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VIDEO AMPLIFIERS WITH INSTANTANEOUS  
AUTOMATIC GAIN CONTROL

by  
WILLIAM E. AYER

Technical Report No. 4  
April 17, 1953

Prepared under Office of Naval Research Contract  
Nonr 225 (10), NR 377 366  
Jointly supported by the U.S. Army Signal Corps,  
the U.S. Air Force, and the U.S. Navy  
(Office of Naval Research and the Bureau of Ships)



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Electronics Research Laboratory  
Stanford University  
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## SUMMARY

Circuits are described which allow essentially complete control of the output/input amplitude characteristic of a multi-stage video amplifier handling both positive and negative input signals. The incremental gain of each stage is determined instantaneously by the signal current through the tube. Germanium diodes are employed in the cathode circuits to introduce degeneration for signals above a predescribed amplitude. The gain reduction achieved in this manner lasts only as long as the signal, so that recovery time is not adversely affected.

Design relations are given for single stages, and for paired stages employing negative voltage feedback. This latter configuration, while more complicated from the design standpoint than individual stages, permits a wide choice of gain characteristics (including zero or negative gain for large signals), and is much less susceptible to power supply ripple.

The process of cascading several stages or pairs of stages to realize a desired output/input characteristic is discussed. Some of the problems that may be encountered with several cascaded stages are indicated, and solutions outlined.

Using the techniques described, amplifiers of 100 decibel dynamic range are readily constructed. As a typical example, amplifiers of this dynamic range and 100 decibel small-signal gain have been built using three dual triodes. The over-all bandwidth for this case was approximately 500 kc. With the gain characteristic adjusted so that the output signal is proportional to the logarithm of the input signal, such an amplifier may be used to provide a crystal-video receiver with cathode-ray tube display without resorting to a manual gain control.

The ability of these amplifiers to handle signals of both polarities simultaneously greatly reduces the overshoot problem usually encountered in pulse amplifiers. When pulse-width fidelity is required, an overshoot may be introduced near the front end to prevent the pulse stretching which otherwise occurs for strong signals.

An important feature of the circuit arrangements used is the stability with respect to changes of diode and tube characteristics with age and temperature.

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# VIDEO AMPLIFIERS WITH INSTANTANEOUS AUTOMATIC GAIN CONTROL

## 1. INTRODUCTION

In many microwave receiving systems, the incoming signal is detected directly, and then amplified to provide a usable output signal. When the received signals are of a pulsed nature, a video amplifier must be used following the detector to reproduce the envelope of the input signal. Arrangements of this sort are often called crystal-video systems.

Work at this laboratory with crystal video systems has prompted an investigation of video amplifiers possessing special characteristics. The principal results of this investigation are reported in this paper.

Prior to the current investigation, a considerable amount of work was done by others along the same lines of attack. Mr. J. R. Wilkerson of Airborne Instruments Laboratory was the first, to the author's knowledge, to work with the diode degeneration scheme, a form of which is employed here. Mr. W. R. Rambo (formerly of Airborne Instruments Laboratory and now at Stanford) made major contributions to the development. The chief work of the author has been in extending the circuitry to allow bipolar operation, and to develop design relations.

## 2. BASIC CIRCUIT ARRANGEMENT

The circuit of a single-stage amplifier is shown in Fig. 2.1. The quiescent currents through the diodes are established by the resistors  $R_{t1}$  and  $R_{t2}$ , which are large in comparison with  $R_k$ . The condenser  $C_k$  is large, so that it performs essentially as a battery for pulse amplification. The grid bias for the tube is obtained from the divider  $R_{b1}$  and  $R_{b2}$ .

For small signals the cathode-to-ground impedance is typically of the order of a few hundred ohms, as determined by the sum of the dynamic resistances of the diodes.  $R_k$  is always several thousand ohms, so that it has little effect upon the impedance to ground when the diodes are conducting. The equivalent circuit is shown in Fig. 2.2. Under these conditions the stage exhibits maximum gain. As the size of the incoming signal is increased, the current through the tube changes. For positive input signals, the current through  $D_2$  increases, while the current through  $D_1$  decreases. When the grid



signal is sufficiently large, conduction through  $D_1$  will cease, and for all larger signals the equivalent circuit will be that shown in Fig. 2.3. For large negative input signals the diode  $D_2$  will be out of conduction, and the operation may again be understood from Fig. 2.3

For small signals, then, the cathode degeneration is small, and the incremental gain is large. As the input level is raised, conduction through one of the other of the diodes stops, the cathode degeneration increases to that provided by the cathode resistor  $R_k$ , and the incremental gain of the stage drops. The shape of the output/input characteristic of a single stage is shown in Fig. 2.4. The upper curve illustrates the behavior that would be expected on the basis of "ideal" diodes (constant resistance for all forward voltages, and infinite resistance for all back voltages). The incremental gain would be large and constant up to the transition point, and then drop abruptly to a lower constant value for larger signals.

In the practical case of high-conduction germanium diodes, the dynamic resistance of the series diodes increases as the transition point is approached, so that the incremental gain falls off before the actual transition occurs. The shape of the output/input curve under these conditions is illustrated by the lower curve of Fig. 2.4.

In designing stages for specific applications, considerable freedom is available. Thus, the small-signal gain may be varied over wide limits. Similarly, the output voltage at which the gain transition occurs, and the incremental gain for large signals, may be adjusted to meet various requirements. Stages of different characteristics may be cascaded to provide a desired over-all characteristic. For example, it is frequently desired that the output voltage of the amplifier be approximately proportional to the logarithm of the input voltage. This characteristic may be closely approximated by using a fairly large number of identical stages, so that there are many transitions or "knees" in the output/input curve. In view of the gradual nature of the transitions realized with available germanium diodes, the approximation is better than one based upon straight-line segments. The logarithmic characteristic is obviously one of many which may be achieved. Several others will be discussed later.

The control of gain characteristic for both positive and negative input signals which is provided by the circuits described is important

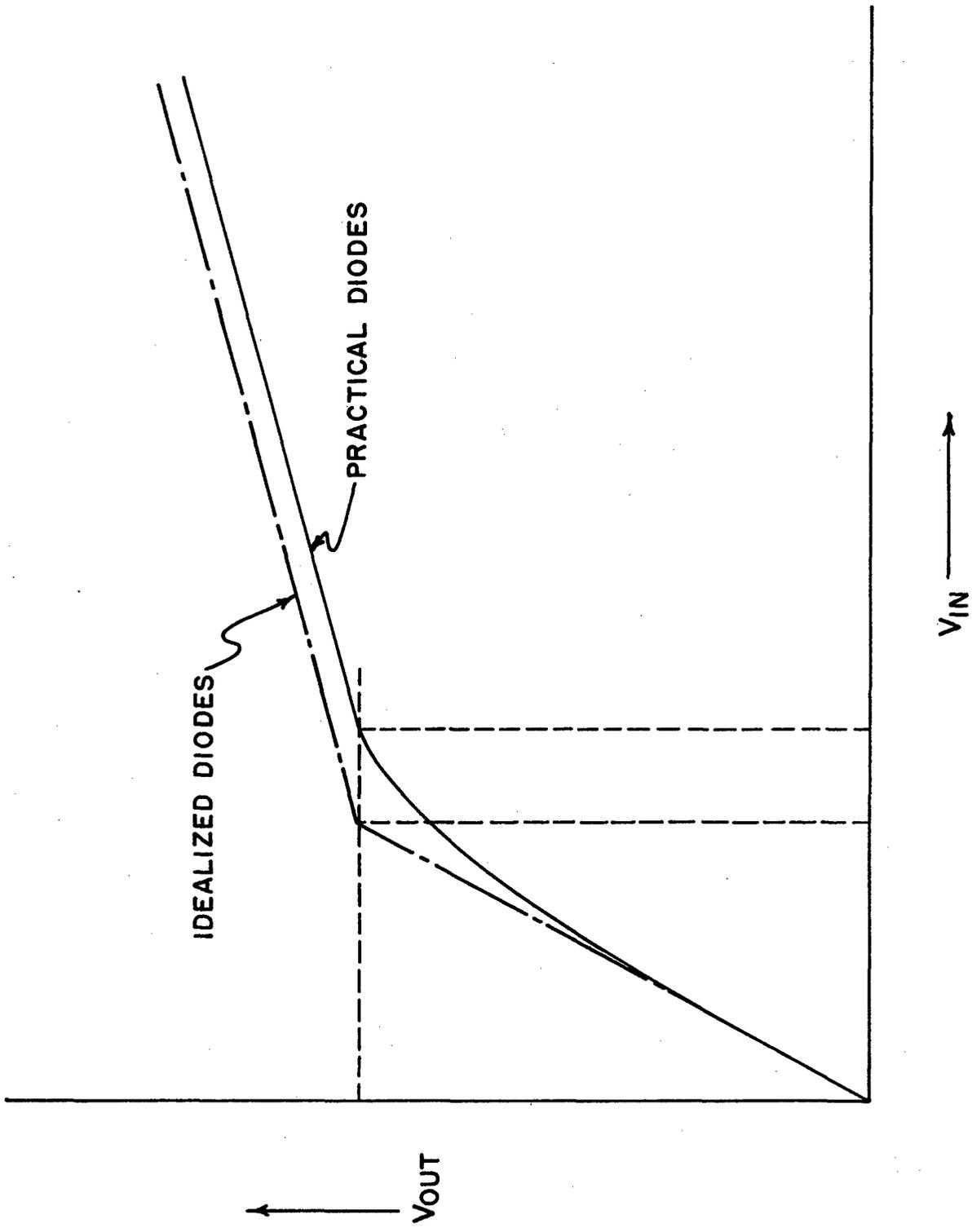


FIGURE 2.4 OUTPUT / INPUT CHARACTERISTIC FOR ONE STAGE

in many practical applications wherein pulses of only one polarity are to be amplified. First, it is often very convenient to introduce a short time constant somewhere in the amplifier to reduce stray low-frequency signals from power supplies, etc. If this is done in an amplifier which has large dynamic range for signals of only one polarity, the overshoot produced by such a short time constant can easily paralyze the system, resulting in a long recovery time. If the gain is controlled for signals of both polarities, this problem is greatly simplified.

In some applications it is necessary to preserve pulse width. When large signals are applied to an amplifier with IAGC, the weaker trailing portion of a received signal will be amplified more than the main pulse, so that very large amounts of stretching may occur. This may be circumvented with the bipolar amplifier by purposely introducing an overshoot early in the amplifier. The end of the pulse is thus defined by the baseline crossover point, and will be faithfully reproduced at the output.

#### Design Procedure

The design of a stage to provide a given output/input variation may be carried out in a fairly direct manner. Some cut-and-try is generally necessary, but convergence to the proper component values is rapid, once a reasonably good approximation has been made.

Basic specifications are the small-signal gain, the output voltage at transition, and the large signal gain. It is assumed that the operating point of the tube has been chosen, and that the gain characteristic specified is within the capabilities of the tube.

First, a reasonable value for the plate load resistor should be assumed. A satisfactory value for the first trial may be found by computing the plate load required to realize the desired small-signal gain without cathode degeneration, and then doubling this answer to allow for the dynamic resistance of the cathode diodes under quiescent conditions. In the typical case of triode amplifiers this estimated plate load is given by<sup>1</sup>

---

<sup>1</sup>The derivations of the equations of this section are indicated in Appendix A.

$$R_L = 2 \frac{A_1 R_p}{(\mu - A_1)} \quad (2.1)$$

where  $A_1$  is the desired small-signal gain for the stage.  $R_p$  is the dynamic plate resistance, and  $\mu$  the amplification factor, evaluated at the operating point.

Next, the transition point may be approximated. This is done by finding the change in tube current required through  $R_L$  to give the desired output transition voltage. Since the cathode Resistor  $R_k$  will generally be large in comparison with the dynamic resistance of the diodes, most of the tube signal current will come from the diode circuit. Thus, the output voltage at the time one of the diodes goes out of conduction will be given to a good approximation by

$$V_t \approx i_o R_L \quad (2.2)$$

where  $i_o$  is the quiescent diode current. When greater accuracy is desired in determining the output transition voltage, the expression

$$V_t = i_o R_L + v_o R_L / R_k \quad (2.3)$$

may be used, where  $v_o$  is the quiescent diode voltage drop. In typical applications involving high- $\mu$  tubes, the large-signal incremental gain is given approximately by

$$A_2 = R_L / R_k \quad (2.4)$$

so that (2.3) and (2.4) may be used to complete the first approximate evaluation of  $R_L$ ,  $R_k$ , and  $i_o$ .

When these tentative values have been determined, the final design may be started. First, the dynamic resistance of the series diodes for the quiescent current  $i_o$  should be found. This is simply twice the resistance for a single diode. Fig. 2.5 shows the measured dynamic resistance and voltage drop  $v_o$  as functions of  $i_o$  for 1N56 germanium diodes.

At this point, the plate load resistor may be recalculated. An equivalent plate resistance of the tube with cathode degeneration may be found from

$$R'_p = R_p + (1 + \mu) R \quad (2.5)$$

where  $R_p$  is the actual plate resistance, and  $R$  designates the resistance between cathode and ground. In this case  $R$  would be the total dynamic

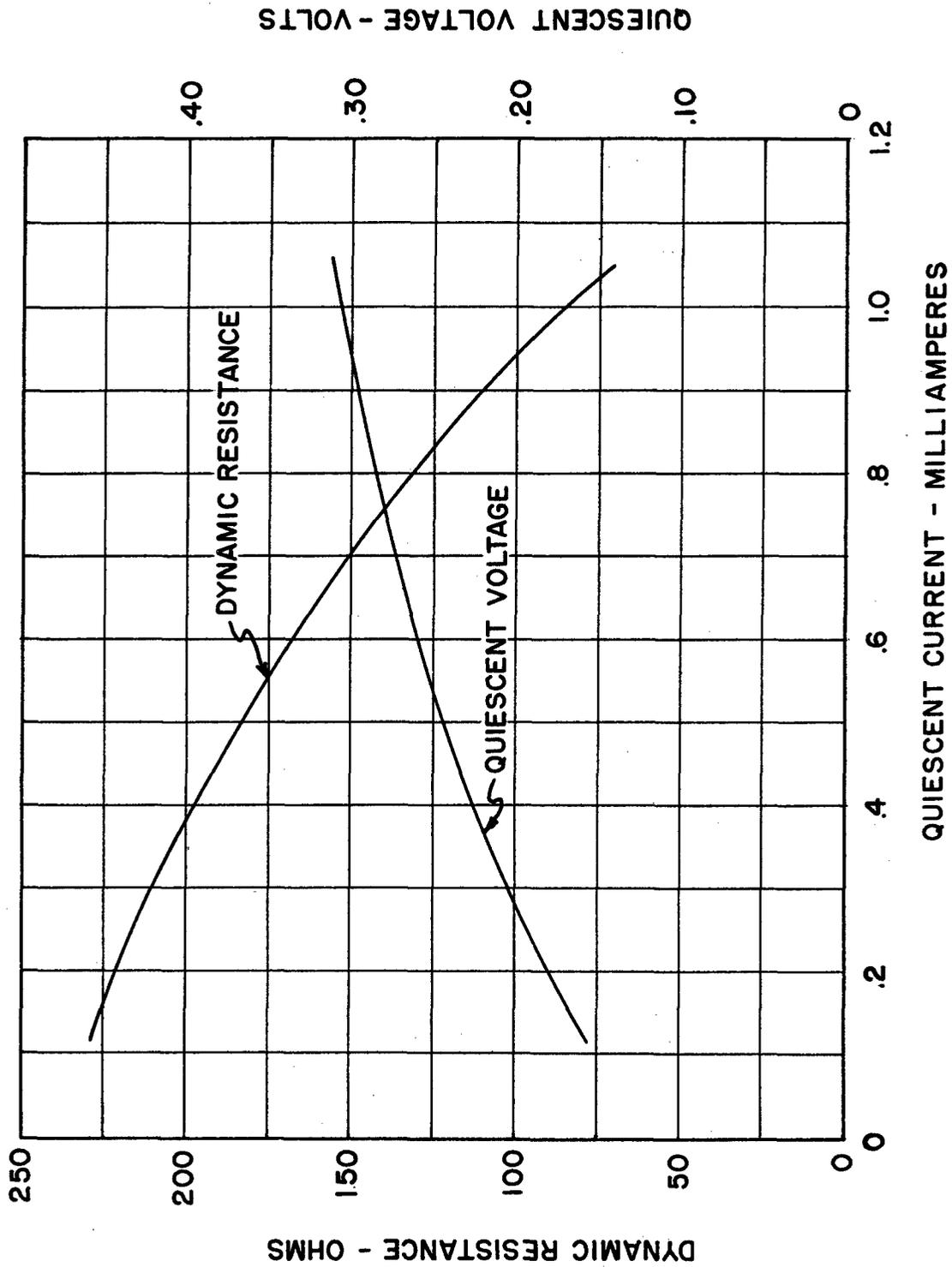


FIGURE 2.5 CHARACTERISTICS OF IN56  
GERMANIUM DIODES

resistance of the diodes with the quiescent current found from (2.2) or (2.3). The corrected plate load resistance value may now be found from

$$R_L = \frac{A_1 R'_p}{(\mu - A_1)} \quad (2.6)$$

where  $R'_p$  is the effective plate resistance obtained from (2.5) above.

With the new plate load a slightly different  $i_o$  will be required to re-establish the proper transition point, which, in turn, will result in a different dynamic diode resistance. This "closing in" on the proper values for  $R_L$  and  $i_o$  generally goes very rapidly, however.

When  $R_L$  and  $i_o$  have been established, the small-signal gain and the transition point are fixed, and it remains only to obtain the correct value for the incremental gain  $A_2$ . The circuit element used to control  $A_2$  is the cathode resistor  $R_k$ . Its value may be found from

$$R_k = \frac{R_L (\mu - A_2) - R_p A_2}{(1 + \mu) A_2} \quad (2.7)$$

To complete the design, values must be found for  $R_{t1}$  and  $R_{t2}$ ,  $R_{b1}$ , and  $R_{b2}$ ,  $R_g$ ,  $C_g$ , and  $C_k$ . Since the quiescent diode current  $i_o$  has been determined, and the operating tube current  $i_p$  has been chosen, the zero-signal potential of the cathode is given by

$$V_{ko} = (i_p - i_o) R_k \quad (2.8)$$

The potential at the junction of the two diodes will be  $V_{ko}$  less the drop across  $D_2$  for the current  $i_o$ . This latter voltage is small for high-conduction germanium diodes--typically of the order of 0.3 volts for 1N56's. The voltage drop across  $D_1$  will be equal to that across  $D_2$ , but of opposite polarity, so that the voltage at the lower end of  $R_{t1}$  will be equal to  $V_{ko}$ . Since  $R_{t1}$  is to carry a current  $i_o$ , it may be found from

$$R_{t1} = \frac{(V_b - V_{ko})}{i_o} \quad (2.9)$$

The size of  $R_{t2}$  may be determined in a similar way. The voltage at its upper end will be  $V_{ko} - v_o$ , where  $v_o$  is the quiescent drop across diode  $D_2$ . Since  $R_{t2}$  is to carry the quiescent currents of both diodes, the proper value is given by

$$R_{t2} = \frac{(V_c + V_{ko} - v_o)}{2i_o} \quad (2.10)$$

The grid-circuit components  $R_{b1}$ ,  $R_{b2}$ ,  $R_g$  and  $C_k$  are readily determined. First, it is usually convenient to make the sum of  $R_{b1}$  equal to or greater than  $R_{t1}$ . Since the voltage across  $C_k$  is generally small in comparison with the positive supply voltage, this choice simplifies design in that no appreciable fraction of the current through  $R_{t1}$  will be diverted from  $D_1$ . It also helps to maximize the discharge time-constant of  $C_k$ . The division of resistance between  $R_{b1}$  and  $R_{b2}$  is determined by the operating grid bias chosen for the tube, and the potential  $V_{ko}$  existing at the junction of  $R_{b1}$ ,  $R_{t1}$ ,  $D_1$ , and  $C_k$ . If the desired grid bias is represented by  $V_g$ , the values of the bias resistors may be found from

$$\frac{R_{b1}}{(R_{b1} + R_{b2})} = \frac{V_g}{V_{ko}} \quad (2.11)$$

Since the sum of the two resistances is assumed to have been established, the individual values may be found from (2.11).

The values for  $R_g$  and  $C_k$  are determined by the usual considerations involved in pulse amplifier design, and need not be discussed here.

The choice of the condenser  $C_k$  is governed by the consideration that the voltage at its upper end should not vary appreciably in the presence of strong signals. When one of the diodes is out of conduction, a current  $i_o$  flows into or out of the condenser, tending to change its voltage. The magnitude of this voltage change under specified conditions is easily found from

$$\Delta V = \frac{i_o t}{C_k} \quad (2.12)$$

where  $t$  is the length of the pulse. The manner in which a voltage variation across  $C_k$  affects the operation of the amplifier is rather complex, but the primary result is a change in cathode potential. The quiescent cathode voltage following a strong positive input signal, for example, is increased by a somewhat smaller amount than the condenser voltage change given by (2.12). The result at the output of the stage is thus an overshoot. In practice  $C_k$  is generally made as large as convenient on the basis of available condensers, typically of the order of tens of microfarads.

### 3. A TYPICAL STAGE DESIGN

The design procedure just described may be illustrated by a practical example. Two stages of an amplifier are shown in Fig. 2.6. The plate load resistor  $R_L$  was chosen to provide an over-all bandwidth of a five-stage amplifier of approximately 500 kc, corresponding to a stage bandwidth of 1.3 Mc. A small-signal gain of about 85 db for the five stages was required. On this basis the voltage gain per stage ( $A_1$ ) would be 7.1. The output voltage at the gain transition was to be approximately 3.5 volts, and the large-signal gain ( $A_2$ ) approximately 1.

Using (2.3), (2.4), (2.5), and (2.6), and the diode dynamic resistance curve of Fig. 2.5, it was found that the desired performance could be obtained with a plate load resistor of 4.7k and a quiescent diode current of 0.7 ma. A tube current of 4 ma was used, giving a plate resistance of 18k and an amplification factor of 60. The required grid bias for this operating point was 2.8 volts. Substituting  $2(150) = 300$  ohms from Fig. 2.5 in (2.5) and solving (2.6), the required plate load was found to be 4.9k. The effective interstage capacitance was estimated to be 25 micro-microfarads, giving an upper 3 db point of 1.35 Mc. The cathode resistor  $R_k$  was next calculated using (2.7).

The values of the remaining circuit elements may be found easily. The quiescent cathode voltage from (2.8) is

$$(4.0 - 0.7)\text{ma}(4.7)\text{k} = 15.5 \text{ volts}$$

$R_{t1}$  will then be, from (2.9)

$$(250 - 15.5)\text{v}/0.7\text{ma} = 335\text{k}.$$

The other diode bias resistor  $R_{t2}$  will be, from (2.10)

$$(250 - 15.5 - .3)\text{v}/1.4\text{ma} = 189\text{k}.$$

The grid bias divider composed of  $R_{b1}$  and  $R_{b2}$  may be determined using (2.11). This relation may be rewritten as

$$R_{b1} = \frac{V_g/\sqrt{V_{ko}}}{R_{b2} - V_g/\sqrt{V_{ko}}} \text{ megohms,}$$

with  $R_{b2}$  also in megohms. This form is sometimes more convenient, as it allows  $R_{b2}$  to be chosen as a standard value. In the example,  $R_{b2}$  was assumed to be 1 megohm. For a grid bias voltage of 2.8 volts and a quiescent cathode voltage of 15.5 volts,  $R_{b1}$  is found to be 220k.

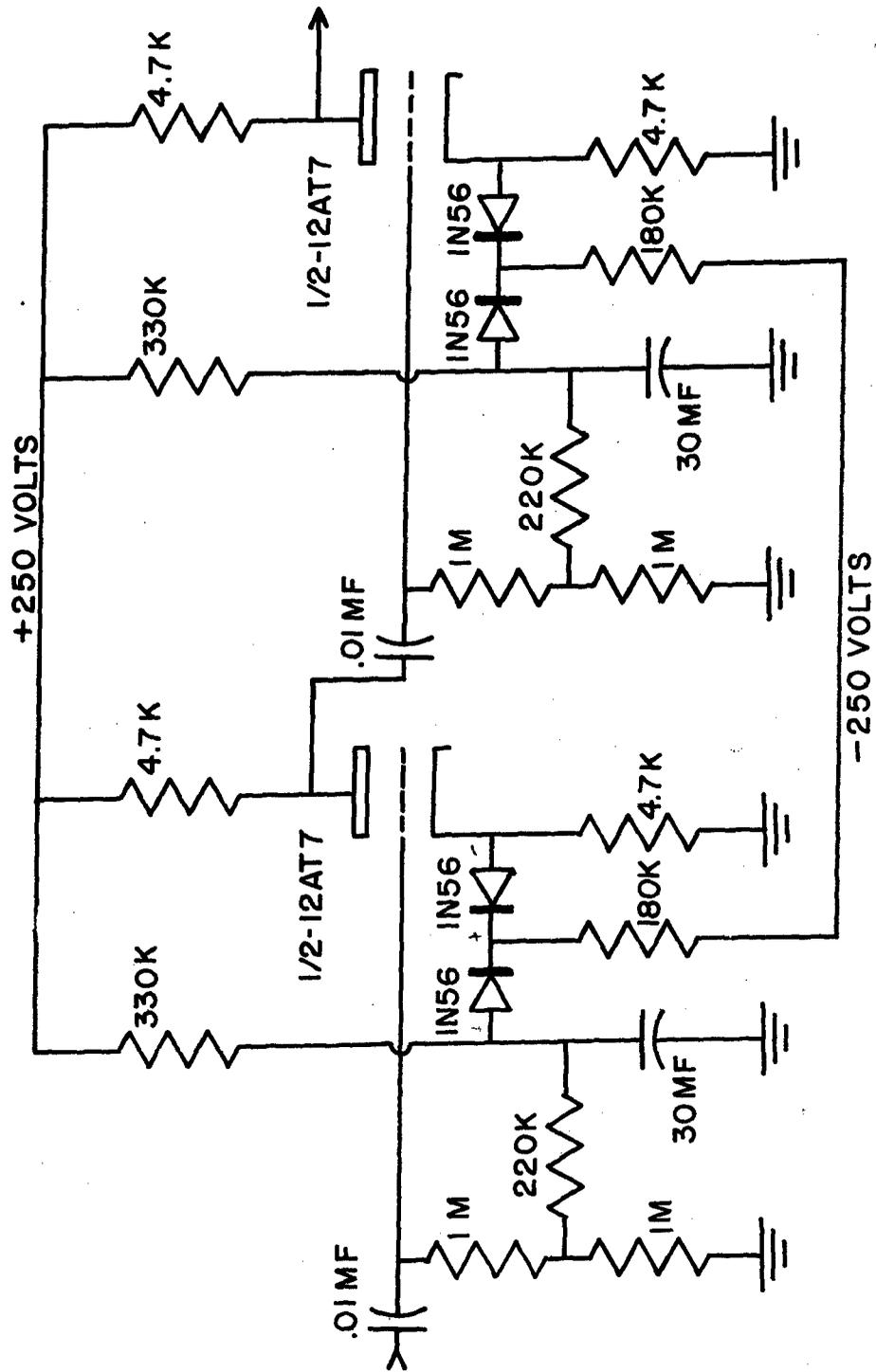


FIGURE 2.6 TWO STAGES OF A TYPICAL AMPLIFIER

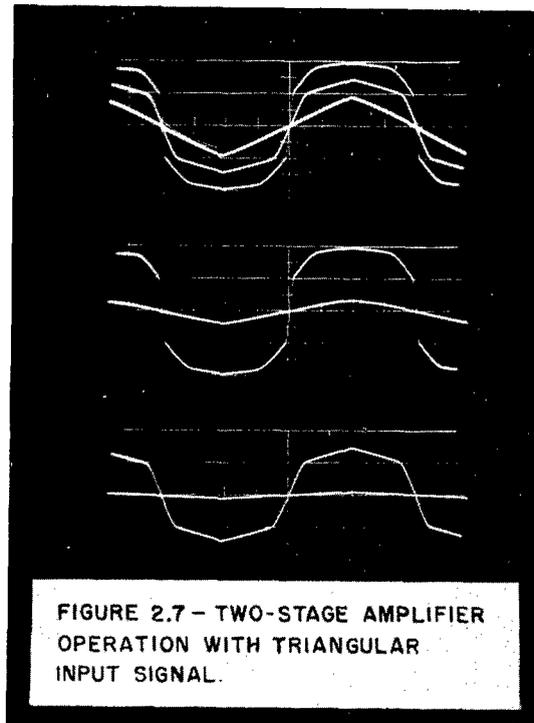
The cathode condenser  $C_k$  was made 30 microfarads. This value was chosen partly because dual units of convenient size were available. From (2.12) the change in condenser voltage during a 5-microsecond pulse (of sufficient amplitude to drive one of the diodes out of conduction) would be

$$\frac{0.7 \text{ ma } 5 \times 10^{-6} \text{ sec}}{30 \times 10^{-6} \text{ farads}} = 116 \text{ microvolts.}$$

In practice it has been found that voltage changes up to about 500 microvolts produced no appreciable degradation of the performance of high-gain multistage amplifiers, so that the 30 microfarad value for  $C_k$  is adequate in most cases. Trouble may be encountered in some applications when an overshoot is purposely introduced early in the amplifier. Under such conditions, the overshoot from a strong signal would be stretched in passing through the amplifier, so that the "off-time" of some of the later diodes might be many times greater than the length of the pulses being amplified. The important time in determining voltage change on  $C_k$  would then be the duration of the overshoot, and this could be very large. In such cases, the overshoot should be made as small as possible, consistent with the requirements to be met, and  $C_k$  should be made as large as possible. Additional improvement can sometimes be realized by limiting all signals above a prescribed amplitude before entering the amplifier.

The choice of the interstage coupling components is based upon normal pulse amplifier considerations of desired low-frequency response, allowable grid-circuit resistance, and physical size of the coupling condenser. Values of 1 megohm for  $R_g$  and 0.01 microfarads for  $C_g$  have been found satisfactory for most applications which required pulse lengths of the order of 0.5 to 5 microseconds to be handled.

Fig. 2.7 shows a series of oscilloscope photographs of input and output signals for the two stages of Fig. 2.6. The input signal was a triangular wave with a fundamental frequency of 1200 cycles/second. With this sort of signal the diodes are out of conduction for long periods, so that appreciable change in cathode condenser voltage takes place. This causes some shift in the positions of the gain transitions. The lower picture shows the input and output signals for the condition of only the second stage going through a gain transition. The center picture is for a larger input



signal, both stages being driven into their low-gain regions. The upper picture shows the change in output signal shape as the input signal is varied. The lower of the three curves shows linear performance, neither of the stages having passed a transition. The upper two waveforms are for successively larger input signals, and are identical to the output signals shown in the lower two photographs.

The measured performance of the two-stage amplifier with pulse signals is shown in Fig. 2.8. A pulse width of 5 microseconds and a repetition rate of 2000 cycles/second were used. The output has been plotted linearly in voltage, and the input linearly in decibels. The closeness to which a true logarithmic response is approximated is somewhat surprising in view of the fact that a 50-decibel input range is being covered with only two gain transitions (three segments). The maximum departures from the logarithmic characteristic are approximately 2 decibels.

#### 4. MODIFICATIONS OF THE BASIC CIRCUIT

There are several variations of the basic circuit which may be used to advantage in various applications. First, there is the obvious possibility of using pentodes in place of triodes. The design procedure in the

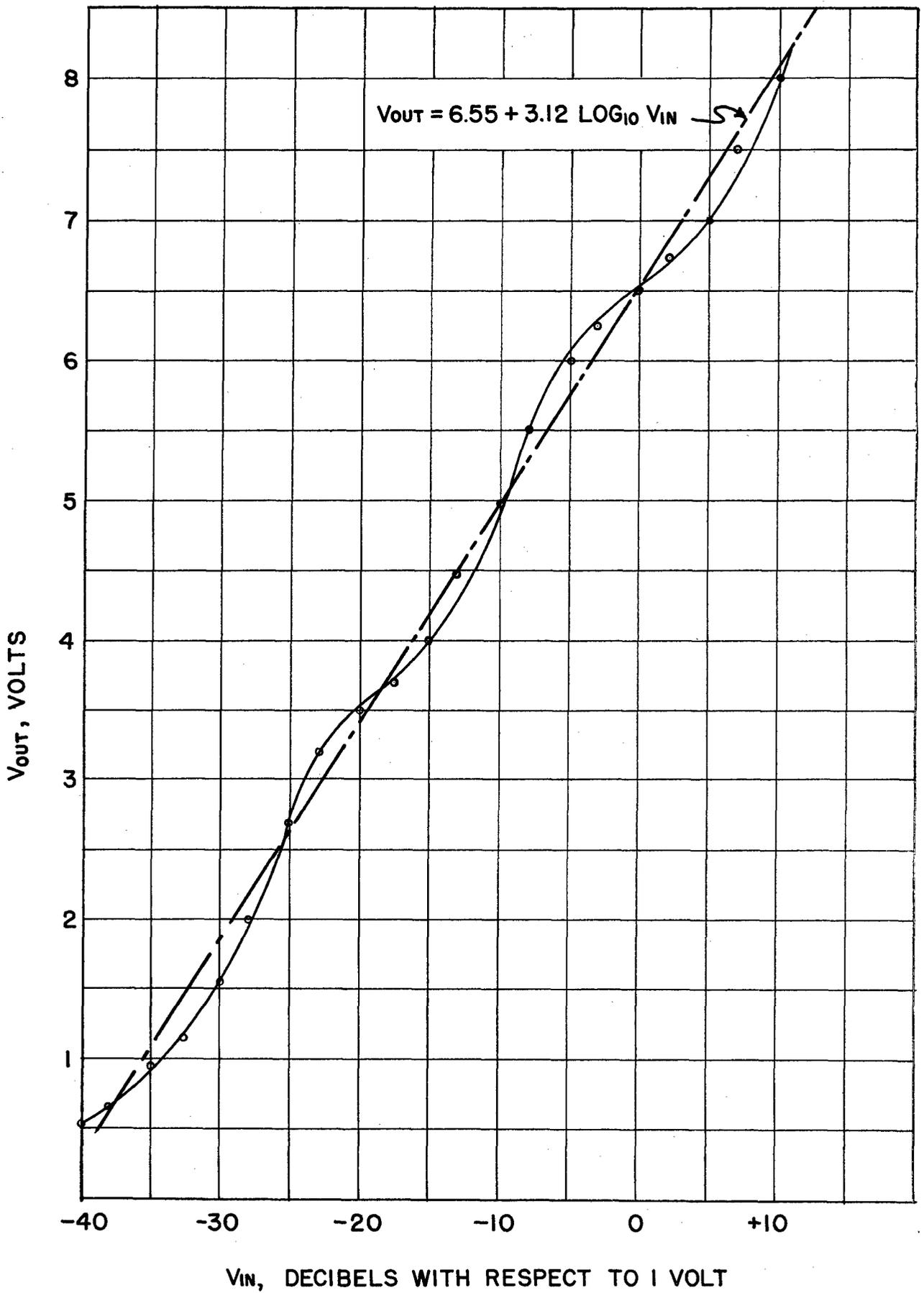


FIGURE 2.8 INPUT/OUTPUT CHARACTERISTIC OF THE TWO-STAGE AMPLIFIER OF FIGURE 2.6

case of pentodes will be slightly different than that for triodes because of the presence of screen current in the cathode circuit. The quiescent diode current for a given output transition voltage will thus be somewhat larger, as will the quiescent cathode voltage. The gain-bandwidth considerations will, of course, be different. To provide a given gain at some specified bandwidth, a considerably smaller number of tube sections will be required than in the triode case. Typically, this ratio approaches one to two, so that the number of pentodes would be only slightly greater than the number of dual triodes to do the same job. In applications requiring the pass-band to go up to several megacycles, the instability of triodes, due to their high grid-plate capacitance, would enter the picture, and pentodes would be the best choice. At the lower bandwidths (up to 1 Mc or so) triodes are quite satisfactory. For many applications, the use of triodes is advantageous because of the relatively large number of stages used. Since there is one gain transition per stage, a large number of stages results in an output characteristic made up of a large number of segments. A relatively good approximation to a desired over-all characteristic may thus be obtained.

There are several other minor modifications of the basic circuit of Figs. 2.1 and 2.6 which have frequently been employed. The negative high-voltage supply may be eliminated, and the common diode bias resistor  $R_{t2}$  returned to ground. The effect of this change is twofold: it reduces the value of  $R_{t2}$  to such an extent that the large-signal incremental gain is affected, and it causes a slight reduction in the stability of the circuit with respect to changes in components. For large positive signals  $R_{t2}$  will be effectively in parallel with  $R_k$  so that the large-signal gain will be increased if  $R_{t2}$  is not several times greater than  $R_k$ . (in the two-stage amplifier of Fig. 2.6,  $R_{t2}$  would have to be 10.5k to maintain the original diode currents.) Since  $R_k$  alone determines the incremental gain for large negative input signals, the result of this change is to make the large-signal gain less for positive than for negative signals. Since the polarity of the signal is reversed in going through each stage, this difference in gain tends to average out in a multistage amplifier. The over-all input/output characteristic will not be as smooth as that obtainable using a negative supply, but for many applications the operation is entirely satisfactory.

The reduction of  $R_{t2}$  also causes some loss of the circuit's ability to adjust to changes in tube and diode characteristics, although this has not been found at all serious in practical amplifiers where  $R_{t2}$  has been of the order of 10k to 20k.

A second type of circuit change which may be made is to keep the negative high-voltage supply, and to return both  $R_{t2}$  and  $R_k$  to it. This allows  $R_k$  to be made very large, which means that the large-signal incremental gain can be made very small. Thus the circuit of Fig. 2.6 could be modified in this way, raising  $R_k$  to 80k. The large signal gain would then be (using (2.7)) only 0.057. This corresponds to rather effective limiting action, and is easily obtained circuitwise.

Another modification of the basic circuit which may be found useful is to bias the two diodes unequally so that the gain transitions occur at different voltages for positive and negative input signals. This brings out the fact that the over-all characteristic of an amplifier may be adjusted separately for positive and negative input signals. The principal circuit change is in the values of the diode bias resistors  $R_{t1}$  and  $R_{t2}$ . With unequal quiescent currents, the dynamic resistances of the two diodes will, of course, be different. The sum of these resistances would be used to determine the cathode degeneration in the small-signal case. Output signal voltage at transition would be found from (2.2) or (2.3), with the appropriate quiescent current and voltage substituted for  $i_o$  and  $v_o$ , respectively.

## 5. THE FEEDBACK PAIR

Up to this point the discussion has been limited to simple grounded-cathode stages. For two reasons which will be stated shortly, it is often advantageous to operate the stages in pairs with voltage feedback between grid and plate of the second tube. This sort of arrangement has been described in the literature<sup>1</sup> as a feedback pair. The particular arrangement which has been found useful in this work as shown in Fig. 5.1.

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<sup>1</sup>Several forms of this circuit, though not the exact one used here, are discussed by Van Voorhis in "Microwave Receivers," vol. 23 of MIT Radiation Lab. Series, pp. 524-526.

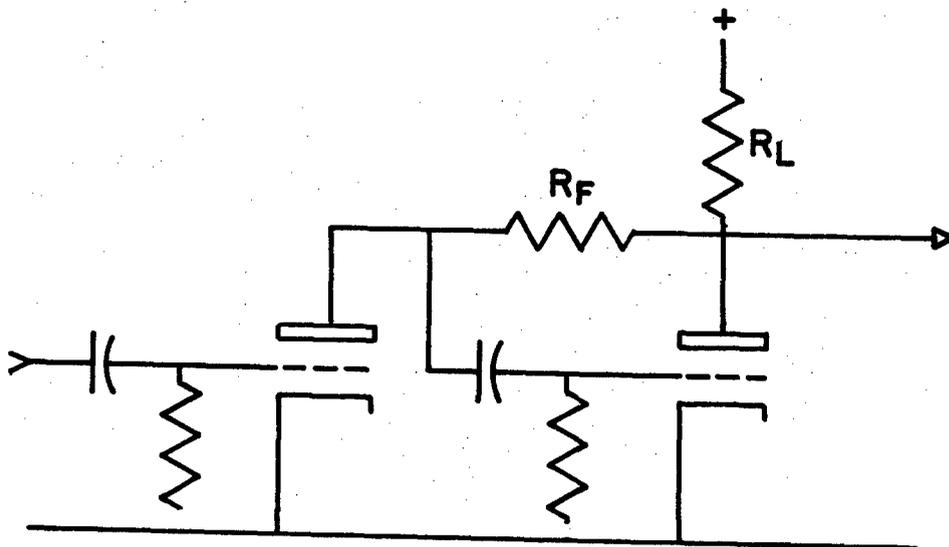


FIGURE 5.1 THE FEEDBACK PAIR

In this circuit, the plate voltage for the first tube is supplied from the plate of the second tube through the feedback resistor  $R_F$ . Since the second stage has grid-to-plate voltage feedback, it will have a resistive input impedance determined by the amount of feedback. This input resistance is given by<sup>1</sup>

$$R_{in} = \frac{R_F R_2 + R_F R_L + R_2 R_L}{R_2 + (1 + \mu_2) R_L} \quad (5.1)$$

where  $R_2$  is used to designate the effective plate resistance of the second tube. In circuits with no cathode degeneration,  $R_2$  would be just the dynamic plate resistance of the tube. With cathode degeneration such as would be introduced in using the IAGC circuits,  $R_2$  would be given by Eq. (2.5).

Since the plate load of the first stage is provided by the input resistance of the second, the first stage gain is

$$A_1 = \frac{\mu_1 R_{in}}{R_{in} + R_1} \quad (5.2)$$

where  $R_1$  denotes the effective plate resistance, and  $\mu_1$ , the amplification

<sup>1</sup>The derivations of the equations in this Section are indicated in Appendix B.

factor, of the first tube. The first stage voltage gain given above is measured between grid and plate of the first tube.

The voltage gain of the second stage may be found from

$$A_2 = \frac{\mu_2 R_L (\mu_2 R_F - R_2)}{R_F R_2 + R_F R_L + R_2 R_L} \quad (5.3)$$

where  $\mu_2$  is the amplification factor of the second tube, and the remaining notation is as previously defined.

The voltage gain of the pair is just the product of  $A_1$  and  $A_2$ . This may be written as

$$A_T = \frac{\mu_1 \mu_2 R_L (\mu_2 R_F - R_2)}{R_F R_2 + R_F R_L + R_2 R_L + R_1 R_2 + (1 + \mu_2) R_L} \quad (5.4)$$

Another quantity of particular interest is the impedance seen at the plate of the second tube. The output resistance due to the feedback is given by

$$R_{out} = \frac{R_1 + R_F}{R_1/R_2 (1 + \mu_2) + R_F/R_2 + 1} \quad (5.5)$$

The actual impedance seen at the plate of the tube would be the parallel combination of  $R_{out}$  and  $R_L$ .

The advantages of the feedback pair for this type of amplifier may now be stated. First, note that the gain of the second stage is proportional to the difference between  $\mu_2 R_F$  and the effective plate resistance of the second tube,  $R_2$ . If the IAGC diode arrangements are incorporated in the cathode circuits of the pair, it is seen that the large-signal gain of the second stage may be varied over wide limits by changing the second stage cathode resistor  $R_{k2}$ , which, in turn, controls  $R_2$ . It is readily possible to make the large-signal gain (which will be designated by  $A_{22}$ ) zero, or even negative, if desired.

Since this same difference term appears in the over-all gain expression given by (5.4), the gain of the pair above the first transition point may be controlled to a large extent by  $R_{k2}$ . In the particular case of  $A_{22}$  equal to zero ( $\mu_2 R_F = R_2$ ), the total gain  $A_T$  is also zero for all signals greater than that required to drive the second stage to its gain transition. Pairs arranged in this way thus operate as limiters, and as such are very useful in certain applications. For this special case the value of

$R_{k2}$  may be readily found using Eq. (2.5). Thus

$$R_F = \frac{R_{p2} + (1 + \mu_2)R_{k2}}{\mu_2} \quad (5.6)$$

which, in the case of typical high- $\mu$  triodes may be reduced to

$$R_{k2} = R_F \quad (5.7)$$

with good accuracy.

The second advantage of the feedback pair results from the low second-stage output resistance resulting from its grid-to-plate resistive feedback. As a result of this low resistance, the feedback pair is relatively insensitive to power supply ripple. This consideration is of importance at low signal levels (stages operating at maximum gain), and this is the condition under which maximum discrimination against supply voltage ripple is realized. As will be shown later, the ripple output of a feedback pair is typically something like 20 decibels less than that of a pair of stages without feedback, but of similar gain characteristics.

#### Design Considerations

The techniques for determining component values in the basic grounded-cathode circuit are applicable, in general, to the feedback pair, but several new considerations must be dealt with. The necessary departures in design procedure are best brought out by following through the action of a pair with IAGC circuits incorporated.

For small input signals the output will be proportional to input, and the two stages will operate at maximum gain. As the input level is raised, a point will be reached at which one of the diodes in the second-stage cathode circuit will cease to conduct, so that a change in over-all incremental gain will occur. One point to note here is that all of the change in second tube current required to reach this first transition point does not flow through the load resistor  $R_L$ , some going into the feedback resistor  $R_F$ .

When the gain of the second tube drops, the input resistance of the stage increases. Accordingly, the gain of the first stage increases, so that the decrease in over-all gain is much less than the decrease in second stage gain. (This is to be expected because of the gain-stabilizing pro-

perties of negative voltage feedback.) As the input signal is increased further the first stage goes through its transition, and the over-all incremental gain decreases again. The over-all characteristic may be stated in terms of the individual stage parameters. The required stage gain characteristics to realize a desired over-all performance are illustrated in Fig. 5.2. The small-signal gain of the pair,  $A_1$ , is equal to the product

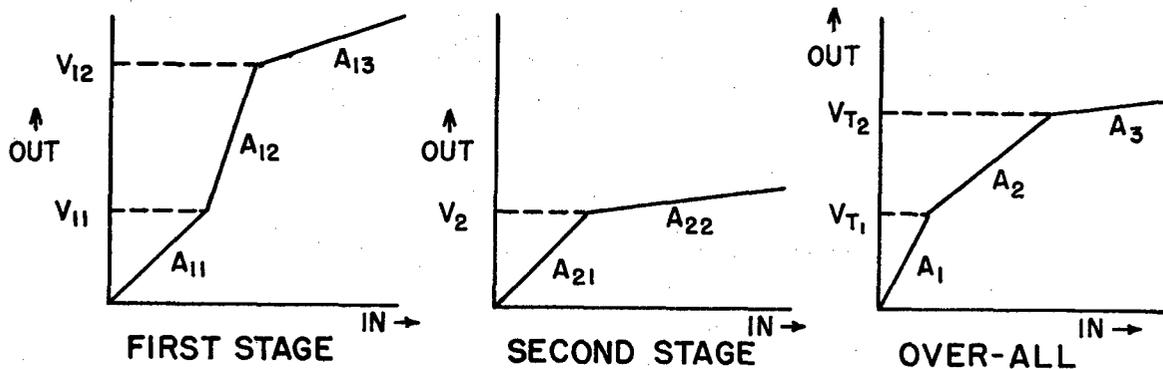


FIGURE 5.2 GAIN CHARACTERISTICS OF PAIR WITH IAGC

of the small-signal gains of the first and second stages,  $A_{11}A_{21}$ . The intermediate pair gain  $A_2$  is given by the product of the first-stage intermediate gain, and the second-stage large-signal gain,  $A_{12}A_{22}$ . The large-signal gain of the pair is given by the product of the first-stage large-signal gain and the second-stage large-signal gain,  $A_{13}A_{22}$ . The first output transition voltage  $V_{T1}$  is equal to  $V_2$ ; the second,  $V_{T2}$ , will occur at a voltage  $V_{T1} + (V_{12} - V_{11})A_{22}$ .

When the desired over-all characteristic is known, the design of the individual stages must be worked out in a cut-and-try fashion, due to the dependence of first-stage operation upon that of the second stage. Thus to obtain the proper intermediate gain, the large-signal gain  $A_{22}$  of the second stage must be less than it would be in an amplifier made up of separate grounded-cathode stages, to overcome the increase in first-stage gain from  $A_{11}$  to  $A_{12}$ . With this lower  $A_{22}$  gain,  $A_{13}$  must then be increased to provide the desired final gain  $A_3$ . As a rough working rule, the cathode-to-ground resistors in a feedback pair will be approximately twice the size used in separate stages to obtain a given response.

As mentioned earlier, the quiescent diode currents in the second stage must be somewhat greater than would be used with separate stages to realize the same output voltage at the first transition. The required quiescent current may be found from

$$i_{o2} = V_{T1} \left[ 1/R_L + (1 + 1/A_{21})/R_F \right] - v_{o2}/R_{k2} \quad (5.8)$$

The first term represents the change in current through the load resistor to provide the desired transition. The second term accounts for the additional current which must be supplied to the feedback resistor. The last term represents the increment of tube current which is supplied from the cathode resistor  $R_{k2}$ , as discussed in Section 2 and Appendix A.

The diode currents required in the first stage may be specified in a similar way by the relation

$$i_{o1} = \frac{V_{T1}}{R_{in1}A_{21}} + \frac{V_{T2} - V_{T1}}{R_{in2}A_{22}} - \frac{v_{o1}}{R_{k1}} \quad (5.9)$$

Here  $R_{in1}$  designates the input resistance of the second stage below its gain transition, and  $R_{in2}$  the input resistance above the gain transition, as calculated from Eq. (5.1).

From the preceding discussion of the feedback pair, it is apparent that the design procedure is considerably more complicated than for the basic circuit without feedback. Perhaps the most difficult phase of the design is to achieve a rough approximation to the desired performance. Once this has been obtained, the directions and approximate amounts of component changes may be estimated, and a suitable final arrangement achieved without particular difficulty. To facilitate the process of obtaining a first approximate circuit, a general procedure is given below.

Once reasonable parameters ( $\mu$  and plate resistance) have been chosen for the tube to be used, an approximate value for the feedback resistor  $R_F$  may be found from (5.4). To do this, values must be assumed for the dynamic resistances of the cathode diodes so that the effective plate resistances of the tubes ( $R_1$  and  $R_2$ ) may be found. Also, a value must be assumed for  $R_L$ . (The effect of a poor value for  $R_L$  is not very serious as it has relatively little effect upon the small-signal over-all gain  $A_1$ .)

When a first value for  $R_F$  is determined, the small-signal gain of the

second stage may be found from (5.3). An approximate value for the output voltage at the first gain transition may now be found from Eq. (5.6) by neglecting the last term. The value of  $R_2$  above the first transition may be found using (5.4). The second cathode resistor  $R_{k2}$  may now be found using Eq. (2.5). The actual value of  $V_{T1}$  may now be determined using (5.6). The required value for  $R_1$  may now be found from (5.4).  $R_{k1}$  may now be found from (2.5). The output voltage at the second transition can now be determined directly, using (5.1), (5.3), and (5.7).

Using this procedure, the desired over-all gains ( $A_1$ ,  $A_2$ , and  $A_3$ ) will be realized, but the output transition voltages ( $V_{T1}$  and  $V_{T2}$ ) will not generally be correct. The process of correcting the transition points involves changing the quiescent diode currents and recalculating the over-all response. If the initial assumptions are not too far out, the final values can be found with one or two such recalculations.

The values for the diode bias resistors, grid bias and coupling resistors, etc., may be determined directly from the appropriate relations given in Section 2.

The circuit of a typical feedback pair designed for a logarithmic output/input characteristic is shown in Fig. 5.3. The measured perform-

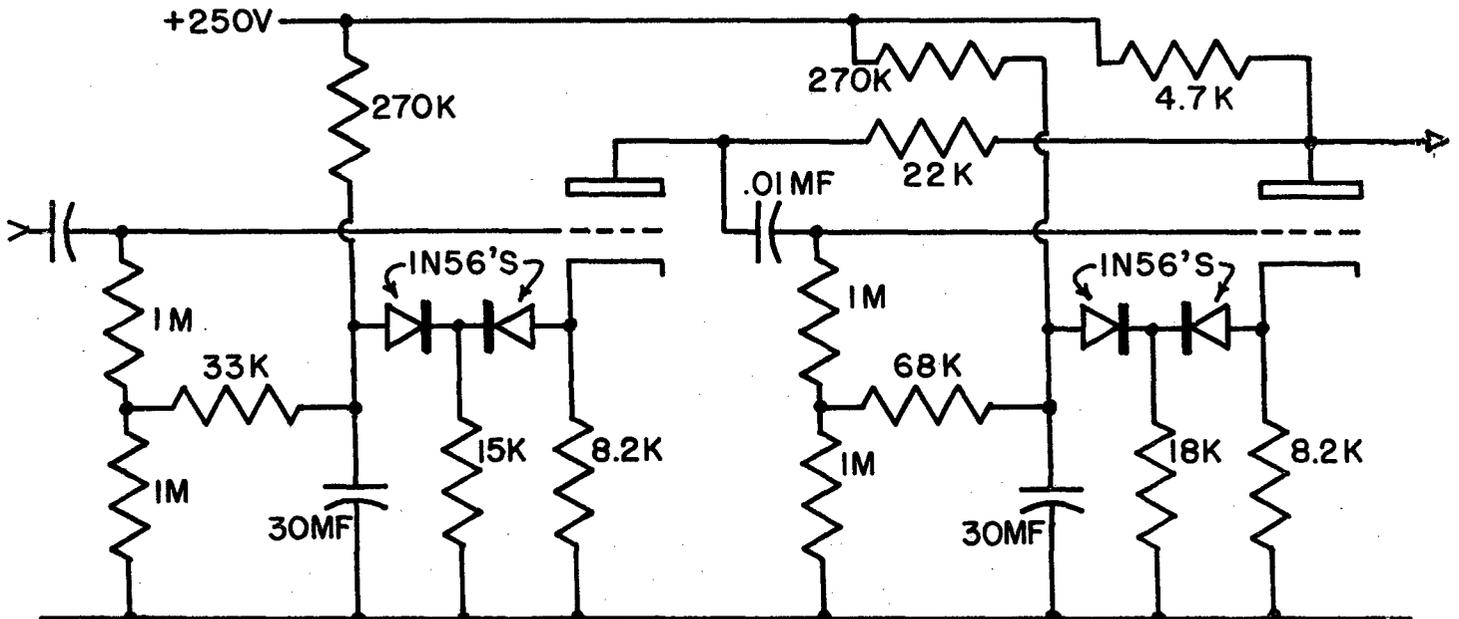


FIGURE 5.3 TYPICAL FEEDBACK PAIR WITH IAGC

ance with positive input pulses is as follows: Small-signal voltage gain,  $A_1 = 36$ . Output voltage at first transition,  $V_{T1} = 3$  volts. Gain above first transition,  $A_2 = 5.7$ . Output voltage at second transition,  $V_{T2} = 6.5$  volts. Gain above second transition,  $A_3 = 1.0$ . The performance above the first transition is slightly different with negative input pulses because no negative supply is used for diode bias. The relatively small resistors used between the diode junctions and ground thus cause the gains to be different for positive and negative signals, as discussed in Section 4.

In this circuit the first tube operates with a plate current of 3.5 ma, and the second with 4.5 ma. The quiescent diode currents are approximately 0.75 ma. Since the plate voltage for the first tube is supplied through the 22k feedback resistor, it operates at a relatively low voltage. In this design the first tube operates with a plate voltage of 135 volts, and a cathode voltage of 22.5 volts. The grid bias is about 0.75 volts. The plate voltage of the second stage is 212 volts, while the cathode voltage is 31 volts. The second tube bias is 2.0 volts. Under these conditions each tube has an amplification factor of about 60, and a plate resistance of approximately 15k.

To illustrate the insensitivity of the feedback pair to power supply ripple, a comparison between the circuit of Fig. 5.3 and that of Fig. 2.6 will be made. For this purpose the fraction of supply ripple appearing at the first plate of each type of amplifier will be calculated. For these computations the amplifiers will be assumed to be operating at their maximum (small-signal) gain, since this is the condition under which ripple is bothersome. First, for the basic circuit of Fig. 2.6, the effective plate resistance for the calculated operating conditions may be found using Fig. 2.5 and Eq. (2.5). This is found to be 36k. The fraction of available ripple which will appear at the plate will be determined by the voltage division between plate load and effective plate resistance. This is found to be 0.86. In the case of the feedback pair, the fraction of the total ripple voltage at the second plate is first found using Fig. 2.5 and Eqs. (2.5) and (5.5). The effective plate resistances of the two stages is found to be 32k. From (5.5) the output resistance of the second stage due to feedback is found to be 0.86k. The fractional ripple at the second plate

is found as above to be 0.155. The ripple on the first plate will be less than this because of the dividing action of the 22k feedback resistor and the 32k effective output resistance of the first tube. The fractional ripple at the first plate is thus found to be 0.095, so that the feedback pair gives an improvement of  $0.86/0.095 = 9.05$ , or 19.1 decibels.

The amplifier of Fig. 5.3 may be converted into a limiter by changing only two resistors. For this type of operation the feedback resistor is made 8.2k as indicated by (5.7). This changes the operating conditions of the first tube considerably, so that the resistor determining its grid bias must be changed. The original 3.5 ma plate current is obtained if the 33k bias resistor is raised to 68k.

## 6. CONSIDERATIONS RELATING TO COMPLETE AMPLIFIERS

Up to this point attention has been focused upon the problem of designing stages or pairs of stages to operate in a certain manner. There are a few points pertaining to the cascading of stages which should be mentioned at this time.

When several identical stages are cascaded, the increments of output voltage between successive transitions is not constant. This may be understood by considering an example. Let the small-signal gains be  $A_1$ , the stage output transition voltages be  $V$ , and the large-signal gains be  $A_2$ . As the input signal to the amplifier is increased from zero, the last stage will go through its gain transition at an amplifier output voltage of  $V$ . The output of the preceding stage at this point will be  $V/A_1$ . When its output increases to  $V$  the second transition will occur. The output of the preceding stage will thus increase by an amount  $V - V/A_1$  before the second transition occurs at the output, so that the output voltage between first and second transitions will be  $A_2V(1 - 1/A_1)$ . At the time the second transition occurs at the output, the output of the second from last stage is again  $V - V/A_1$ . The next increment of output voltage is thus  $A_2^2V(1 - 1/A_1)$ , etc.

In the special case of  $A_2 = 1$  which is frequently used in practice, the first increment at the output is  $V$ , and all remaining increments are of amount  $V(1 - 1/A_1)$ . When  $A_2$  is significantly different from unity,

the size of the output voltage increments will change progressively as the input signal is increased.

The same sort of increment variation will be present in the case of feedback pairs. Here the first two transitions are determined by the pair design. Assuming identical pairs, the output voltage increment between second and third transitions will be  $\left[ V_{T1} - V_{T1}/A_1 - (V_{T2} - V_{T1})/A_2 \right] A_3$ . The fourth increment will be  $\left[ V_{T2} - V_{T1} \right] A_3$ . The fifth increment will be the same as the third, except that  $A_3$  appears squared. Similarly, the seventh, ninth, etc., have  $A_3^3$ ,  $A_3^4$ , etc. The sixth, eighth, etc., output voltage increments will be the same as the fourth, except for  $A_3^2$ ,  $A_3^3$ , etc. terms.

In a typical case of identical pairs with equal increments between first and second transitions and unity large-signal gain,  $A_3$ , it is found that there is a repetitive variation in the size of the output increments. If, for example, the pair characteristics are  $A_1 = 36$ ,  $A_2 = 6$ ,  $A_3 = 1$ ,  $V_{T1} = 3$  volts, and  $V_{T2} = 6$  volts, the output voltage increments are found to be 3, 3, 2.4, 3, 2.4, 3. etc. If all but the last pair are altered to have output transition voltages of 3.6 volts and 6.6 volts, the output increments will all be 3 volts.

Another consideration relating to the cascading of several stages or pairs is that of over-driving in the later tubes. Difficulty of this sort may be encountered when the grid of the last tube of a high-gain chain is driven negative. If the signal at this point exceeds the quiescent cathode voltage, clipping will occur. This situation will be worst when feedback pairs are used because of the higher gain of the first stage, as illustrated in Fig. 5.3.

There are several ways to get around this difficulty. First, other considerations permitting, the size of the output voltage increments of all stages may be reduced. This, of course, results in a smaller amplifier output voltage for a given large input signal. Second, it may be possible to reverse the signal polarity so that the larger signals at the last grid are positive. Since the stage operates with a large unbypassed cathode resistor, very large positive signals may be applied without drawing grid current. Another possible solution when pairs are being used is to convert the last two stages to the basic circuit (no feedback). In typical

cases, this will reduce the signal at the last grid by a factor of two.

## 7. CONCLUSIONS

One of the most important aspects of the operation of the type of amplifier described here is its stability with respect to variations in the characteristics of the tubes and germanium diodes employed. Considerable difficulty from these sources was experienced in earlier circuits which involved diodes connected directly between the cathode of the tubes and ground. A "floating" arrangement is used here in which the voltage across a large condenser adjusts to changes in diode and tube characteristics. In practice, it has been found that tubes and germanium diodes may be selected at random with no appreciable change in circuit performance. The drop in the back-resistance of germanium diodes which occurs at high temperatures causes no appreciable change in circuit operation, since the diodes are effectively in parallel with the stage cathode resistors which are typically 10k or less.

APPENDIX A  
DESIGN RELATIONS FOR THE BASIC IAGC CIRCUIT

The standard expression for the voltage gain of a grounded-cathode triode stage is

$$A = \frac{\mu R_L}{R_L + R_P} \quad (\text{A.1})$$

When this is solved for  $R_L$ , the expression given in (2.1), without the factor of 2, results. Typically the degeneration due to the conducting diodes in the cathode circuit is such that the gain  $A$  is realized when the plate load resistor is doubled. The accuracy of (2.1) is dependent upon the quiescent diode current, the characteristics of the diodes, the tube parameters, etc., so that it is useful only for starting the design.

The simple relation given by Eq. (2.2) is usually of adequate accuracy for the practical determination of output voltage at the gain transition. The answer will always be somewhat low, since some change in the cathode potential is necessary to overcome the diode voltage drops. A correction term of

$$v_o R_L / R_K \quad (\text{A.2})$$

may be added to (2.2) to improve the accuracy. The actual change in cathode voltage is somewhat greater than  $v_o$  because of the increased current through the diode not being driven out of conduction. The slight under-correction realized with Eq. (2.3) compensates for the gradual decrease in gain before the actual transition voltage is reached, so that the output of the stage for large signals may be found with good accuracy if the output voltage at transition is determined from (2.3).

An equivalent circuit representation of a triode with cathode degeneration may be found using Thevenin's theorem. The circuit for signal frequencies is shown in Fig. A1, and the equivalent in Fig. A2.

The equivalent generator voltage is first found by assuming that the load ( $R_L$ ) is infinite. The current through the circuit is then zero, so that the open-circuit voltage is just  $-\mu V_g$ . The impedance seen from the output terminals is then found with the internal generator  $V_g$  shorted out. For this determination  $R_L$  in Fig. A2 is replaced by a zero-impedance generator, and the  $V_g$  term in the equivalent generator is dropped. The net driving

(5.8) may be derived in a direct manner. The first term,  $V_{T1}/R_L$  might be termed the useful part of the total diode current, since it sets the output transition. The additional current that must be supplied to the feedback resistor is given by  $V_{T1}(1 + 1/A_{21})/R_F$ . This current is greater than  $V_{T1}/R_F$  by  $1/A_{21}$  because of the action of the feedback. Thus, when there is unit signal voltage at the plate of the tube, there must be a signal of opposite polarity and magnitude  $1/A_{21}$  at the grid, to which the feedback resistor is connected. The gain here is the small-signal gain of the second stage. The last term  $v_{o1}/R_{k2}$  accounts for the increased tube current which comes from the cathode resistor  $R_{k2}$  in pulling the diode out of conduction.

The quiescent diode current required in the first stage may be specified in terms of the output transition voltages of the pair. This relation is given by (5.9). The first term represents the change in the tube current of the first stage up to the first output transition. The output voltage at the first plate will be  $V_{T1}/A_{21}$  when  $V_{T1}$  is reached, so that this portion of the current will be  $V_{T1}/(R_{in1}A_{21})$ . The second term represents the additional change in first-tube current necessary to provide a second output transition voltage of  $V_{T2}$ . This term is similar to the first, the incremental change in output voltage appearing in the numerator, and the input resistance and gain of the second stage in the denominator. Since the second stage is above its gain transition, the appropriate values  $R_{in2}$  and  $A_{22}$  are used.

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