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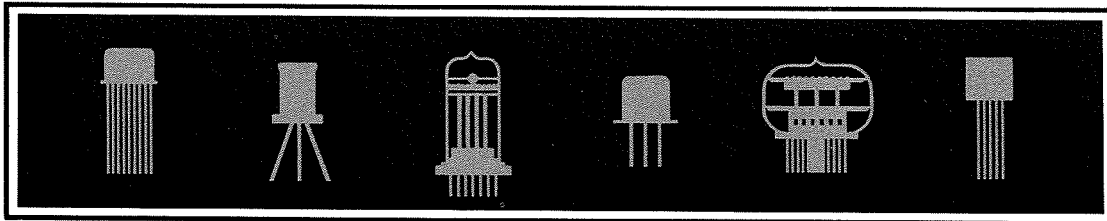
A PRACTICAL APPROACH TO TRANSISTOR AND VACUUM TUBE AMPLIFIERS

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DISPLAY DEVICES DEVELOPMENT



Two articles published in past issues of Service Scope contained information that, in our experience, is of particular benefit in analyzing circuits. The first article was "Simplifying Transistor Linear-Amplifier Analysis" (issue #29, December, 1964). It describes a method for doing an adequate circuit analysis for trouble-shooting or evaluation purposes on transistor circuits. It employs "Transresistance" concept rather than the complicated characteristic-family parameters. The second article was "Understanding and Using Thévenin's Theorem" (issue #40, October, 1966). It offers a step-by-step explanation on how to apply the principles of Thévenin's Theorem to analyze and understand how a circuit operates.

Now we present three articles that will offer a practical approach to transistor and vacuum-tube amplifiers based on a simple DC analysis. These articles will, by virtue of additional information and tying together of some loose ends, combine and bring into better focus the concepts of "Transresistance."

Part I
THE TRANSISTOR AMPLIFIER
INTRODUCTION

Tubes and transistors are often used together to achieve a particular result. Vacuum tubes still serve an important role in electronics and will do so for many years to come despite a determined move towards solid state circuits.

Whether a circuit is designed around vacuum tubes or transistors or both, it is important to recognize the fact that the two are in many ways complementary. It is wrong to divorce vacuum tubes and transistors as separate identities each peculiar to their own mode of operation. Indeed, as this series of articles will show

Let us consider the general equation for current through a P.N. diode junction.

$$I = I_o \left[\exp \frac{V}{\rho V_e} - 1 \right] \quad (1)$$

where $V =$ applied volts

$I_o =$ reverse bias current

$\rho =$ constant between 1 & 2

and $V_e = \frac{kT}{q}$ where $k =$ Boltzmann's

Const., 1.38×10^{-23} Joule/° Kelvin

$T =$ absolute temperature in degree Kelvin at room temperature, i.e., $T = 300^\circ K$

$q =$ electronic charge 1.602×10^{-19} Coulomb.

$$V_e = \frac{300}{11600} = 0.026 \text{ volts}$$

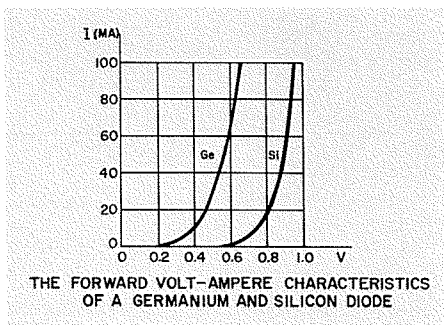


Figure 1.

Figure 1 shows a typical forward volt/amp characteristic for germanium and silicon diodes. Figure 2 is a plot of the collector current or the base current versus the base-to-emitter voltage of a transistor; point A on this curve is a typical operating point.

there is an analogy between the two. It is true of course, that the two are entirely different in concept; but, so often we come across a situation where one can be explained in terms of the other that it is very desirable to recognize this fact.

Transistor and vacuum tube data give us very little help in the practical sense. Parameter Curves and electrical data show the behavior of these devices under very defined conditions. In short, they are more useful to the designer than the technician. We are often reduced to explaining most circuits in terms of an ohms law approach;

OBJECTIVE

The objective of this paper is to present a practical approach to Transistor and Vacuum-tube amplifiers based on a simple DC analysis.

The articles will be published in the following sequence.

1. The Transistor Amplifier.
2. The Vacuum-tube Amplifier.
3. An analysis of a typical Tektronix hybrid circuit (Type 545B vertical) based on conclusions reached in (1) and (2).

As a corollary they will bring forward some important relationships between vacuum tubes and transistors.

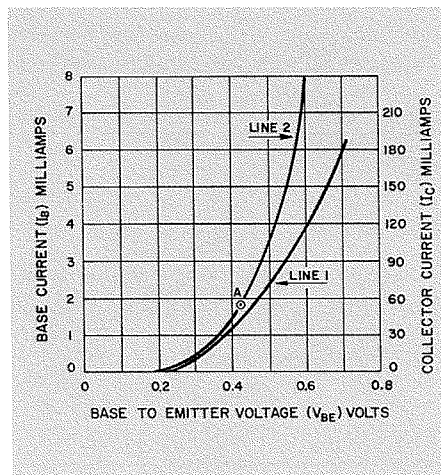


Figure 2. Line (1) is a plot of the base current versus the base-to-emitter voltage (V_{BE}). Line (2) is a plot of the collector current versus the base-to-emitter voltage (V_{BE}). Point "A" is a typical operating point.

so, it seems pointless not to pursue this approach to its logical conclusion.

In this first article we will look at a transistor amplifier as a simple DC model; and then, in the second article, look at a vacuum-tube amplifier in a similar light. We will assume that both devices are operated as linear amplifiers and then use the results in a practical way.

One must bear in mind that this approach cannot be assumed in all cases. It is, as it is meant to be, a simple analysis but the results will prove to be a valuable tool in trouble-shooting and understanding circuits.

One is quite justified in looking at a transistor in terms of the two-diode concept, refer to Figure 3. Therefore, assuming diode A to be forward biased and diode B to be reverse biased, as would be the case if we were to operate the transistor as a linear amplifier, the current through diode A will conform to equation (1). Let us take a closer look at Figure 2.

We define conductance in the general case as

$$g = \frac{I}{V}$$

and therefore at our operating point "A" the dynamic conductance

$$g' = \frac{\Delta I}{\Delta V} \quad (2)$$

hence

$$g' = \frac{I_o \exp \frac{V}{\rho V_e}}{\rho V_e} = \frac{I + I_o}{\rho V_e} \quad (3)$$

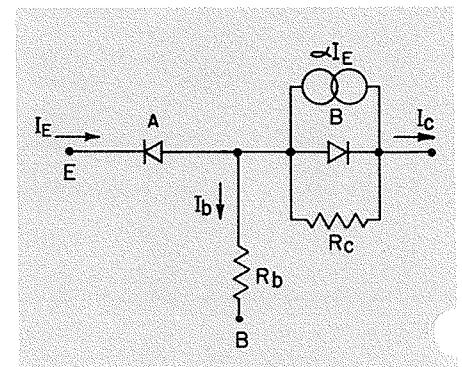


Figure 3. Illustration of the two-diode concept of a transistor.

but $I \gg I_0$ then $g' = \frac{I}{\rho V_e}$

or $g' = \frac{I}{0.026\rho}$ mhos (4)

The term "ρ" takes into account the recombination of carriers in the junction region. It is approximately unity for germanium and approximately 2 for silicon. At a typical operating point this term can usually be neglected. Therefore, we may say that

$g' = \frac{I}{26}$ if I is in milliamps. (5)

Now resistance is the reciprocal of conductance and therefore the value of conductance at point "A" can be given in terms of resistance

$r_e = \frac{26}{I}$ Ω's (6)

This resistance (r_e) is commonly known as the dynamic emitter resistance.

At this point we will depart from our simple model and look at the transistor in another form; but, bear in mind our first thoughts. Transistor parameters are derived from various equivalent circuits depending upon the configuration i.e., common emitter, common base, or common collector. We will not consider any detailed analysis in this approach; but, to understand the approach it is necessary to know how these parameters are derived. It will be simple enough to derive another set of parameters once we have our basic model constructed.

The simplest and easiest equivalent circuit of a transistor is the "Tee" equivalent. It is a very good approximation about the behavior of a transistor, especially at DC and low frequencies. We can also represent either the common emitter or the common base simply by interchanging R_b and R_e . Figure 4 is a "Tee" equivalent circuit of

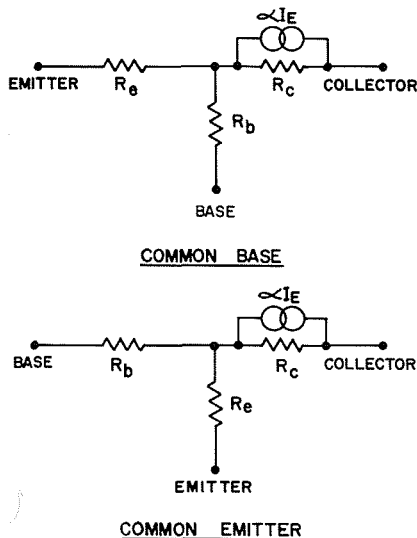


Figure 4. "Tee" equivalent circuits for the common-base and common-emitter configurations of transistors.

the common emitter and the common base configurations.

Firstly, let us define the term β (the small-signal current gain) as

$\beta = \frac{\Delta I_c}{\Delta I_b}$ (7)

and since $I_E = I_c + I_b$

then $I_E = I_c (1 + \frac{1}{\beta})$ (8)

usually $\beta \gg 1$ then $I_E \approx I_c$

Equation (8) shows us that only $\frac{1}{\beta}$ of the emitter current flows into the base. Hence, it is reasonable to suppose that any impedance in the emitter, when viewed from the base, will be β times as great; and, any impedance in the base, when viewed from the emitter, will be β times as small. That is to say, the dynamic resistance multiplied by β must equal R_e in our equivalent "Tee" circuit.

Hence $R_e = \beta R_b$

Our equivalent circuit shows a resistance R_b . This resistance is known as the base-spreading resistance. It is a physical quantity and can be expressed in terms of resistivity associated with the base-emitter junction. It can vary between a few ohms to hundreds of ohms, depending upon the type of transistor; and therefore, must be taken into consideration. Looking into the emitter we see it as an impedance whose value is divided by β and appears in series with the dynamic emitter resistance (r_e). Hence the emitter current encounters an impedance in the base/emitter junction which is equal to the sum of the dynamic resistance plus $\frac{R_b}{\beta}$, the latter term we will designate R_r and the sum of these two resistances we will designate R_t .

Hence $R_t = r_e + R_r$ (9)

The value of R_r can vary anywhere between 2Ω to 24Ω depending on the value of R_b . R_b is difficult to measure and rarely given in electrical data on transistors. A figure of 250Ω 's is a typical value at low frequencies. Therefore, if β were 50 then R_r would be 5Ω 's.

Now if we look into the base in the common emitter or the common collector configuration it is reasonable to suppose we will see the resistance (R_t)—plus any other impedance which may be wired to the emitter terminal—multiplied by β , then

$R_{in} = \beta(R_t + R_E)$ (10)

where R_E = the external emitter resistance.

If $R_E \gg R_t$ then $R_{in} = \beta R_E$

So far we have had very little to say about R_e shunted by the current generator $\propto I_E$. If our equivalent "Tee" circuit con-

sisted of resistances alone, it would be passive; i.e., it could supply no energy of its own. But a transistor can amplify energy to the signal. To represent this we have shown a current generator shunting R_e . The value of R_e will depend on the circuit configuration; i.e., tens of kilohms for a common emitter configuration, to many megohms for a common base configuration. In our approach it is not necessary to pursue this matter any further since we will not be considering a transistor in any extreme condition.

Now in a more practical sense, let us look at Figure 5, a typical common-emitter configuration.

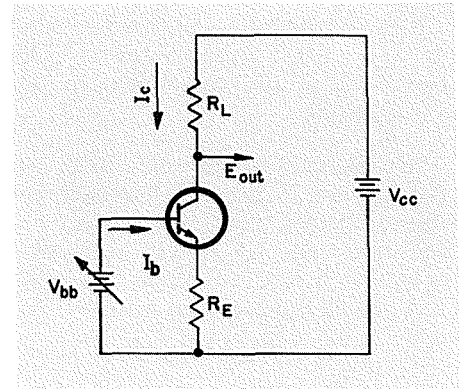


Figure 5. A typical common-emitter circuit.

Now we will assume $R_e \gg R_L$.

Now by inspection

$E_{out} = V_{cc} - I_c R_L$ (11)

hence $\Delta E_{out} = -\Delta I_c R_L$ (12)

The input impedance we see looking into the base of a transistor in the common emitter configuration is

$R_{in} = \beta(R_E + R_t)$ (10)

also $\Delta I_b = \frac{\Delta V_{bb}}{R_{in}}$
 $= \frac{\Delta V_{bb}}{\beta(R_E + R_t)}$ (13)

we also recall that

$\beta = \frac{\Delta I_c}{\Delta I_b}$ (7)

hence $\Delta I_c = \beta \Delta I_b$ (14)

Therefore substituting equation (13) in equation (14)

$\Delta I_c = \beta \frac{\Delta V_{bb}}{\beta(R_E + R_t)}$ (15)

and from equation (15)

$\Delta V_{bb} = \Delta I_c (R_E + R_t)$ (16)

we define the voltage gain as

$$A(v) = \frac{\Delta E_{out}}{\Delta V_{bb}}$$

Then from equation (12) and equation (16)

$$A(v) = - \frac{\Delta I_c R_L}{\Delta I_c (R_E + R_t)}$$

$$= - \frac{R_L}{R_E + R_t} \quad (17)$$

and if $R_E \gg R_t$ then

$$A(v) = - \frac{R_L}{R_E} \quad (18)$$

If we analyze the common-base configuration in a similar manner we arrive at the same result with the one exception that the sign is positive.

The conclusion we can draw from this analysis is that the gain of a transistor stage is set by external conditions provided that the emitter resistance is sufficiently great enough to "swamp" our internal resistance (R_t). In the absence of an emitter resistance

$$A(v) = \frac{R_L}{R_t}$$

There is one very important fact we must remember about R_E . R_E will be that impedance in which the signal current will flow to the AC ground. We define an AC ground point as that point in a circuit at which the power level of the signal has been reduced to zero.

We normally encounter three types of an AC ground:

1. An Actual AC Ground.

This is the chassis point or the DC ground point. It is as well to remember the

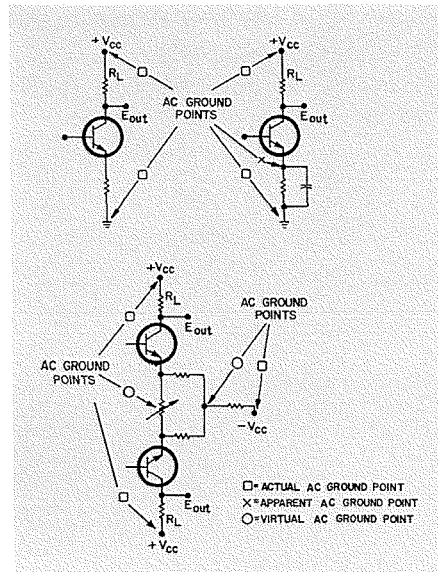


Figure 6. Illustrating the three types of AC ground normally encountered in electronic circuits.

power supply can be placed in this category so far as the signal is concerned.

2. An Apparent AC Ground.

The apparent AC ground may be represented by any point in a circuit which acts as to represent a low impedance between that point and the actual AC ground thereby bypassing the signal to an actual AC ground. A large value capacitor is a typical example should one side be returned to an actual AC ground.

3. The Virtual AC Ground.

The virtual AC ground point is perhaps the most difficult to recognize. It may best be explained as that point in a circuit where we have two signals of equal amplitude and frequency but exactly opposite in phase. Figure 6 will help clarify these points.

Figure 8 summarizes the results of our DC analysis of the common emitter, common base and common collector.

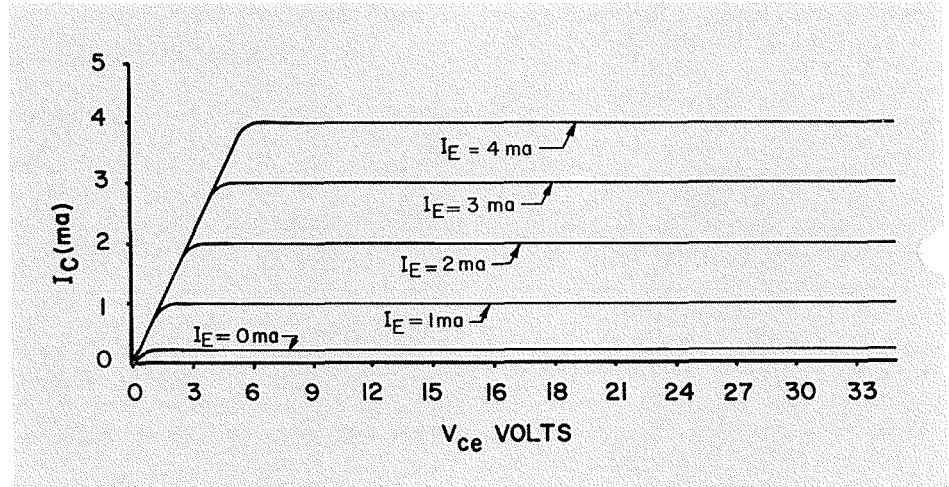


Figure 7. We define the parameter R_c in the common-base "Tee" configuration as;

$$R_c = \frac{\Delta V_{ce}}{\Delta I_c} \Bigg|_{I_E} \text{ ohms}$$

Where ΔV_{ce} is the change in the collector voltage because of the change in collector current ΔI_c , when we hold the emitter current I_E constant.

Once the collector becomes saturated, the change in I_c is very small for a large change in V_{ce} . Hence, R_c is a very large resistance and does not modify the DC equivalent circuit to any extent. For this reason it was omitted from Figure 8. Therefore; $R_{out} = R_L$ (Common Base).

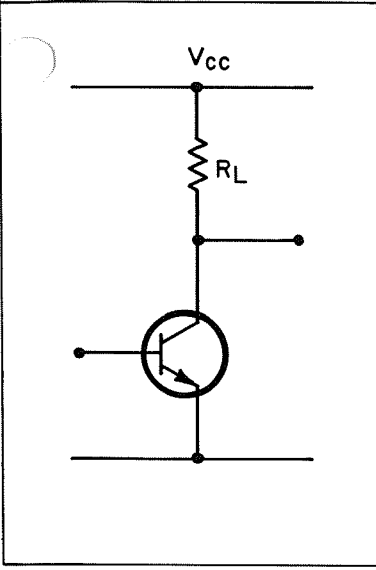
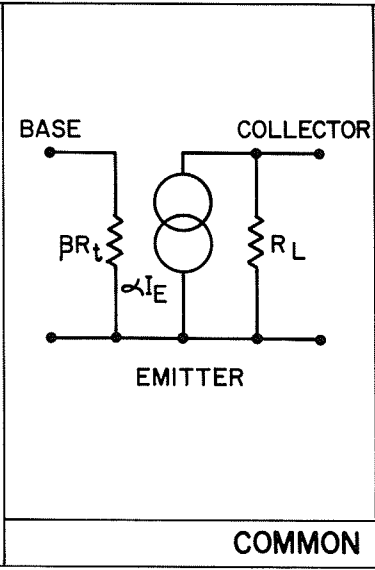
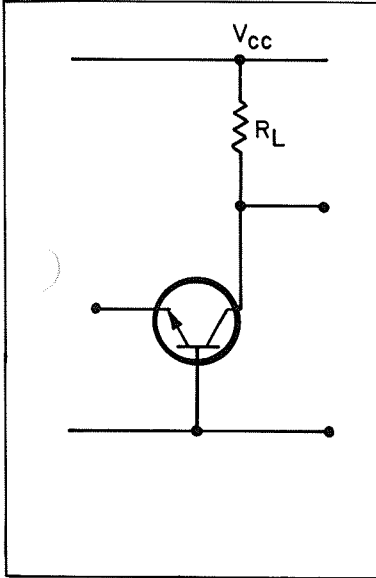
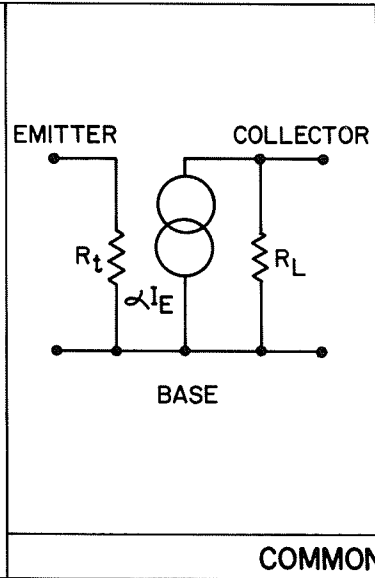
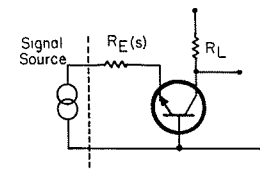
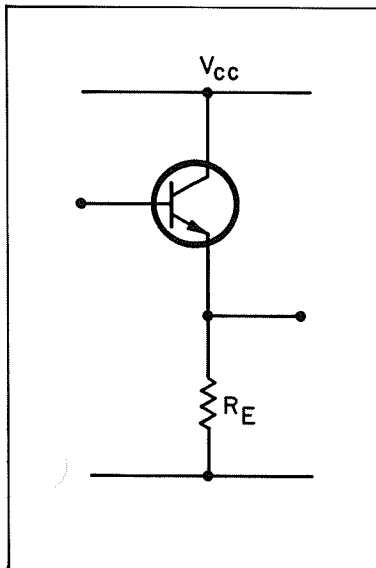
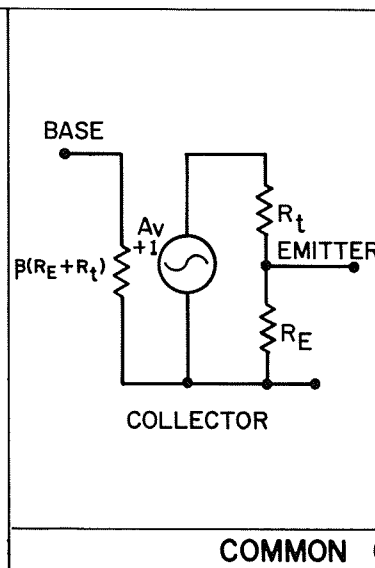
BASIC CIRCUIT	EQUIVALENT CIRCUIT	EQUATIONS	REMARKS
		<p>GAIN = $A(v) = -\frac{R_L}{R_t}$</p> <p>RIN = βR_t</p>	<p>A resistance (R_E) between the emitter terminal of the transistor to the AC ground will modify the gain equation and the input impedance; then,</p> <p>$A(v) = -\frac{R_L}{R_t + R_E}$ and</p> <p>$R_{in} = \beta (R_t + R_E)$.</p>
COMMON EMITTER			
		<p>GAIN = $A(v) = +\frac{R_L}{R_t}$</p> <p>RIN = R_t</p> <p>ROUT = R_L</p>	<p>The equivalent resistance $R_{E(s)}$ between the input signal source and the emitter terminal of the transistor will modify the gain equation and the input impedance as seen from the signal source; then,</p> <p>$A(v) = +\frac{R_L}{R_t + R_{E(s)}}$ and</p> <p>$R_{in} = R_t + R_{E(s)}$.</p> 
COMMON BASE			
		<p>GAIN = $A(v) = \frac{R_E}{R_E + R_t}$</p> <p>RIN = $\beta (R_E + R_t)$</p> <p>ROUT = $(R_t + \frac{1}{\beta})$ in parallel with R_E</p>	<p>The actual value of R_{out} will depend on what resistance is connected to the base. Let us assume the base is directly coupled to the preceding stage. The equivalent output impedance of the preceding stage becomes the numerator over beta in the second term in the parenthesis and the output impedance of the stage under consideration R_{out} is modified accordingly; e.g., if the output impedance of the previous stage is 1000Ω, then</p> <p>$R_{out} = (R_t + \frac{1000}{\beta})$ in parallel with R_E.</p>
COMMON COLLECTOR			

Figure 8.

**PART 2
THE VACUUM TUBE AMPLIFIER**

In the previous article (Part I, "The Transistor Amplifier") of this series, it was shown that the gain of a linear transistor amplifier is set by external conditions. The same reasoning can also be applied to vacuum tubes. The equivalent circuit of a vacuum-tube amplifier is shown in Figure 9. The current that is produced in the plate circuit by the signal (E_g) acting on the grid is taken into account by postulating that the plate circuit can be replaced by a generator, $-\mu E_g$ having an internal resistance (r_p). We may also consider a vacuum-tube amplifier in terms of the constant-current form by replacing the voltage generator in the constant-voltage form with a current generator ($g_m E_g$) shunting the internal resistance (r_p).

These two approaches are valid in every respect but they do not convey much to us in the practical sense. Let us now consider a vacuum-tube amplifier from another approach.

In an amplifier which has its grid referenced to ground all plate-circuit impedances, R_L and r_p , when viewed from the cathode are multiplied by the term

$$\frac{1}{\mu + 1}$$

cathode impedances when viewed from the plate circuit are multiplied by the term ($\mu + 1$). Therefore, the impedance we see looking into the cathode must be

$$\frac{r_p + R_L}{\mu + 1}$$

where μ equals the amplification factor of the tube.

Hence it is reasonable to suppose that the voltage E_c , reference Figure 10, appears across this impedance we see looking into the cathode.

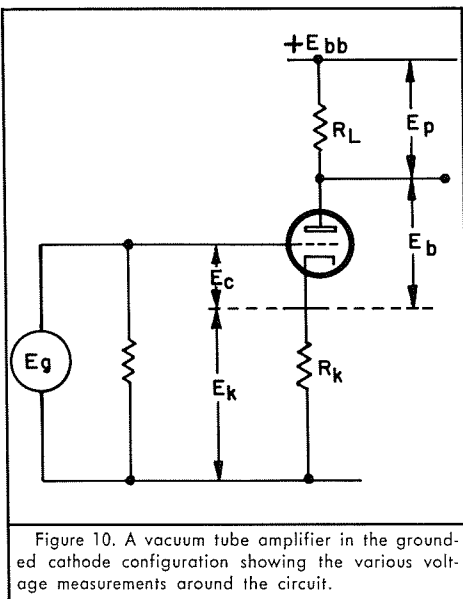


Figure 10. A vacuum tube amplifier in the grounded cathode configuration showing the various voltage measurements around the circuit.

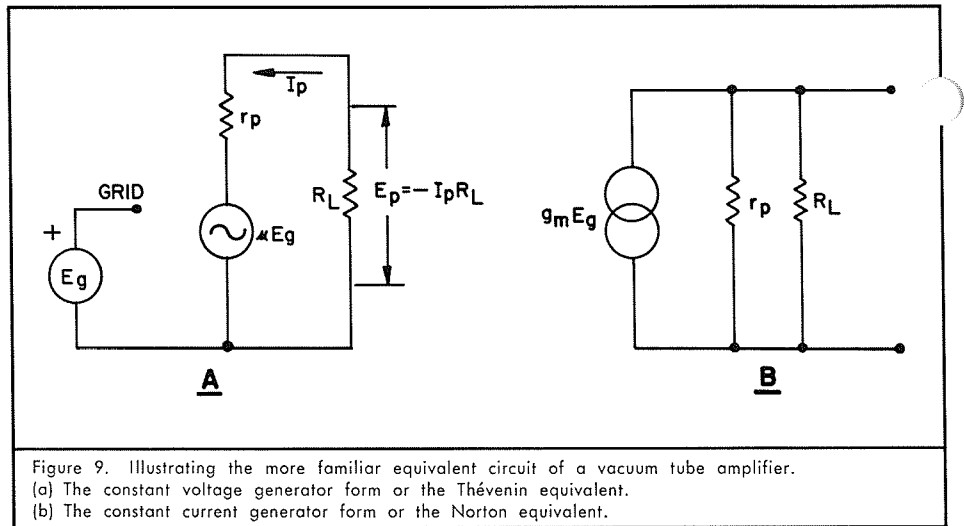


Figure 9. Illustrating the more familiar equivalent circuit of a vacuum tube amplifier. (a) The constant voltage generator form or the Thévenin equivalent. (b) The constant current generator form or the Norton equivalent.

The Triode Amplifier (Grounded Cathode)

We will now look at a triode amplifier in terms related to our equivalent circuit. The common component is of course, the plate current. The change in this current due to the action of a control grid will determine the output voltage across the load impedance (R_L).

$$\text{Now } E_g = E_c + E_k \quad (19)$$

That is to say

$$E_g = I_p \left[\frac{r_p + R_L}{\mu + 1} \right] + I_p R_k$$

$$\text{Or, } E_g = I_p \left[\left(\frac{r_p + R_L}{\mu + 1} \right) + R_k \right] \quad (20)$$

$$\text{Also, } E_{bb} = E_b + E_p + E_k \quad (21)$$

$$\text{or } E_b = E_{bb} - E_p - E_k \quad (22)$$

$$\text{and } E_p = -I_p R_L \quad (23)$$

We define the voltage gain $A_{(v)}$ as

$$A_{(v)} = \frac{E_p}{E_g} \quad (24)$$

$$\begin{aligned} \text{Then } A_{(v)} &= - \frac{I_p R_L}{I_p \left[\left(\frac{r_p + R_L}{\mu + 1} \right) + R_k \right]} \\ &= - \frac{R_L}{\left(\frac{r_p + R_L}{\mu + 1} \right) + R_k} \quad (25) \end{aligned}$$

We now have arrived at an equation for gain which is a ratio of impedances. The same approach may be applied to the grounded-grid configuration and we arrive at a similar result, except the sign is positive.

The Pentode Amplifier

In the triode amplifier all the cathode current will flow through the output load impedance (R_L). However, in the case of the pentode and other multigrad tubes, some of this current is diverted into the screen. Equation (23) defines the output voltage

in terms of the plate current. Therefore, to derive the actual gain figure we must determine the actual amount of cathode current which will finally reach the plate and become signal current. This figure can be arrived at from a graphical analysis of the mutual-conductance curves. In most cases, about 72% of the cathode current reaches the plate to become signal current. A typical example is a type 12BY7 pentode. However, this figure can be as high as 90% for some types—for example a 7788 pentode. The ratio of the plate current (I_p) to the cathode current (I_k) is η .

$$\text{plate efficiency factor, i.e., } \eta = \frac{I_p}{I_k}$$

Now let us reexamine what effect this fact must have on the gain of a pentode amplifier as compared to a triode amplifier. The impedance we see looking into the cathode of a pentode is the same as for a triode.

$$\text{That is } \frac{r_p + R_L}{\mu + 1}$$

however $r_p \gg R_L$ and therefore R_L can usually be neglected in this equation.

$$\text{That is to say } \frac{r_p}{\mu + 1} \approx \frac{1}{g_m}$$

and since conductance is the reciprocal of resistance we will call this impedance r_k .

$$\text{i.e. } r_k = \frac{1}{g_m} \quad (26)$$

We have seen that the gain equation of the triode amplifier is defined in terms of the parameters μ and r_p . We should not lose sight of the fact that μ and r_p are related to the plate current and therefore when these parameters are transferred to cathode dimensions these terms must be multiplied by the plate efficiency factor (η). That is to say the impedance we see looking into the cathode r_k must be multiplied by (η). With these facts in mind let us

now derive the gain equation for a pentode amplifier.

We recall that:

$$E_p = E_{bb} - E_p - E_k \quad (22)$$

$$\text{and } E_p = -I_p R_L \quad (23)$$

$$\text{also } E_g = E_c + E_k \quad (19)$$

$$= \eta r_k I_k + I_k R_k \quad (27)$$

$$\text{but } I_k = \frac{I_p}{\eta} \quad (28)$$

Therefore substituting equation (28) in equation (27)

$$\begin{aligned} E_g &= \frac{\eta r_k I_p}{\eta} + \frac{I_p R_k}{\eta} \\ &= I_p \left(r_k + \frac{R_k}{\eta} \right) \end{aligned} \quad (29)$$

and since the voltage gain

$$\begin{aligned} A_{(v)} &= \frac{E_p}{E_g} \quad (24) \\ &= - \frac{I_p R_L}{I_p \left(r_k + \frac{R_k}{\eta} \right)} \\ &= - \frac{R_L}{r_k + \frac{R_k}{\eta}} \end{aligned} \quad (30)$$

The same remarks we made about the external emitter resistor R_E (refer to Part No. 1, The Transistor Amplifier) apply equally as well to the cathode resistor, R_k ; namely, R_k will be that impedance in which the signal current will flow to the AC ground.

In the case of the grounded plate (the cathode follower) we do not need to consider the plate efficiency factor if the amplifier is triode connected, therefore, the "gain" can be considered in terms of a simple divider network which can never be greater than unity.

$$A_{(v)} = \frac{R_k}{R_k + r_k} \quad (31)$$

The Push-Pull Amplifier

We can view a push-pull amplifier in a similar light by recognizing the existence of a virtual AC ground point between the cathodes of $V_{(1)}$ and $V_{(2)}$ as shown in Figure 11. Therefore, the gain of a push-pull triode amplifier will be:

$$A_{(v)} = \frac{R_{L(1)} + R_{L(2)}}{r_{k(1)} + r_{k(2)} + R_{k(1)} + R_{k(2)}} \quad (32)$$

where subscripts (1) and (2) are associated with $V_{(1)}$ and $V_{(2)}$.

And if:

$$R_{k(1)} = R_{k(2)}$$

$$\text{and } r_{k(1)} = r_{k(2)}$$

which is usually the case; then,

$$A_{(v)} = \frac{R_{L(1)} + R_{L(2)}}{2r_k + 2R_k} \quad (33)$$

$$\text{Where } r_k = \frac{r_p + R_L}{\mu + 1} \quad (\text{either } V_{(1)} \text{ or } V_{(2)})$$

$$\text{and } R_k = R_{k(1)} \text{ or } R_{k(2)}$$

With a push-pull pentode amplifier we must consider the plate-efficiency factor (η). Therefore,

$$A_{(v)} \text{ pentode} = \frac{R_{L(1)} + R_{L(2)}}{2r_k + \frac{2R_k}{\eta}} \quad (34)$$

$$\text{where } r_k = \frac{1}{gm} \text{ either } V_{(1)} \text{ or } V_{(2)}$$

$$R_k = R_{k(1)} \text{ or } R_{k(2)}$$

η = plate-efficiency factor of either $V_{(1)}$ or $V_{(2)}$.

The Cascode Amplifier

The cascode amplifier fundamentally consists of two tubes connected in series, see Figure 12. Normally we usually fix the grid of $V_{(1)}$ at some positive voltage.

The key to understanding this type of circuit is to consider $V_{(2)}$ as a voltage-activated current generator. All the current delivered by $V_{(2)}$ passes through the output load impedance R_L . Any change in voltage appearing at the grid of $V_{(2)}$ appears as a change in current across R_L . We can derive the gain equation in the same way as we did for a pentode amplifier. There is no need to consider (η) if both tubes are triodes.

$$A_{(v)} \text{ (stage)} = \frac{R_{L(1)}}{R_{k(2)} + r_{k(2)}} \quad (35)$$

$$\begin{aligned} \text{where } r_{k(2)} &= \frac{r_{p(2)}}{\mu_{(2)} + 1} \\ &= \frac{1}{gm_{(2)}} \end{aligned}$$

where the subscripts (1) and (2) are associated with $V_{(1)}$ and $V_{(2)}$.

One of the advantages of this type of circuit is that the internal impedance which shunts R_L is extremely high.

In this respect the triode cascode amplifier closely approximates a pentode amplifier. If we compare the plate-current versus plate-voltage curves of both devices we see a close resemblance.

The Hybrid Cascode Amplifier

Figure 13 is a typical configuration consisting of a vacuum tube V_1 and a transistor, Q_1 , connected in series. We can apply much the same approach as we did for the cascode vacuum-tube amplifier. Let us assume the base to emitter junction of Q_1 to be forward biased. The collector current of Q_1 becomes the plate current of V_1 . Therefore, any change occurring at the base of Q_1 is reflected as a change in plate current in V_1 .

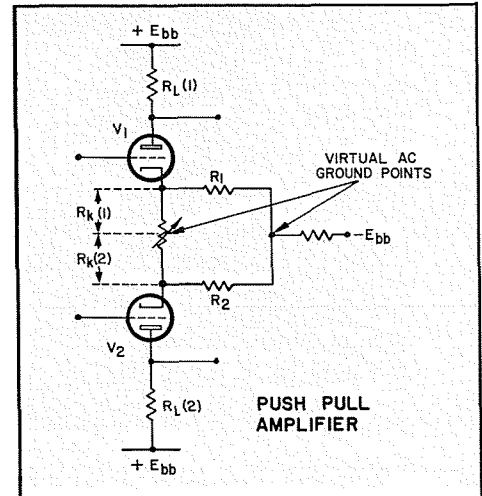


Figure 11. A typical push-pull triode amplifier. We normally encounter two virtual AC ground points between the cathodes of V_1 and V_2 . It may be necessary to consider the effect of the virtual AC ground point at the junction of R_1 and R_2 . If R_1 or R_2 is large in value compared respectively to $R_{k(1)}$ or $R_{k(2)}$ then we can neglect this virtual AC ground and consider R_k in terms of $R_{k(1)}$ or $R_{k(2)}$. However, if this is not so, R_k will be the parallel combination of $R_{k(1)}$ and R_1 or $R_{k(2)}$ and R_2 .

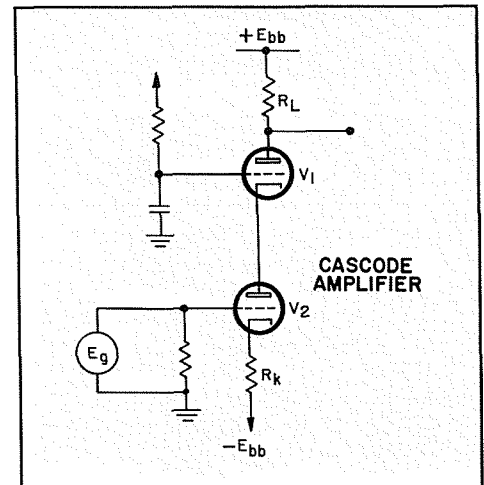


Figure 12. Illustrating a cascode amplifier using two triodes.

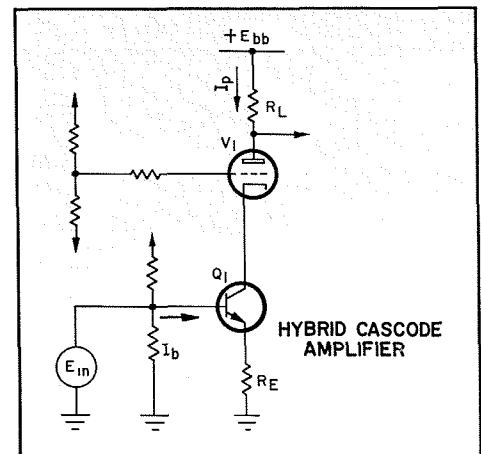


Figure 13. A typical hybrid cascode amplifier using a transistor and a vacuum tube.

BASIC CIRCUIT (TRIODE)	EQUIVALENT CIRCUIT	EQUATIONS
		$\text{GAIN} = A(v) = - \frac{R_L}{\frac{r_p + R_L}{\mu + 1} + R_k}$
GROUNDING CATHODE		

		$\text{GAIN} = A(v) = + \frac{R_L}{\frac{r_p + R_L}{\mu + 1} + R_k}$ $R_{IN} = R_k + \frac{r_p + R_L}{\mu + 1}$
GROUNDING GRID		

		$\text{GAIN} = A(v) = + \frac{R_k}{R_k + r_k}$ $R_{OUT} = r_k \text{ in parallel with } R_k$ $\text{WHERE: } r_k = \frac{r_p}{\mu + 1}$
GROUNDING PLATE		

PENTODE AMPLIFIER	PUSH PULL AMPLIFIER	CASCODE AMPLIFIER	HYBRID CASCODE AMPLIFIER
$\text{GAIN} = A(v) = \frac{R_L}{r_k + \frac{R_k}{\eta}}$ <p>WHERE:</p> <ul style="list-style-type: none"> R_L = LOAD RESISTANCE $r_k = \frac{1}{g_m}$ R_k = CATHODE RESISTOR (Refer Text) η = PLATE EFFICIENCY FACTOR 	<p>TRIODE PAIR</p> $\text{GAIN} = A(v) = \frac{R_{L(1)} + R_{L(2)}}{2r_k + 2R_k}$ <p>WHERE:</p> $r_k = \frac{r_p + R_L}{\mu + 1}$ <p>PENTODE PAIR</p> $\text{GAIN} = A(v) = \frac{R_{L(1)} + R_{L(2)}}{2r_k + \frac{2R_k}{\eta}}$ <p>WHERE: $r_k = \frac{1}{g_m}$</p> <p>SUBSCRIPTS (1) AND (2) ARE ASSOCIATED WITH V_1 AND V_2</p>	$\text{GAIN} = A(v) = \frac{R_{L(1)}}{r_k(2) + R_{k(2)}}$ <p>WHERE:</p> $r_k = \frac{1}{g_m(2)}$	$\text{GAIN} = A(v) = \frac{R_L}{R_E + R_f}$ <p>WHERE:</p> <ul style="list-style-type: none"> R_L = LOAD RESISTANCE * $R_f = r_e + R_r$ * R_E = EXTERNAL EMITTER RESISTANCE * REFER PART I "THE TRANSISTOR AMPLIFIER"

Figure 16

We recall (Part 1, The Transistor Amplifier, Eq. 10) that the input impedance we see looking into the base of a transistor the common-emitter configuration is:

$$R_{in} = \beta (R_E + R_t) \quad (10)$$

Now $E_{in} = I_b R_{in}$

$$= I_b \beta (R_E + R_t) \quad (36)$$

$$\text{also } \beta = \frac{I_c}{I_b}$$

$$\text{or } I_c = \beta I_b \quad (37)$$

therefore substituting equation (37) in equation (36)

$$E_{in} = I_c (R_E + R_t) \quad (38)$$

now the collector current I_c becomes the plate current of V_1 . Then,

$$E_{in} = I_p (R_E + R_t) \text{ since } I_p = I_c \quad (39)$$

$$\text{also } E_p = -I_p R_L \quad (23)$$

and since

$$A_{(v)} \text{ (stage)} = \frac{E_p}{E_{in}}$$

then from equations (23) and (39)

$$\begin{aligned} A_{(v)} \text{ (stage)} &= - \frac{I_p R_L}{I_p (R_E + R_t)} \\ &= - \frac{R_L}{R_E + R_t} \quad (40) \end{aligned}$$

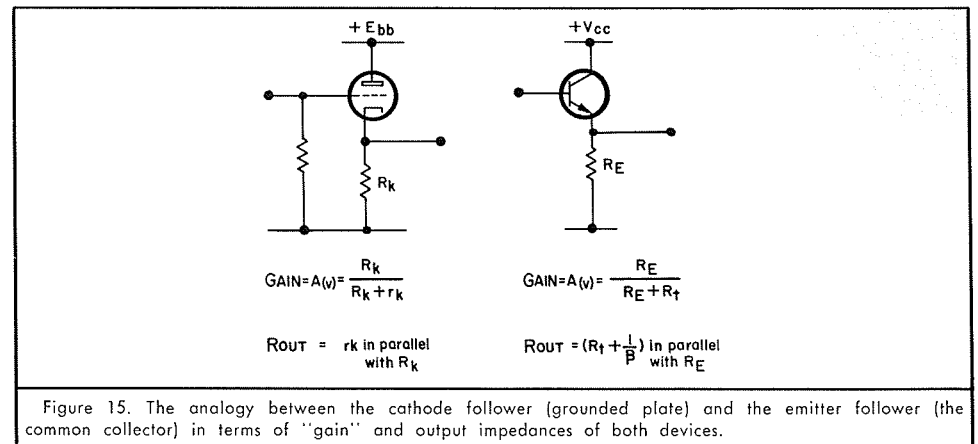
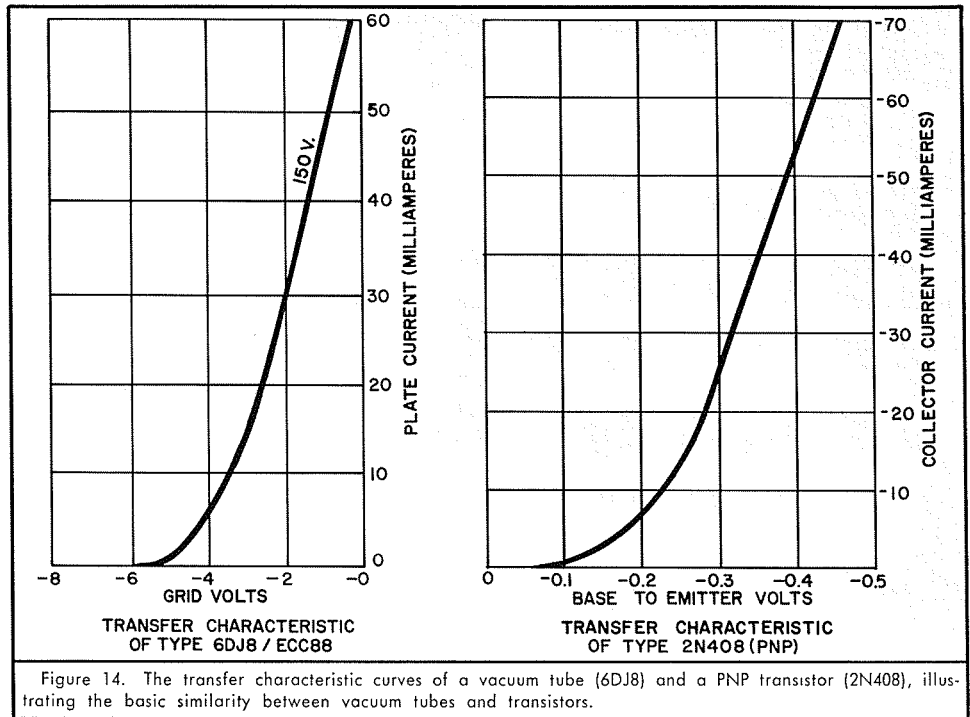
If the vacuum tube is not a triode but some other multigrad tube such as a pentode, the gain equation will have to be multiplied by the plate efficiency factor (η).

The same remarks concerning the output impedance of the vacuum-tube cascode amplifier can be applied to the hybrid counterpart.

Summary

We have shown that the gain of a linear amplifier, transistor or vacuum tube, is a ratio of impedances. We can, of course, derive the gain equations for both devices in terms of mutual conductance. In fact, if we compare the transfer curves of both devices, Figure 14, we see a striking similarity. V_{DB} and E_c can be thought of in the same terms and in like manner I_p and I_c perform identical functions. Our analysis of both devices has shown that this fact is not coincidence.

It is not unreasonable to say that when we compare the cathode-follower (grounded-cathode) against the common-collector configuration, Figure 15, we can think of both devices as being identical in operation—differing only in concept. The same argument can be put forward about the com-



mon-base amplifier and the grounded-grid amplifier. So too, the common-emitter amplifier and the grounded-cathode amplifier if we chose to ignore the input impedances of both devices.

Figure 16 summarizes the results of our analysis of the grounded cathode, grounded grid, and grounded plate amplifiers.

It is not surprising we sometimes find ourselves explaining one device in terms of another. Nature has a charming way of making most things interdependent upon one another. Recognize this fact and most tasks become a little easier.

PART 3 A DC ANALYSIS OF A TYPICAL TEKTRONIX HYBRID CIRCUIT

As a typical example of a Tektronix, Inc. hybrid circuit on which to demonstrate our DC analysis, we have chosen the vertical amplifier of a Type 545B Oscilloscope. This circuit is representative of the hybrid circuit one encounters so often in electronic instrumentation today.

The Type 545B vertical amplifier is a hybrid push-pull amplifier operating in a class A mode. It incorporates a few extra circuits such as trigger pick-off amplifiers necessary to accomplish its function, but, basically it is a hybrid push-pull amplifier.

To begin our analysis of the amplifier, the first thing we must do is select a portion of the amplifier circuit which will give us the information necessary for us to make our first calculation. We are going to analyze the whole circuit so we can choose our point of entry. The input circuit is as good a point as any. Bear in mind that, for our purpose, this is not the only point of entry. Any point on the circuit which will give us useful information would do.

A quiescent DC voltage of +67 volts is the nominal voltage at the output of the plug-in amplifiers used in the Type 545B oscilloscope. This voltage appears at terminals 1 and 3 of J11 in Figure 17, and thus, at the grids of V494A and V494B, a 6DJ8 dual triode. The input cathode follower (V494 A & B) has a bias of about 4 volts; therefore, both cathodes will be at +71 volts. The base voltage of Q514 and Q524 is then fixed at 71 volts. This sets the emitter voltages of Q514 and Q524 at one junction drop more negative (they are both NPN transistors) than the base. Therefore, the voltage at the emitter of Q514 and Q524 is 70.5 volts. T500 is a small toroidal transformer used for high-frequency common-mode rejection. The DC BALANCE Control, R495, sets the quiescent condition. We mean by this that the trace is centered.

We have made certain assumptions about the bias of a vacuum tube and the base-to-emitter voltage drop of a transistor. This is quite justifiable since we know what function the device performs. One helpful hint about transistors is that you can expect a base-to-emitter voltage drop of about 0.5 to 0.6 volts for a silicon transistor and about 0.2 volts for a germanium transistor.

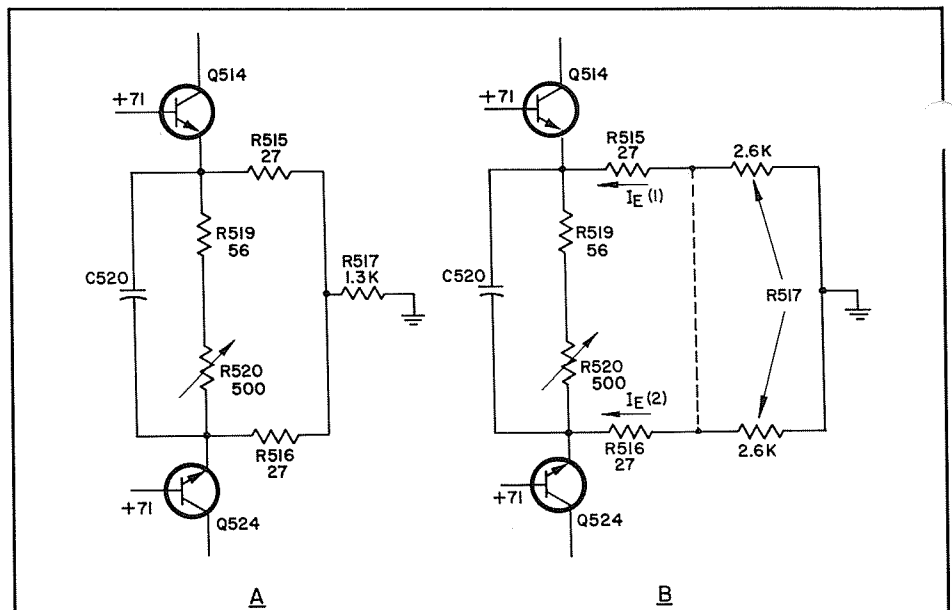


Figure 18. The circuit which will determine the DC emitter currents for either Q514 or Q524. (A)—The actual circuit as shown in Figure 17. (B)—The equivalent DC circuit considering R517 as two resistors through which the individual emitter currents will flow.

We are now able to calculate the emitter current of either Q514 or Q524. The DC-emitter current will flow through R515 or R516 and into R517 to ground. Since the emitter currents of Q514 and Q524 both pass through R517, we may think of R517 being made up of two resistors, each of 2.6 k Ω in value, in which the individual emitter currents will flow, refer to Figure 18: Therefore,

$$I_E (1) \text{ or } (2) = \frac{70.5 \times 10^3}{2.627 \times 10^3} \text{ mA}$$

$$= 27 \text{ mA}$$

We can now calculate the value of r_e , the dynamic-emitter resistance,

$$r_e = \frac{26}{I_E} = \frac{26}{27}$$

$$= 0.96 \Omega$$

to this we can add our constant, R_r , of say, 4 Ω . We recall that:

$$R_t = r_e + R_r \quad (9)$$

therefore:

$$R_t = 0.96 + 4 = 4.96 \Omega$$

or approximately 5 Ω . We have now established the value of the emitter current and the value of R_t for Q514 and Q524.

Our next step is to find the value of R_E . We must know this value in order to calculate gain. You will recall that R_E will be that impedance through which the signal current will flow to the AC ground. Let us take another look at the resistive network between the emitters of Q514 and Q524. The signal currents flowing in this circuit will be equal and opposite at two points, refer to Figure 19. These points are virtual AC-ground points; therefore, the impedance seen by the signal current from the emitters of Q514 or Q524 will be the parallel combination of 153 Ω and 27 Ω or approximately 23 Ω to the AC ground points. Hence, R_E for Q514 or Q524 will be 23 Ω .

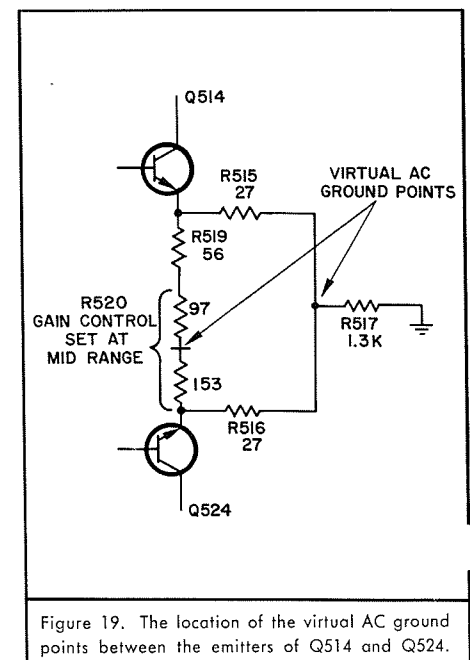


Figure 19. The location of the virtual AC ground points between the emitters of Q514 and Q524.

We have now calculated from this part of the circuit all of the information we need to progress further into the circuit. Let us turn our attention to the circuit around Q513 and Q523. The first thing we notice is that the base of Q513 and Q523 are tied together at an AC-ground point. You will recall that the impedance we see looking into the emitter of the common-base configuration is R_t . In order to calculate R_t we must, of course, calculate r_e and add our constant for R_r of 4Ω ; r_e will be a function of the actual value of current flowing into the emitter. 27 milliamps has been set in the emitter circuit of Q514 and Q524; but not all of this current will flow into the

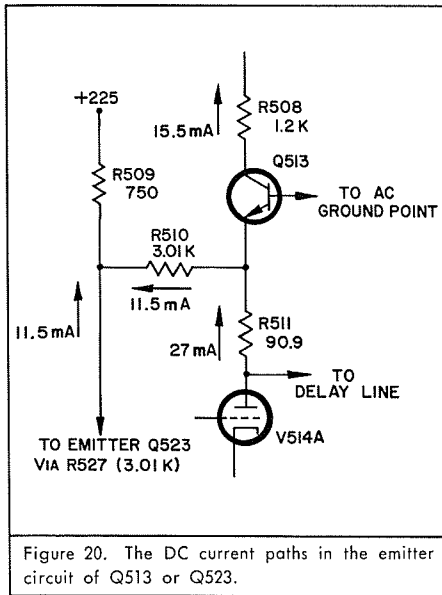


Figure 20. The DC current paths in the emitter circuit of Q513 or Q523.

emitter of Q513 and Q523. 11.5 milliamps will flow through R510 and R527, refer to Figure 20. The actual value of current into Q513 or Q523 will be 15.5 milliamps. Therefore, the impedance (R_t) we see looking into the emitter of Q513 and Q523 will be

$$R_t = r_e + R_r \quad (9)$$

$$= \frac{26}{15.5} + 4\Omega$$

$$= 5.68\Omega$$

This impedance of 5.68Ω plus R511 or R526 (90.9Ω) constitutes part of the load impedance of the hybrid cascode amplifier Q514, V514A or Q524, V514B and the necessary matching impedance for the delay line.

There is one point we should make clear here. We have assumed a value of 4Ω for R_r which you will recall is equal to $\frac{R_b}{\beta}$. R_r can vary from between 2Ω to 24Ω depending upon the type of transistor (refer to Part 1, "The Transistor Amplifier" SERVICE SCOPE #42, February 1967). This is one of those few times we should be really a bit more specific about assuming a value of R_r . The sum of the impedances 5.68Ω and 90.9Ω should be equal to 93Ω since our delay line is a 186Ω balanced line. Therefore, we have a difference of 3.58Ω between the theoretical value and the calculated value, or an error of approximately 3.7%. This error has been due in part to our presupposed value of R_r to be 4Ω . Such an error could not be tolerated in design work but it is acceptable here for our purpose of DC analysis. Bear this limitation in mind when you apply this analysis.

There is another point we must clear up. What is the load impedance of the hybrid cascode amplifier Q514, V514A or Q524, V514B? Clearly it will be that impedance or impedances connected from the plate of V514A or V514B to the AC ground. We are using a balanced delay line of 186Ω , (93Ω to a side), referenced to the AC ground. Therefore, the delay line impedance (93Ω) must shunt R511 in series with R_t (or R526 in series with R_t) making an effective load impedance of approximately 47Ω in the plate circuit of V514A or V514B. We now have all the necessary information to calculate the gain to this point.

$$A_{(v)} = \frac{R_{L(1)} + R_{L(2)}}{R_{E(1)} + R_{E(2)} + R_{t(1)} + R_{t(2)}}$$

$$= \frac{47 + 47}{23 + 23 + 5 + 5}$$

$$= \frac{94}{56}$$

$$A_{(v)} = 1.68$$

Q523 is the trigger pick-off amplifier and Q543 is an emitter follower providing isolation between the vertical amplifier and the trigger circuits.

The trigger pick-off amplifier Q523 is one part of a transistor cascode amplifier. The input stage is Q514 and Q524. Normally, the gain of a transistor cascode amplifier is the ratio of R_L to $R_E + R_t$. The gain in this case must be multiplied by 0.5 for the following reason. The signal current is equally divided at the plate of V514B, half of the signal current will flow through the delay line impedance (93Ω) and the other half through R526 and finally through the load impedance of Q523. The load impedance will be that impedance which is connected to the AC ground. The collector of Q523 is connected to the base of Q543. The impedance we see looking into the base of Q543 is

$$R_{in} = \beta (R_E + R_t) \quad (10)$$

If we choose to neglect the input circuit of the trigger amplifier we see that R_E in this case is R547 $6.5k\Omega$. A beta of 50 is a close figure to use for Q543, and since $R_E \gg R_t$ then,

$$R_{in} = \beta R_E$$

$$= 50 \times 6500\Omega$$

$$= 325k\Omega$$

This impedance shunts R544 ($75k\Omega$) and L528 a $1.5k\Omega$ wire-wound resistor. We may then, for all practical purposes, consider L528 the collector load resistance (R_L); therefore,

$$A_{(v)} = 0.5 \left[\frac{R_L}{R_{E(1)} + R_{E(2)} + R_{t(1)} + R_{t(2)}} \right]$$

$$= 0.5 \left[\frac{1500}{23 + 23 + 5 + 5} \right]$$

$$= 13.3$$

Q534 is the beam-indicator amplifier. Its function is to drive two neon lamps situated above the CRT on the front panel of the oscilloscope. These neons indicate the position of the trace in a vertical direction. In the quiescent condition the voltage at the junction of R535 and R536 is 287 volts. Both indicator neons, B538 and B539, have 287 volts across them, not enough voltage to strike either neon. (This type of neon has a striking voltage in excess of 68 volts.)

When we apply a negative signal to the vertical input of the oscilloscope, the base of Q524 is driven negative and the base of Q514 moves in a positive direction by a similar amount. Therefore, the current through R530 decreases and the current through R507 increases. The voltage at the emitter of Q534 increases and the voltage at the base of Q534 decreases. As a result, the base-to-emitter junction of Q534 becomes reverse biased and Q534 ceases to conduct.

Therefore, the voltage at the junction of R535 and R536 rises towards 350 volts striking neon B539 which indicates trace has shifted down.

R513 and R523 and the DC SHIFT control R502 are thermal-compensation networks associated with Q514 and Q524. The thermal time constants are long and the visible result appears on the CRT display as a DC shift in trace position after

a step function. The DC SHIFT control is adjusted for the best dynamic thermal compensation, typically about 1% tilt.

We will now analyze the output circuits to the right of the delay line, refer to Figure 17. The first thing we must do is to calculate the voltage at the base of Q594 or Q584. The voltage at the junction of R532 and R533 (174 volts) will set the base voltage of Q513 and Q523. Assuming a junction drop of 0.5 volt the voltage at the emitter of Q513 and Q523 will be 173.5 volts. The current through R511 and R526 is 27 milliamps, hence the voltage drop across these resistors will be

$$\frac{90.9 \times 27}{1000}$$

$$\approx 2.5 \text{ volts}$$

therefore, the voltage at the plate of V514A and V514B is

$$173.5 - 2.5 = 171 \text{ volts.}$$

This 171 volts is directly coupled to the base of Q594 and Q584 via the delay line. The voltage at the emitter of both Q594 and Q584 is then 170.5 volts. We will now calculate the current flowing into the emitter of Q594 or Q584. Figure 21 shows a step-by-step approach in solving this problem. The simplest approach is to use Thévenin's Theorem to simplify the resistive network R569, R570, R571 and R572. The result is we have a V_{oc} of +100 volts and a Z_{th} of 900 Ω to the junction of R574 and R576. Therefore, looking from the emitter of either Q594 or Q584 we see an impedance of 13.3 Ω in series with 1800 Ω to +100 volts.

$$I_E = \frac{(170.5 - 100) 10^3}{1.8 \times 10^3} \text{ mA}$$

$$= \frac{70.5}{1.8}$$

$$= 39 \text{ mA}$$

we now calculate r_e

$$r_e = \frac{26}{I_E} = \frac{26}{39}$$

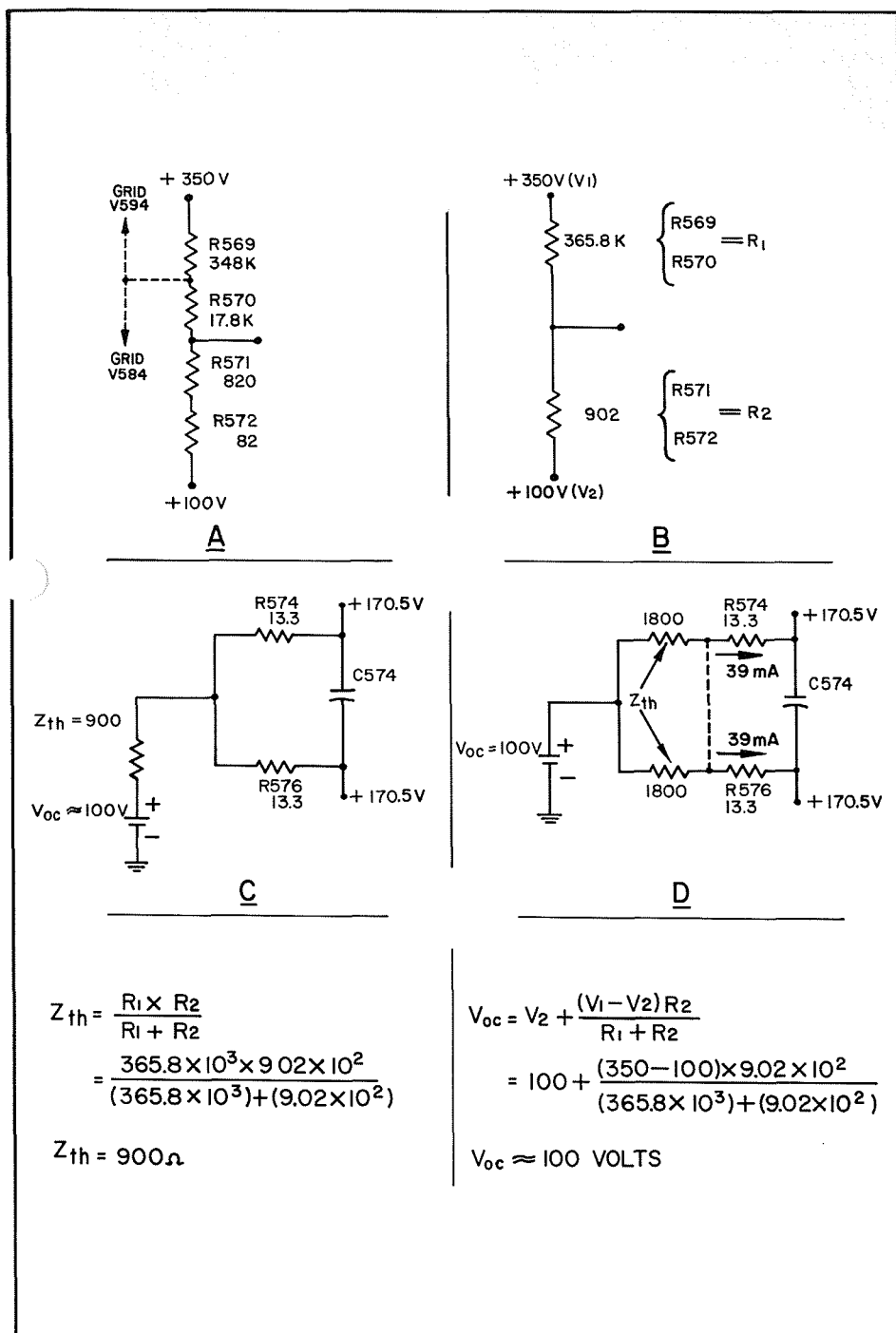
$$\approx 0.7 \Omega$$

and to this we add our constant R_r of 4 Ω ; therefore,

$$R_t = r_e + R_r \quad (9)$$

$$= 0.7 + 4.0$$

$$= 4.7 \Omega's$$



$$Z_{th} = \frac{R_1 \times R_2}{R_1 + R_2} = \frac{365.8 \times 10^3 \times 9.02 \times 10^2}{(365.8 \times 10^3) + (9.02 \times 10^2)}$$

$$Z_{th} = 900 \Omega$$

$$V_{oc} = V_2 + \frac{(V_1 - V_2) R_2}{R_1 + R_2} = 100 + \frac{(350 - 100) \times 9.02 \times 10^2}{(365.8 \times 10^3) + (9.02 \times 10^2)}$$

$$V_{oc} \approx 100 \text{ VOLTS}$$

Figure 21. Illustrating the use of Thévenin's Theorem to simplify a network consisting of a voltage source and a resistive network. (A)—The network whose Thévenin equivalent is to be determined. (B)—Determining the equivalent source impedance (Z_{th}) and the equivalent voltage source (V_{oc}). (C)—The Thévenin equivalent network of (A) connected to the junction of R574 and R576. (D)—The equivalent circuit considering Z_{th} as two resistors through which the individual emitter currents will flow.

We have only one point in this circuit (a virtual AC ground point) at which the signal currents will be equal and opposite. That point is the junction of R574 and R576 (13.3 Ω resistors). This fact sets R_E at 13.3 Ω. The purpose of the RC network to the right of R574, R576 is to compensate the high frequencies.

The input impedance we see looking into the base of Q594 or Q584 is

$$R_{in} = \beta (R_E + R_t) \quad (10)$$

A beta of 75 for this type of transistor is a close figure to use for practical purposes. Therefore,

$$\begin{aligned} R_{in} &= 75 (13.3 + 4.7) \Omega \\ &= 1350 \Omega \end{aligned}$$

The value of R_{in} is part of a resistive network which will terminate the delay line in its correct impedance. Therefore, before we leave this section we must check to see if our

value of R_{in} is within practical limits. Figure 22 shows a progressive breakdown of this network.

This network will induce a loss between the two stages. The signal is reduced in amplitude by a factor of 0.64 because of the voltage divider network consisting of 57.6 Ω and the parallel combination of 100 Ω, 2800 Ω, and the input impedance into Q594 or Q584.

The gain of the output stage is

$$\begin{aligned} A_{(v)} &= \left[\frac{R_{L(1)} + R_{L(2)}}{R_{E(1)} + R_{E(2)} + R_{t(1)} + R_{t(2)}} \right] \eta \\ &= \left[\frac{1100 + 1100}{13.3 + 13.3 + 4.7 + 4.7} \right] \eta \\ &= \left[\frac{2200}{36} \right] \eta \\ &= 61 \eta \end{aligned}$$

You recall that the gain equation of a hybrid cascode amplifier (refer part 2, "The Vacuum Tube Amplifier," Service Section #43, April 1967) must be multiplied by a plate efficiency factor (η) if the vacuum tube is not a triode. The plate efficiency factor (η) normally varies from between 0.7 to 0.9. In this case (η) is approximately 0.9 - 0.88 to be exact. So finally,

$$\begin{aligned} A_{(v)} &= 61 \times \frac{9}{10} \\ &= 54.9 \end{aligned}$$

The gain of the complete Type 545B vertical amplifier is

$$\begin{aligned} A_{(v)} \text{ (total)} &= 54.9 \times 1.68 \times 0.64 \\ &= 59 \end{aligned}$$

Summary

This brings to a close this series of three articles dealing with a practical approach to transistor and vacuum-tube amplifiers. This approach has been offered as a direct method of trouble shooting and understanding circuits. There are limitations as to application as we have seen. However, these limitations do not impair the practical approach we must apply to our everyday maintenance and trouble shooting problems.

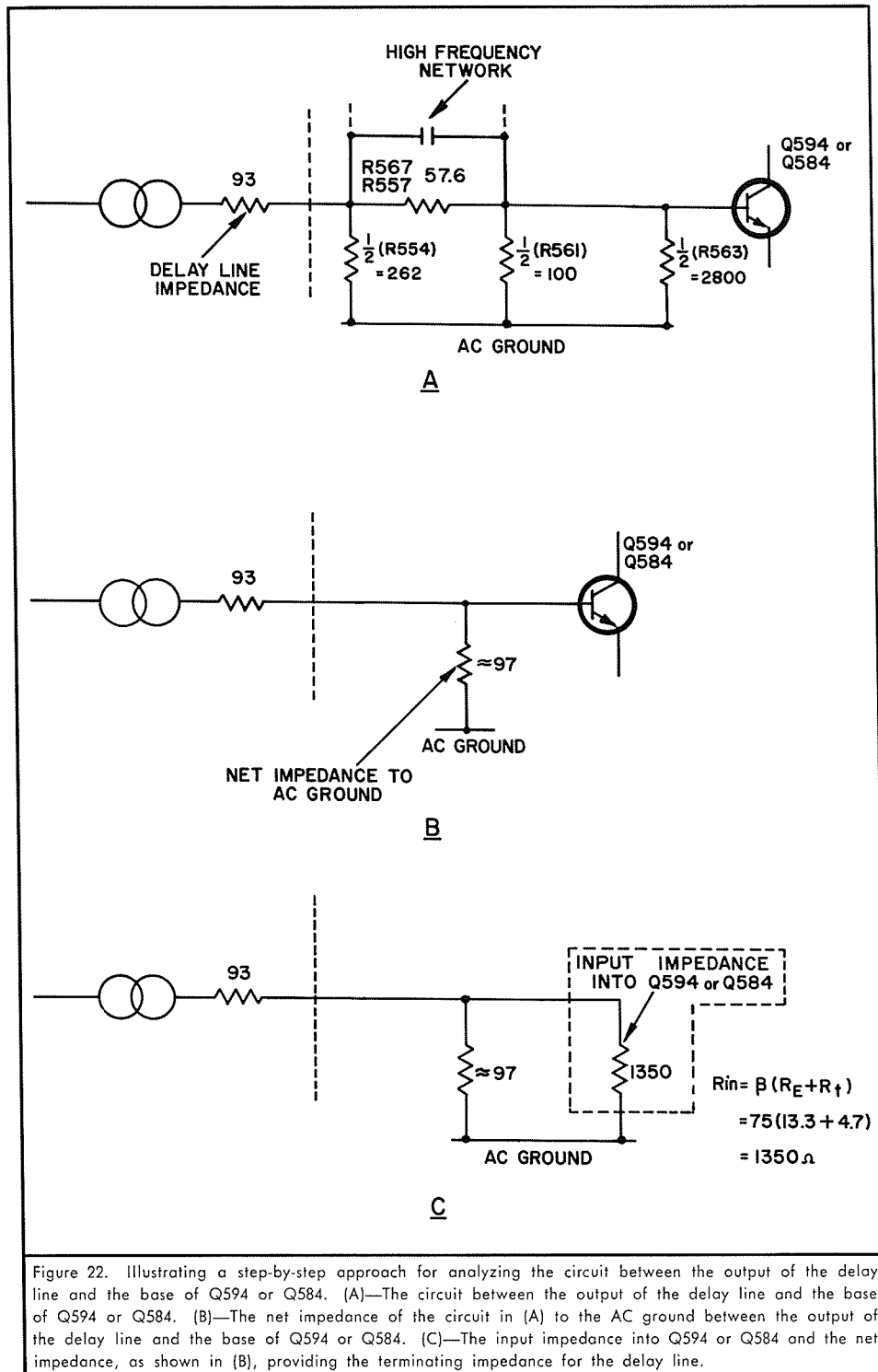
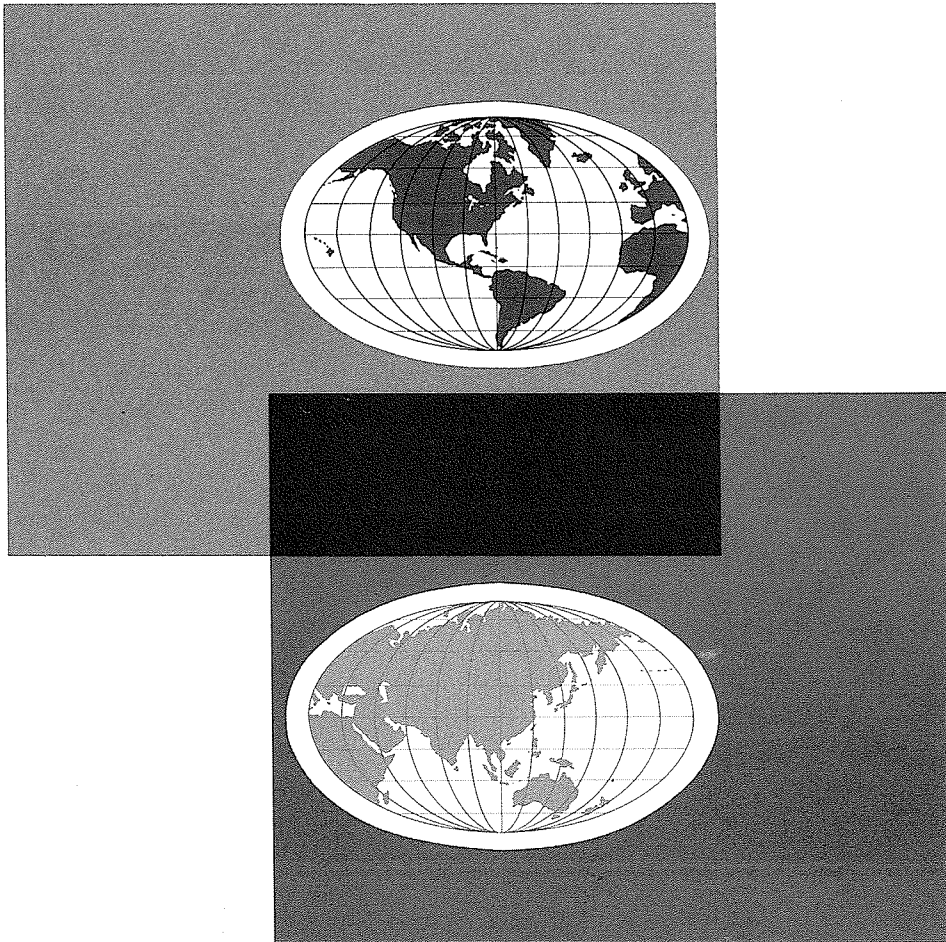


Figure 22. Illustrating a step-by-step approach for analyzing the circuit between the output of the delay line and the base of Q594 or Q584. (A)—The circuit between the output of the delay line and the base of Q594 or Q584. (B)—The net impedance of the circuit in (A) to the AC ground between the output of the delay line and the base of Q594 or Q584. (C)—The input impedance into Q594 or Q584 and the net impedance, as shown in (B), providing the terminating impedance for the delay line.

LIST OF SYMBOLS

<p>ΔE_{out} Voltage gain defined as $\frac{\Delta E_{out}}{\Delta E_{in}}$</p> <p>$E_b$ Average or quiescent value of plate voltage</p> <p>E_{bb} Plate supply D-C voltage</p> <p>E_c Average or quiescent value of grid voltage</p> <p>E_g Effective or maximum value of varying component of grid voltage</p> <p>E_k Average or quiescent value of cathode voltage</p> <p>E_p Effective or maximum value of varying component of plate voltage</p> <p>gm Mutual conductance</p> <p>I_b Base current</p> <p>I_c Collector current</p> <p>I_E Emitter current</p> <p>I_k Cathode current</p> <p>I_p Plate current</p>	<p>R_b Base spreading resistance (Tee Equivalent)</p> <p>R_c Collector resistance (Tee Equivalent)</p> <p>R_e Emitter resistance (Tee Equivalent)</p> <p>R_E External Emitter resistance (refer to text)</p> <p>$R_{E(s)}$ The equivalent resistance between the signal source and the emitter terminal of the transistor in the common base configuration</p> <p>r_e Dynamic emitter resistance</p> <p>R_k Cathode resistance (refer to text)</p> <p>r_k The impedance seen looking into the cathode of a vacuum tube and defined as</p> $\frac{r_p + R_L}{\mu + 1} \approx \frac{1}{gm} \text{ (if } r_p \gg R_L)$ <p>R_L Load resistance</p> <p>rp Dynamic plate resistance</p> <p>R_r $\frac{R_b}{\beta}$</p>	<p>R_t The "Transresistance" resistance ($r_e + R_r$)</p> <p>V_{bb} Base voltage</p> <p>V_{cc} Supply voltage</p> <p>V_{ce} Collector to emitter voltage</p> <p>α (alpha) The total forward current gain of the transistor as viewed from an external circuit. Normally defined as the ratio of I_c/I_E</p> <p>β (beta) The small signal current gain</p> <p>Δ (delta) The change in the variable with which it is associated</p> <p>η (eta) Plate efficiency factor (refer to text)</p> <p>μ (mu) Amplification factor</p>
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