winding should have as few turns as possible to minimize loading of the primary, and the turns ratio between the collector winding and the emitter portion of the primary should provide sufficient feedback to assure oscillation over the entire band at reduced battery voltage with any 2N140, 2N219, 2N411, or 2N412. To assure maximum coupling between the collector winding and the emitter portion of the primary, these windings should be placed next to the core. A 10-turn collector winding and a universal-wound high-Q primary of No. 7/41 Litz wire having a 4-turn emitter portion will provide excellent injection over the band.

Antenna

The amount of signal energy picked up by a ferrite rod and coil antenna is proportional to the volume of the antenna. Since the volume of an antenna for a transistor receiver is generally limited by cabinet-size requirements, the sensitivity that can be achieved depends largely upon the quality of the antenna and upon its position in the cabinet. For maximum sensitivity the core of the antenna should be made of a high-permeability, high-Q ferrite, and should have the largest possible diameter. The tuned winding or primary, should have the highest possible Q, and should be of No. 5/41 or larger Litz wire space wound. The secondary should be wound at the ground end of the primary, and should match the impedance of the tuned primary to the 800-ohm input impedance of the converter circuit at a frequency of 1 megacycle. Because the relative positions and phasing of the primary and secondary windings affect the overall sensitivity of the receiver at the ends of the response band, experiment may be necessary to determine the best arrangement for a given set design[1].

IF Transformer

Although if transformers are generally designed with the stability of the if stages in mind[2], the transformer between the converter and first if amplifier should be designed to extract maximum if energy from the converter. To assure maximum extraction the impedance presented to the converter should be equal to the converter output impedance—that is, 70,000 ohms. This transformer should also be double-tuned so that it presents the lowest possible impedance to the oscillator current throughout the tuning range. For economy, however, a single-tuned transformer may be used with acceptable results.

References;

[1]W. J. Polydoroff, "Magnetic-Field Antenna," Electronic Industries, Volume 17, Number 3, March, 1958.

[2]D. D. Holmes and J. O. Stanley, "Stability Considerations in Transistor IF Amplifiers", Transistors I, RCA Laboratories, Princeton, N. J., 1956.

(With acknowledgements to RCA)

RADIOTRONICS VOL. 22, No. 10, OCTOBER 1957

CORRECTION

We have been asked by Mr. H. R. Wilshire to publish the following corrections to his article entitled "Transistors and Their Parameters".

Page 150, Table 5, the expression for New Symbol h_{rb} under h system Common Collector now reads $\frac{h_{ic}h_{oc}}{-h_{fc}}$. It should read $h_{rc} + \frac{h_{ic}h_{oc}}{-h_{fe}} - 1$.

Page 151, Table 5 continued, the expression for New Symbol h_{rc} under Hybrid π Common Emitter now reads $1 - \frac{r_{b'c}}{r_{b'e}}$. It should read $1 - \frac{r_{b'e}}{r_{b'c}}$

NEW RCA RELEASES

(Continued from page 183)

current and, hence, high effective sensitivity over the entire scanned area. The design of the 7262 utilizes nonmagnetic parts in the front end, an optically flat faceplate free from optical distortion, and an envelope without a side tip.

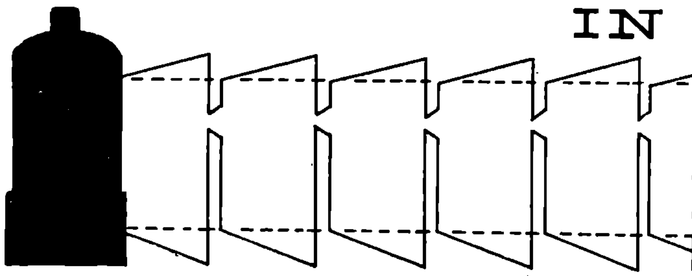
RADIOTRON 7263

The new short vidicon type 7263 is designed to provide reliable performance in black-and-white or colour TV cameras under severe environmental conditions. It has been developed to meet military requirements for a TV camera tube capable of providing high-quality pictures under severe environmental conditions involving shock, vibration, humidity, and altitude. This new tube is tested under environmental conditions according to the techniques of military specifications MIL-E-5272B and MIL-E-5400.

The internal structure of the 7263 is reinforced against shock and vibration by multiple bulb supports and glass beading that locks electrodes in permanent relation to each other. Because of its short length of only $5\frac{1}{8}$ inches, its low-power heater requiring only 0.6 watt (1/3 less than that of any other commercial vidicon), and its fast cathode warm-up time, the 7263 is well suited for use in small, compact transistorized TV cameras.

The 7263 utilizes a photoconductive surface having high effective sensitivity over the entire scanned area and can produce pictures of broadcast quality with as little as one foot-candle of highlight illumination on its faceplate. Other features of the 7263 include an envelope without a side tip, resolution capability of about 600 television lines, and freedom of operation in any position.

VALVE REQUIREMENTS IN Vertical Deflection Circuits



by KARL W. ANGEL

Electron-Tube Division, RCA

This paper discusses linearity and efficiency problems which must be considered in the selection of an output valve for a vertical-deflection power amplifier. Because the purpose of the vertical-deflection power amplifier is to supply a given peak-to-peak current to the vertical yoke winding, the design problem is essentially one of matching the resistive component of the yoke to the valve characteristics in much the same manner as in audio-power amplifier design.

Trace-Period Equivalent Circuit

Fig. 1 shows a simple vertical-deflection circuit and its lumped-parameter equivalent circuit for the trace portion of the deflection cycle. In it, L_1 is the open-circuit transformer inductance, L_o is the yoke inductance plus leakage inductance, and R_o is the yoke resistance plus transformer winding resistance. The yoke current, i_o, is assumed to have a constant rate of change, as shown in Fig. 2a, which will produce a linearly changing magnetic field in the deflecting yoke.[1] The waveforms[2] for the plate voltage (e_p), current flowing in the equivalent shunt inductance (i_l), and plate current ($i_p = i_o + i_l$) in the equivalent circuit for trace are shown in Figs. 2b, 2c, and 2d.

For maximum efficiency[3], the average plate current, $\bar{I}_b$, should be a minimum, $\frac{R_o}{L_o + L_i} = 120$, or

$$\left.\frac{di_p}{dt}\right|_{t=0} = 0$$

If $L_o << L_l$, the average plate current for maximum plate efficiency is 1/3 the peak-to-peak plate current ($I_{p\text{-}p}$), and the maximum plate efficiency of an ideal pentode is 50 per cent.[3]

In practice, the value of dc plate current can be made slightly less than 1/3 the peak-to-peak plate current because class AB operation is used to provide linearity control. Also, the practical efficiency of a pentode circuit is about 40 per cent. because of the 25 per cent. increase in power input necessitated by valve drop. For triode circuits, the practical upper limit of efficiency is about 33 per cent. The transformer efficiency is simply the

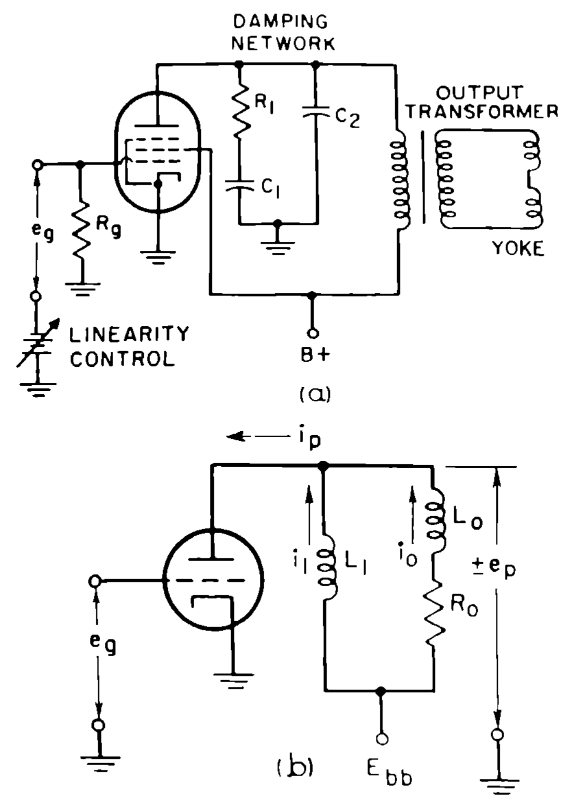


Fig. 1. Vertical-Deflection-Output Circuit and Its Equivalent Circuit During Trace.

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1 For wide-angle deflection of "flat-faced" picture tubes, the yoke current should be slightly "S"-shaped rather than linear.

2 Refer to appendix for mathematical expressions for e_p, i_l, and i_p, and for derivation of expressions for maximum efficiency.

3 Refer to appendix for mathematical expressions for e_p, i_l, and i_p, and for derivation of expressions for maximum efficiency.

ratio of reflected yoke resistance to reflected yoke resistance plus winding resistance and is usually 85 to 90 per cent.

Reactive Power in L_1

In addition to the real power input during trace, the portion of the energy stored in L_1 during trace and dissipated in the damping circuit during retrace must be supplied.[4] In practice, the reactive power is only about 3.6 per cent. of the output power and can be neglected. The peak-to-peak power capability of the amplifier must be increased by about 10 per cent. over the theoretical minimum (for a typical L_o to L_i ratio of 1 to 10) because of the additional energy stored in L_1.

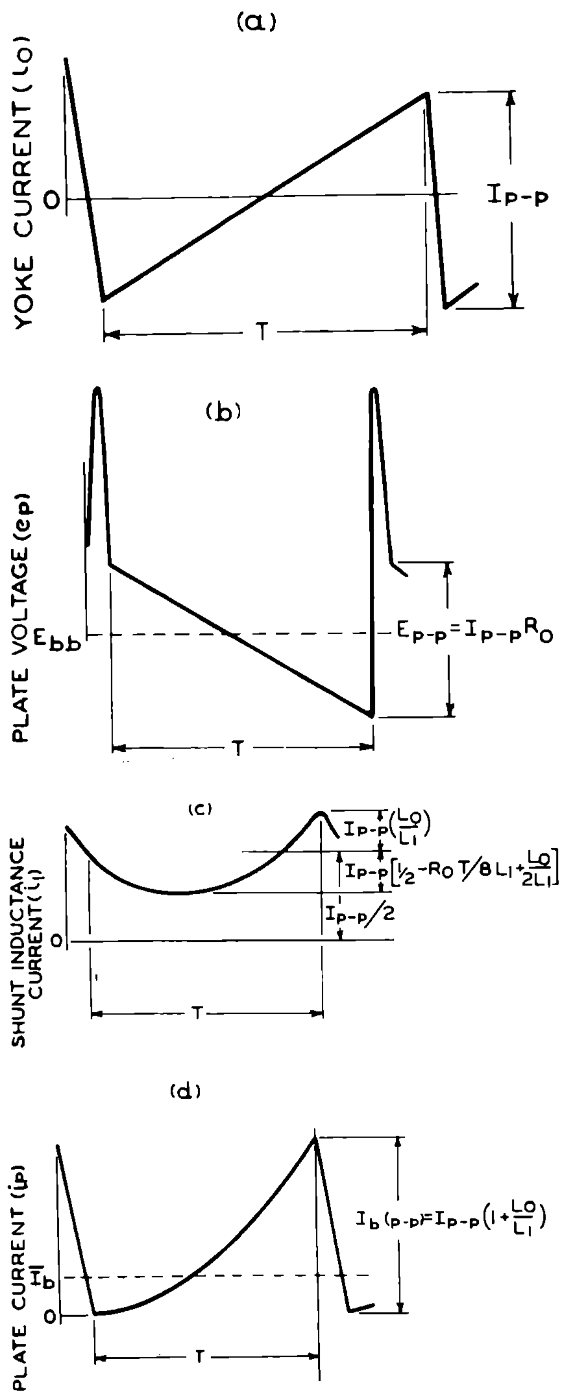

Fig. 2. Vertical-Deflection Current and Voltage Waveforms During Trace.

Yoke Sensitivity and Efficiency

The sensitivity and efficiency of the yoke depend on many factors which need not be considered here. However, it is convenient to examine some of the more straightforward relationships in yoke design.

The amount of deflection caused by the yoke field is proportional to the number of turns and the peak-to-peak current. For a given power input and a given ratio of inductance to resistance in the yoke, the sensitivity[5] is constant for various values of yoke resistance.

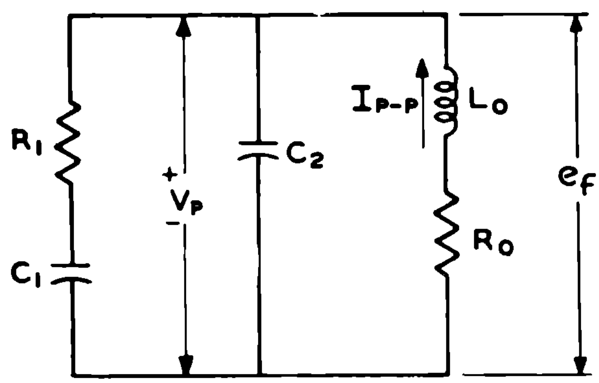

Fig. 3. Vertical-Deflection Equivalent Circuit During Retrace.

For example, let the ratio of inductance to resistance (L/R) equal 1 for two different yokes. Then if the cross-sectional area is the same for both yokes, and if the resistance of yoke B is twice that of yoke A, the current in yoke B can be reduced by $\sqrt{2}$ for the same power sensitivity as yoke B. The efficiency of a yoke, therefore, is measured by the ratio of L to R and is not affected by the absolute value of yoke resistance. The higher this ratio; the more the yoke. Increasing the number of turns to give a higher inductance will require more copper unless the cross-sectional area of the yoke is held constant, in which case the resistance will also increase directly with the square of the turns and the ratio of L to R will be unchanged.

Retrace

After the trace period, the plate current must be brought quickly to zero to prepare the circuit for another trace period. This reduction in current Is effected by driving the valve rapidly into cut-off. The sudden stopping of plate current when

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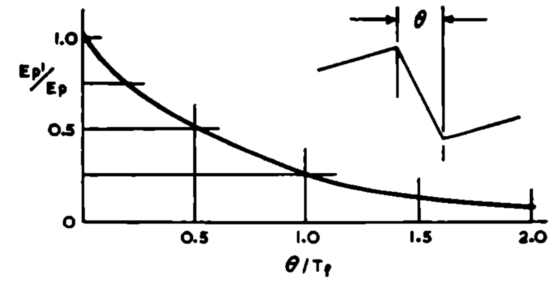
[4] The expressions for Vertical Deflection Power relationships are given in the appendix.

[5] Sensitivity is defined here as the amount of deflection of the electron beam per unit driving power.



Ep'/Ep RATIO OF PLATE PULSE TO PLATE PULSE PRODUCED BY A UNIT STEP
θ/Tf RATIO OF FLYBACK TIME TO FLYBACK TIME REQUIRED BY UNIT STEP

Fig. 4. Flyback Pulse as a Function of Cutoff Rate.

the magnetic energy state is high results in a rapidly collapsing field which generates a large positive plate voltage pulse. The equivalent circuit for retrace is as shown in Fig. 3 if the relatively small current flowing in the equivalent shunt inductance (L_1) is neglected. The stored energy should be dissipated as quickly as possible to obtain rapid retrace. This action is by definition critical damping and is accomplished by the addition of the damping network, R_1, C_1. The capacitance C_2 is the total stray, wiring, and added capacitance, and its value is determined by the pulse rating of the valve plate and the desired flyback time. The capacitance C_1 is added for dc blocking, and its effect is small for values of $C_1 >> C_2$ if critical damping is assumed.

An approximate expression for the flyback voltage (e_f peak) for critical damping in a pentode circuit is:

$$e_f \text{ peak} = \frac{I_{p-p}}{C_2 \omega_r \epsilon} = \frac{I_{p-p}\omega_r L_o}{\epsilon}$$

where e_f peak is in volts, I_{p-p} is in amperes, C_2 in farads, and ω_r is the angular velocity equal to 2π times the resonant frequency of the series circuit L_o, C_2 (Fig. 3) in radians per second. The initial current at the start of retrace ($I_{p-p}/2$) is flowing through L_o. For simplification, the relatively small value of R_o and the initial voltage on C_2 (V_p) are assumed to be zero. The expression given above is the maximum value of e_f peak which occurs at $t = \dfrac{1}{\omega_r}$. In practice, e_f peak will be somewhat less than this value depending on the cutoff rate of the valve. A typical example is a 50-microsecond cutoff time leading to a 20 per cent. reduction in the plate as shown in Fig. 4. All values of plate pulse voltage are above B+.

Design of Damping Networks

The damping network may be calculated by the relationship $R_1 = \sqrt{L_o/4C_2}$ as defined by critical damping. For rapid retrace, capacitor C_2 should be chosen so that the time constant $1/\omega_r$ is as small as possible with e_f peak within the plate pulse rating of the valve and output transformer. The value of C_1 should be about twenty times larger than C_2 if its effect is to be kept negligible. If C_1 is made smaller than 20 times C_2 it will reduce the damping current in R_1, thereby delaying the plate pulse and making the circuit oscillatory unless the value of R_1 is adjusted for the given value of C_1.

An expression for the relationship of R_o, R_1, C_1, and C_2 for critical damping is difficult to determine in the general case so three practical examples, Fig. 5a, have been worked out. These three cases are equivalent for all practical purposes, but in each case different values of R_1 were used to obtain critical damping. In Case III, R_1 is 14 per cent. larger than in Case II in order to compensate for the effect of C_1 which is 0.022 microfarad rather than infinite as in Case II. Also, the peak plate pulse in Case III is slightly larger than in Case II. Case I shows the effect of a finite value of R_o and an infinite value of C_1 with all initial conditions considered. The voltage waveforms for these cases, as shown in Fig. 5b, can be seen to be nearly equal. Retrace must be completed in 665 microseconds (4 per cent. of field period); however, in feedback oscillator circuits in which the plate pulse is used to key the discharge tube, the retrace period should be 250 microseconds ± 20 per cent. for good interlace.*

In triode output circuits the tube acts as the damping resistance. The damping, then, is adjusted by varying a peaking resistor in series with

* The figures quoted here apply of course to the 525-line system in use in U.S.A.

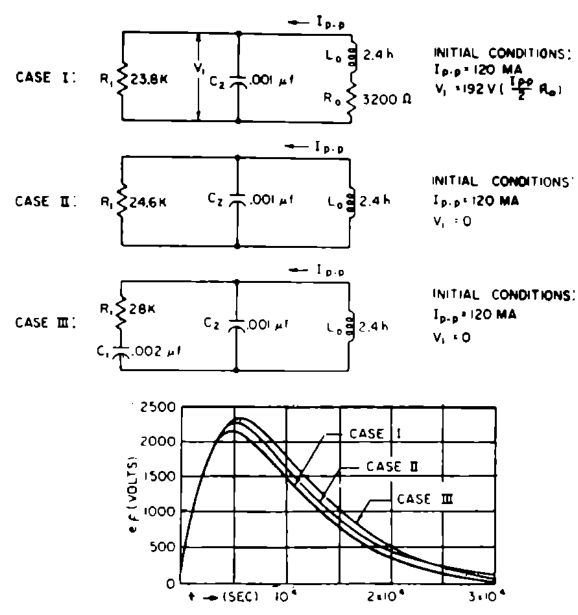


Fig. 5. Examples of Critical Damping in Vertical-Deflection Equivalent Circuits During Retrace and the Plate-Voltage Waveforms for Each Example.

the charge capacitor of the sawtooth generator. This value should be approximately 10,000 ohms with the exact value determined empirically to be just sufficient to allow completion of retrace at the instant peaking pulse is completed. The equivalent circuit for the sawtooth generator and peaking resistor is shown in Fig. 6. The expression for the flyback pulse for a triode output stage cannot be readily analyzed. However, if the flyback pulse is assumed to be a square pulse, then the flyback current will be sawtooth. The expression for the minimum theoretical flyback pulse is: $e_{f(min)} = L_o I_{p-p}/T_r$. Where $e_{f(min)}$ is the pulse voltage above B+ in peak volts, L_o is in henries, I_{p-p} in amperes and T_r is the retrace period in seconds.

This value can be closely approached in practice if the duration of the peaking pulse equals the desired retrace period. This period depends on the switching rate of the discharge tube, which in turn, depends on the grid-circuit time constants. The peak value of the triode plate pulse usually is 30 to 40 per cent. less than the plate pulse of a pentode circuit for an equivalent deflection system.

Triode Load Line

If, as shown in Fig. 7, the expression for plate current vs. plate voltage is plotted on the plate family characteristic curves of a typical triode, the required grid waveform can be derived. The resultant non-linear grid-drive waveform approximates the exponential form $1 - \epsilon^{-at}$. The high impedance sawtooth generator used in vertical deflection circuits (Fig. 6) delivers a voltage of this form, but in practice it is most economical to choose a large value for the time constant 1/a so that the sawtooth voltage is nearly linear. Then the design can utilize smaller, less expensive, coupling capacitors and smaller grid resistors which lessen the effects of grid emission, gas, and leakage reverse-grid currents in the output stage.

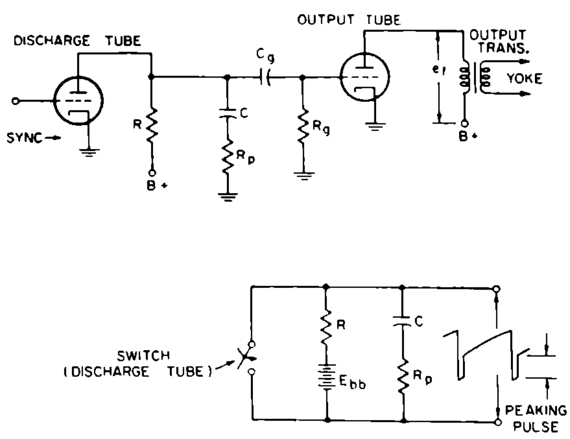


Fig. 6. Triode Vertical-Deflection and Discharge-Tube Circuit and Equivalent Sawtooth Generator Circuit.

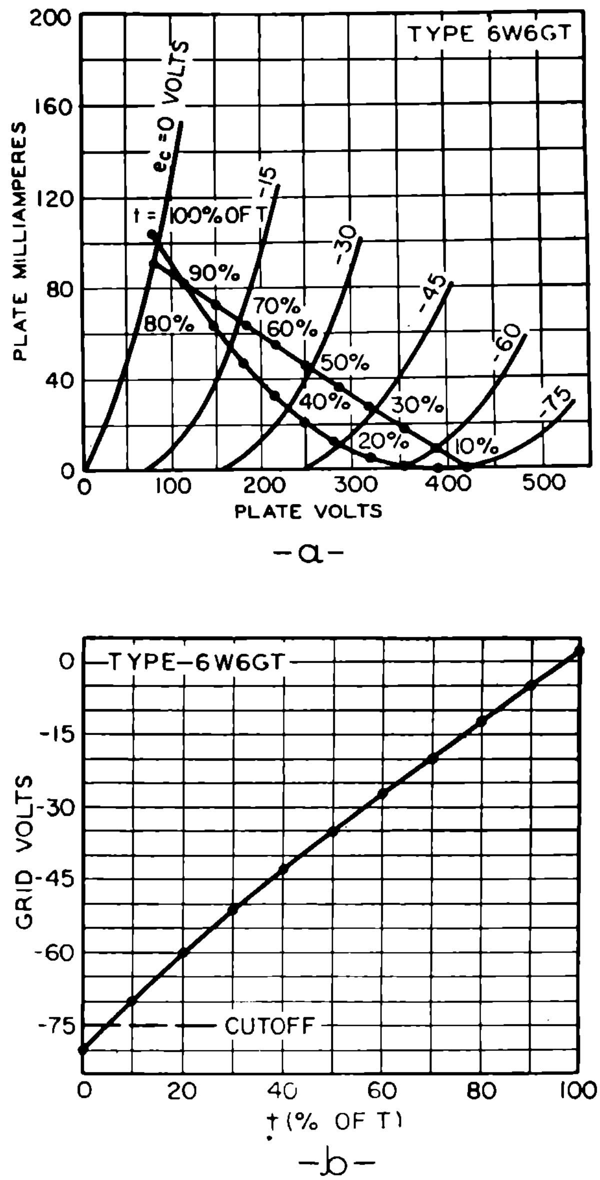


Fig. 7. Load Line for Linear Deflection of a Triode.

The operation of height and linearity controls is identical in principle for both triode and pentode circuits, and is discussed when the pentode example is considered.

Pentode Load Line

The expression for plate current vs. plate voltage is plotted on the plate family characteristic curves of a typical pentode (6CZ5) in Fig. 8a. In the pentode case, the drive must be nearly linear for the last 80 per cent. of the trace period, as shown in Fig. 8b. However, during the first 20 per cent. of trace, a large amount of nonlinearity is required. The voltage shown in Fig. 8c must be added to a sawtooth voltage to drive this typical pentode correctly.

Obtaining the linear drive required by a pentode directly from the high-impedance sawtooth

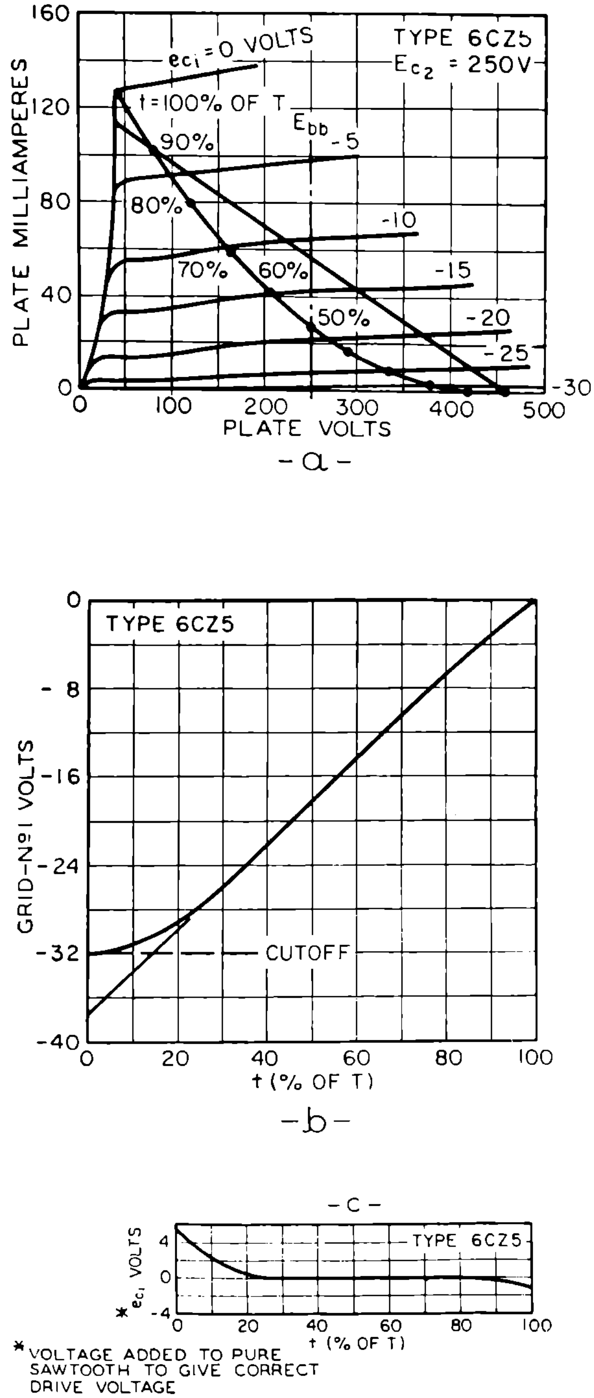


Fig. 8. Load Line for Linear Deflection of a Pentode.

generator is not economical because large coupling capacitors are costly, and because large grid resistors increase the effects of grid emission, grid leakage, and gas reverse-grid currents. When smaller coupling capacitors and grid resistors are used, as in triode circuits, the resulting nonlinearity can be corrected by the addition of an easily obtained linearizing voltage. The linearizing feedback voltage is discussed later.

Stretch at Top of Picture

The grid drive voltage shown in Fig. 8b is extrapolated from the linear portion to cutoff, and is found to cross the time axis after about 15 per cent. of trace. If a sawtooth drive is used, therefore, the tube should be cutoff for the first 15 per cent. of trace or a severe stretch to compression will occur at the beginning of trace (top of picture).

Fig. 9 shows the load line[6] of a vertical-deflection output circuit having a ratio of R_o to

[6] Load line was derived by the method of Finite Operators as described by A. Preisman in "Graphical Construction for Vacuum Tube Circuits".

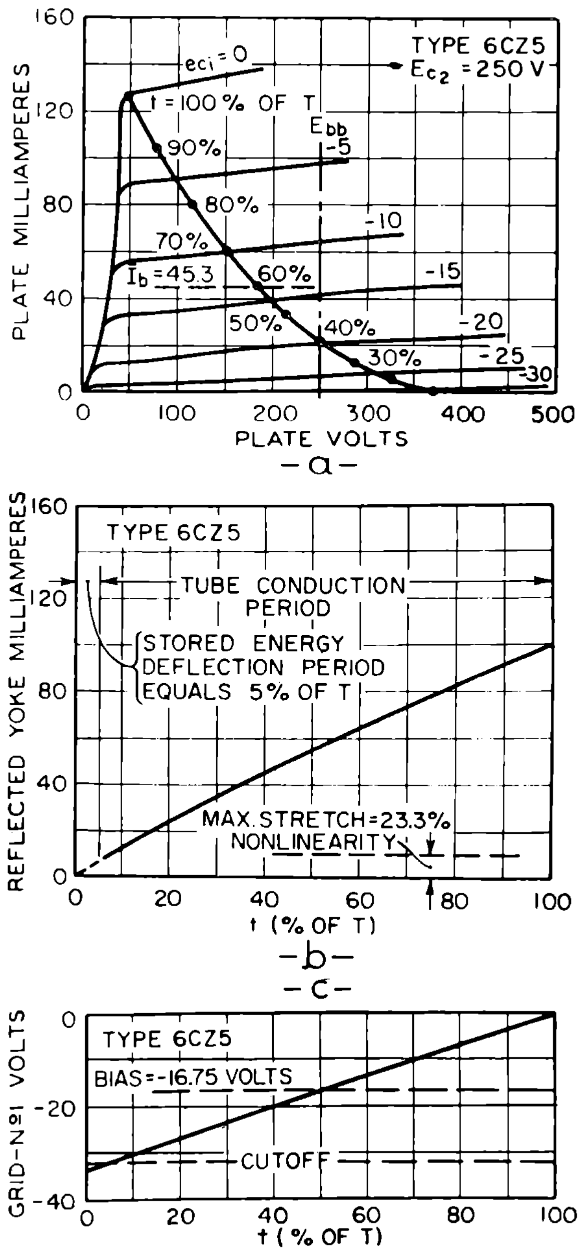


Fig. 9. Pentode Load Line for Picture Stretch.

$L_o + L_1$ equal to 120 (with no dc flowing in the transformer primary), and an average ratio of R_o to $L_o + L_1$ of 141 during the trace period (with an average of 45.3 milliamperes of plate current flowing in the transformer primary). The grid drive is linear and the bias is −16.75 volts. The yoke current referred to the primary, i_o, is stretched by 23.3 per cent. at the beginning of trace.

The initial 5 per cent. of trace occurs when the drive voltage is below cutoff and, therefore, must be entirely supplied by energy stored in the inductance L_o. Fig. 10 shows the vertical deflection equivalent circuit during the stored-energy deflection period (T_1). For good linearity during cutoff, $\left.\frac{di_o}{dt}\right|_{t=0} = 60\ I_{p-p}$ for a trace period of approximately 1/60 of a second. Therefore, the ratio of R_o to L_o plus L_1 must be equal to 120 during the time the valve is cut off.

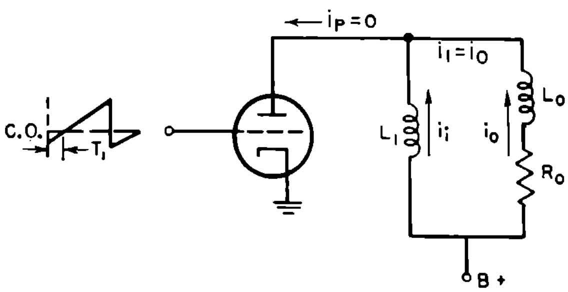


Fig. 10 — Vertical-Deflection Equivalent Circuit During Trace While Valve is Cut Off.

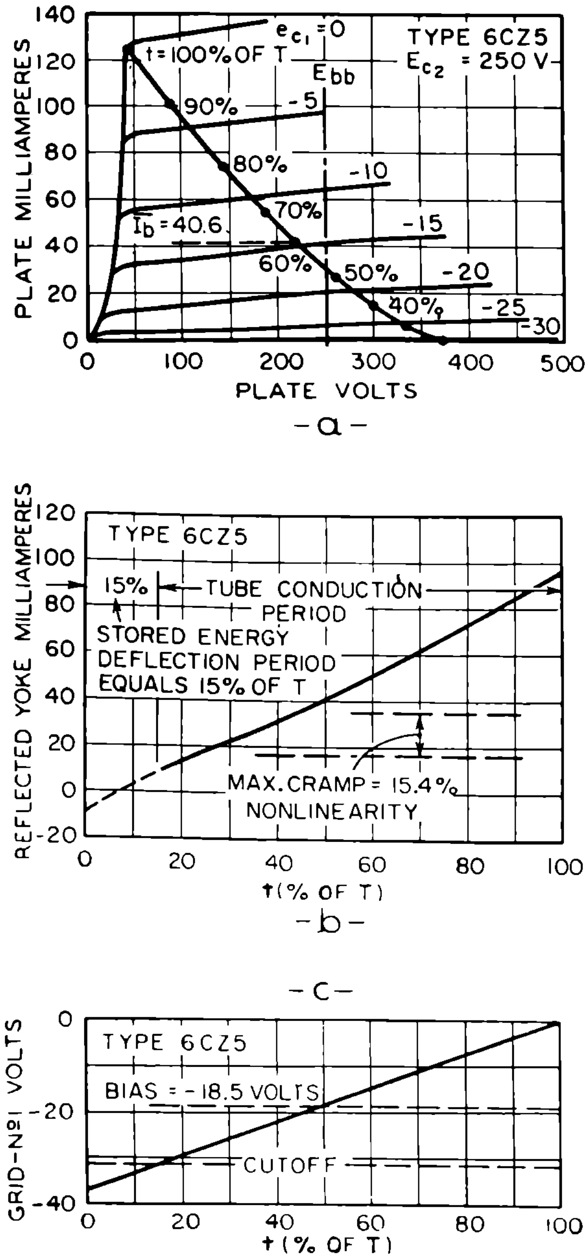


Fig. 11. Pentode Load Line for Top Picture Compression.

Cramp at Top of Picture

In the same circuit, if the bias is increased from −16.75 volts to −18.5 volts (for a constant deflection) as in Fig. 11, the yoke current changes from 23.3 per cent. stretched at the top of the picture to a cramp of 15.4 per cent. one-fourth of the way down from the top of the picture, and a general compression occurs over the top half of the picture. This cramp is caused by the straightening out of the load line due to an increase in the inductive reactance of L_1, which results from an increase in the rate of change of grid drive; i.e., the drive varies from cutoff to zero bias in 85 per cent. of the trace period rather than 95 per cent. as in the first example. This difference reduces the peak-to-peak change in the current i_1 flowing in the shunt inductor. The bias control is called the linearity control because the stretch or compression of the top of the picture is caused by a small bias change.

Because the first 15 per cent. of trace is supplied entirely by stored energy in the inductor L_o, the trace will tend to cramp slightly due to the exponential nature of this current. However, if the ratio of $R_o/(L_o + L_1)$ is kept nearly 120 during cutoff and no larger than 141 during valve conduction, then good linearity can be obtained by adjusting the bias to some value that will allow the valve to conduct during 85 to 95 per cent. of the trace period.

It may be necessary to add a small peaking component to the drive voltage to smoothly blend the switching transition from cutoff to conduction in circuits using sharp cutoff pentodes. Too much peaking will result in an exaggerated variation of linearity with normal product cutoff range. The most convenient source of this peaking voltage is a parallel resistor-capacitor combination in series with the charge capacitor. In most practical cases, the capacitor should be 0.1 microfarad and the resistor should be about 5000 ohms with its exact value adjusted to provide good linearity over the top one-fourth of the picture.

If the ratio of R_o to $L_1 + L_o$ is greater than 120 during cutoff, additional complex wave shaping is required to compensate for the tendency to stretch the top of the picture.

Additional stretching, cramping, and foldover can also result if the retrace damping is not nearly critical but is either oscillatory or over-damped as shown in Fig. 12. If the stored-energy deflection period (T_1) is 10 per cent. of the trace period (T) then the cramping due to the exponential nature of i_o will result in about 2 per cent. nonlinearity. If T_1 is made 15 per cent. of T, the resulting 4 per cent. nonlinearity approaches an objectionable value.

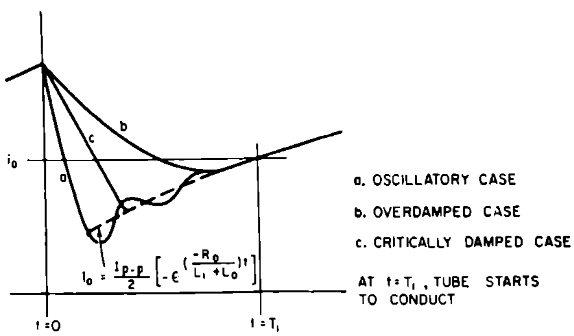


Fig. 12. Vertical-Deflection-Yoke Current Waveform During Cutoff.

The linearity of the yoke current during valve conduction is controlled by the magnetization curve of the output transformer as well as the valve characteristics. To simplify the analysis, however, the magnetizing current is assumed to be constant at 1/3 of the peak-to-peak plate current. The resultant error is relatively small.

The Practical Ratio of $R_o/(L_o + L_1)$ For Maximum Efficiency

In practice the vertical-deflection circuit is operating Class AB rather than Class A_1 and, therefore, the statements made earlier about efficiency should be modified. The ratio of R_o to $L_1 + L_o$ was given as 120 for maximum efficiency. However, if the valve is cut off during 15 per cent. of the trace period, the conduction period is only 85 per cent. of T and then $R_o/(L_o + L_1)$ may be as high as 141 (i.e., 120/0.85). Greater values than 141 will yield greater efficiency, but will require complex linearizing grid voltages. For this relationship, L_1 should be measured with the average dc operating current flowing in it. During T_1, however, the ratio of R_o to $L_o + L_1$ must be 120 for good linearity and the value of L_1 should be measured without dc current flowing in the transformer.

Height control is effected by varying the peak-to-peak grid drive supplied by the sawtooth generator. The proper drive voltage is obtained by adjusting the resistor in series with the battery and charge capacitor and thereby altering the time constant and changing the final voltage developed by the charge capacitor.

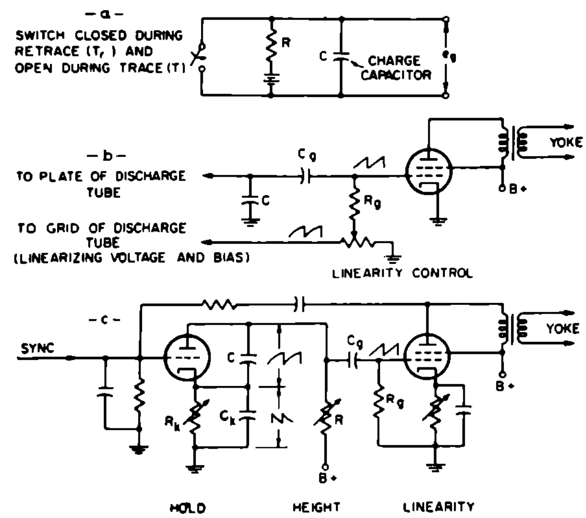


Fig. 13. Vertical-Deflection Sawtooth Generator and Linearizing Feedback Networks.

Linearizing Feedback

As was mentioned previously, it is not practical to obtain a linear drive directly from conventional sawtooth generators as shown in Fig. 13a. An approximate parabola must be added to the grid voltage, e_g, to obtain the required linear drive. Figs. 13b and 13c show simple methods of integrating a sawtooth to obtain an approximate parabola which can be added to e_g to give good linearity.

If Fig. 13b the bias voltage is derived from the grid of the discharge tube, and contains a large exponential waveform. This waveform is integrated by the action of the charge capacitor, C, which produces a parabolic waveform to correct for the voltage produced by the sawtooth generator. The degree of linearizing can be regulated by adjusting the size of the grid resistor R_g. In Fig. 13c the partially bypassed cathode resistor, R_k, generates the parabola which, when added to the saw-tooth generator voltage, gives a linear grid drive. The drive linearity can be corrected by adjusting C_g, R_g, or C_k. Sharp cutoff pentodes may require the use of a parallel resistor-capacitor combination in series with the charge capacitor as explained under the discussion of a cramp at the top of the picture.

The statements that have been made about pentode circuits using linearizing feedback to compensate for the effects of small grid capacitors and resistors can apply equally well to triode circuits. To keep the effects of reverse-grid current as small as possible both triodes and pentodes should use small grid resistors (2.2 megohms or less) and linearizing feedback as required.

Comparison of Triodes and Pentodes[7] in Vertical Output Applications[8]

A. Efficiency:

The maximum plate efficiency of presently-available pentodes is 40 per cent., and that of presently-available triodes is about 33 per cent. If the added power in the screen circuit of the pentode is considered, its efficiency is about 36 per cent. However, the most important criterion is the available power output for a given power input, and in this category presently-available triodes give only 75 per cent. of the deflection power of presently-available pentodes. Very-high-perveance triodes have been recently designed that will give more than 90 per cent. as much power as that available from pentodes.

B. Linearity:

Triodes can be designed into circuits without special linearizing networks. Triode circuit linearity will be more constant for a wide range in line voltage variation than will pentode circuits.

C. Plate Pulse:

Because of the nonlinear tube damping, 30 to 40 per cent. lower plate pulse voltages are obtained from triode circuits.

D. Cost:

Generally, the circuit components associated with triode circuits, including the triode itself, are less expensive than those for pentode circuits.

E. Peaking:

A peaking voltage is not required for pentode circuits, but is often used in special linearizing circuits for sharp-cutoff pentodes.

F. Bias and Drive:

Available triodes required roughly twice the bias and twice the drive required by pentodes. This difference is important for designs using cathode bias or bias from the grid of the horizontal oscillator. The drive available from discharge tubes operating from B boost is over 100 volt peak-to-peak which is adequate for all conventional triodes.

G. Low Line-Voltage Operation:

Pentodes are more efficient than triodes and will deliver approximately 10 per cent. greater power for a given power input (including screen power) than triodes having the highest perveance. However, triodes are less subject to power loss with low line voltages. If pentodes are to maintain their high efficiency, the load must be adjusted as the supply voltage is changed. Triodes are less critical in this respect, and it is possible that a triode having 10 per cent. less power capability at normal line voltage will actually operate on about 3 per cent. lower line voltage than the pentode. The reduced power capability of the triode at normal line voltgae is not important because the design must allow for about 10 per cent. reserve scan in the case of a triode and about 20 per cent. reserve scan for a pentode.

H. Damping:

Pentode circuits require a damping network and triode circuits utilize tube damping.

I. Plate Dissipation Limit per Envelope:

Pentodes can operate with much less cathode area (and heater power) to obtain a given current level at a particular plate voltage and can, therefore, have a cooler operating temperature or a higher plate dissipation limit for a given envelope size.

[7] Pentodes are understood here to include beam power tetrodes.

[8] This comparison considers systems operating from about 260 volts B+.

Bibliography

1. Schade, O. H. — "Magnetic Deflection Circuits for Cathode-Ray Tubes", RCA Review, September 1947.
2. Schlesinger, K. — "Magnetic Deflection of Kinescopes", Proceedings of the IRE, pp. 813-921 — August 1947.
3. Friend, A. W. — "Television Deflection Circuits"—RCA Review, pp. 98-138 — March 1947.
4. Cocking, W. T. — "Electromagnetic Frame Scanning"—Wireless World, pp. 217-222—July 1946, and pp. 289-291 — September 1946.
5. Knight, M. B.—"Practical Analysis of Vertical Deflection Circuits"—Tele-Tech, pp. 62-68-88-90-22 — July 1953.
6. Deutsch, S.—"Magnetic Deflection and Focusing"—Theory and Design of Television Receivers — pp. 356-395 — 1951.
7. Preisman, A. — Graphical Construction for Vacuum Tube Circuits, McGraw-Hill Book Co., Inc. — 1943.
8. "Minimizing Pulse Voltages in Television Vertical Deflection Amplifiers"—RCA Application Note AN-146 — December 1, 1950.

APPENDIX

Please see next page.

Acknowledgement

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Appendix

Current and Voltage Functions:

1. $i_o = I_{p\text{-}p}\left(\frac{t}{T} - \frac{1}{2}\right)$

2. $e_p = -\left(i_o R_o + L_o \frac{d_{io}}{d_t}\right) + E_{bb} = I_{p\text{-}p}\left[\left(\frac{R_o}{2} - \frac{L_o}{T}\right) - \frac{R_o}{T}\, t\right] + E_{bb}$

3. $i_1 = \frac{1}{L_1}\int (E_{bb} - e_p)dt = I_{p\text{-}p}\left[\frac{1}{2} - \left(\frac{R_o}{2} - \frac{L_o}{T}\right)\frac{t}{L_1} + \frac{R_o}{2L_1 T}\, t^2\right] \;;\; i_1\Big|_{t=0} = \frac{I_{p\text{-}p}}{2}$

4. $i_p = i_o + i_1 = I_{p\text{-}p}\left[\left(\frac{1}{T} - \frac{R_o}{2L_1} + \frac{L_o}{L_1 T}\right) t + \frac{R_o}{2L_1 T}\, t^2\right]$

Maximum Efficiency Relationships

1. $\frac{R_o}{L_o + L_1} \approx 120\,\Omega/H_y$ i.e., $\frac{d_{ip}}{dt}\Big|_{t=0} = 0 \;\therefore\; R_o = \frac{2}{T}(L_o + L_1)$, where $T \approx 1/60$ sec.

2. $\bar{I}_{b\,min} = \frac{1}{3} I_{p\text{-}p}$ i.e., $\bar{I}_b = \frac{1}{T}\int_0^T i_p\, dt = I_{p\text{-}p}\left[\frac{1}{2} - \frac{R_o T}{12 L_1} + \frac{L_o}{2L_1}\right]$

   assuming $L_o \ll L_1$ and $R_o = \frac{2}{T}(L_o + L_1)$ then $\bar{I}_{b\,min} = \frac{1}{3} I_{p\text{-}p}$

3. $\eta_{max} = 50\%$ *(Plate Efficiency)*

   $P_{in} = \bar{I} \cdot \frac{I_{p\text{-}p}}{2} \cdot R_o = (I_{p\text{-}p})^2 \frac{R_o}{6}$ $\qquad P_{out} = \frac{1}{T}\int_0^T i_o e_p\, dt = (I_{p\text{-}p})^2 \frac{R_o}{12}$

   $\therefore\; \eta_{max} = \frac{P_{out}}{P_{in}} = 50\%$

Reactive Power in Shunt Inductance (L_1):

1. $P_r = \frac{1}{T}\int_0^T i_1 e_p\, dt = (I_{p\text{-}p})^2\left(\frac{L_o}{2L_1 T}\right)$

2: In a typical example: $L_o \approx \frac{L_1}{10}$, $|R_o| = L_o \times 10^3$

   $\therefore\; P_r \approx \frac{(I_{p\text{-}p})^2 R_o}{12}(.036) = 3.6\%$ of P_o

Required Peak-to-Peak Plate Power

1. $P_{b(p\text{-}p)} = I_{b(p\text{-}p)} E_{p\text{-}p} = (I_{p\text{-}p})^2 R_o\left(1 + \frac{L_o}{L_1}\right) = P_{o(p\text{-}p)}\left(1 + \frac{L_o}{L_1}\right)$

2. In a typical example: $\frac{L_o}{L_1} \approx \frac{1}{10} \;\therefore\; P_{b(p\text{-}p)} \approx 1.1\, P_{o(p\text{-}p)}$

Reflected Yoke Current (i_o) During Trace While Tube is Cut Off (T_1)

1. $i_o = -L_1 = -\frac{I_{p\text{-}p}}{2}\left[\epsilon^{-\left(\frac{R_o}{L_o+L_1}\right)t}\right]$

2. $\left.\frac{d_{io}}{dt}\right|_{t=0} = \frac{I_{p\text{-}p}}{2}\left(\frac{R_o}{L_o+L_1}\right)$

3. For good linearity, $\left.\frac{d_{io}}{dt}\right|_{t=0} = \frac{I_{p\text{-}p}}{T} = 60\,I_{p\text{-}p}$, where $T \approx 1/60$ sec.

4. $\therefore \frac{I_{p\text{-}p}}{2}\left(\frac{R_o}{L_o+L_1}\right) \approx 60\,I_{p\text{-}p}$ or $\frac{R_o}{L_o+L_1} \approx 120\,\Omega/H_y$

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