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The vacuum tube triode at ultra high frequencies

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By

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Academic Dean

PREFACE

From January 1, 1949, to March 18, 1949, I was stationed at the R.C.A. tube plant in Harrison, New Jersey, to work with the Advance Development Section. This period constituted the winter term of the third year in the course in engineering electronics at the U.S. Naval Post Graduate School, Annapolis, Maryland. Because I was interested in working with vacuum tubes in the microwave region, I was assigned by G.M. Rose, in charge of the Advance Development Section, to work with W.A. Harris on a project to measure the input admittance of the pencil tube triode at ultra high frequencies. I became interested in the problems encountered in the operation of the space charge control vacuum tube triode at ultra high frequencies during this time and started research for this thesis in the excellent library at the plant. It is an important subject in the light of present-day trends in television, UHF communication, and radar.

I wish to acknowledge the contribution of G.M. Rose and D.W. Power, who developed the pencil tube, and of the other members of the Advance Development Section, all of whom were always eager to advise and assist me in my work. I am especially grateful to W.A. Harris for his friendly counsel and guidance during my duty with the section.

D.C. Peto

Annapolis, Maryland
May, 1949

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TABLE OF SYMBOLS

a - distance between cathode and grid.

b - distance between grid and anode.

b_N - susceptive component of admittance.

c - velocity of light.

C - capacitance.

d - distance between electrodes.

e - electron charge.

E - potential.

ϵ - dialectric constant.

f - frequency.

F - electric field.

g - conductive component of admittance.

I - current.

K - constant.

l - length.

L - inductance.

m - mass of an electron.

η - efficiency.

q - charge.

r - radius.

t - time.

T - period of a cycle.

τ - transit time.

μ - mu of a tube.

v - velocity.

V - voltage.

W - power.

x - variable distance.

Y - admittance.

Z - impedance.

λ - wave length.

θ - phase angle.

CHAPTER I

INTRODUCTION

1. Historical.

About three hundred years ago, Otto von Guericke, with primitive vacuum tube and electrostatic friction machine, observed the colorful phenomenon of gas discharge in an evacuated glass vessel. Other scientists made similar experiments but it was not until 1869 that Wilhelm Hittorf published a paper "On the Conduction of Electricity by Gases", the first actual scientific study of this remarkable effect. Ten years later the work of William Crookes inspired further serious study of gaseous discharge, as a result of which the entire philosophical background of science was changed by the electron theory of matter. From this evolved the modern electronic tube.

Mouromtseff, (29), traces the development of the main types of modern electronic tube to three independent sources and several independent lines of development. Thus, electron-beam tubes such as the Lenard tube (1894), the Roentgen x-ray tube (1895) and the cathode ray tube (Braun, 1897) are derived from Crookes' studies as are the mercury arc rectifier (Cooper Hewitt, 1902) and its modern derivatives, the thyratron, phanatron, ignitron and excitron. The phototubes originated from the phenomenon observed by Hertz and scientifically studied by Hollwachs (1888), Elster and Geitel (1912) and others. Finally there are the high vacuum tubes which are connected to the Edison effect observed in incandescent lamps (1884) through DeForests audion (1908) and Flemings valve (1904). In this class are the ultra high frequency triodes considered in this paper. For this reason its historical development will be considered more closely.

In 1873 Guthrie noticed that a negatively charged electroscope was discharged when a hot metal ball was brought near it but that a much hotter metal ball was required to discharge a positively charged electroscope. We know now that in the former case positive ions of occluded gas were given off while in the latter case electrons were emitted. Between the years 1882 and 1889 Elster and Geitel, two German scientists, used a two electrode vacuum tube to investigate conduction effects of gases under reduced pressure. They noted the unilateral conductivity but did not develop its rectification possibilities. In 1882, two French scientists, Jamin and Meneuvrier also observed unidirectional flow of electric current, in this case from an anode to a mercury cathode. Apparently unaware of these experiments, Edison, in 1883, while investigating the effect of hot-spots (caused by diminished cross sectional area at certain points) in the carbon filaments of his new incandescent lamps, made an important discovery. The hot-spots blackened the bulbs due to carbon particle emission. To study this he sealed a plate between the legs of a horse shoe - shaped filament. With a galvanometer he detected current flow between plate and the positive end of the filament but not between the plate and the negative end of the filament. This phenomenon is called the Edison effect and some people consider it the real starting point of the modern thermionic tube.

In 1884 and 1885, Preece, in studying the Edison effect, showed that the current was a function of filament temperature, interelectrode distance, and interelectrode potential difference, but not of the type of metal used in the plate. In 1884 Hittorf showed that small potential difference was required if the cathode was incandescent. In 1885 Goldstein reported that with low pressure and cold cathode, the highest voltages caused no perceptible current flow whereas with hot cathode, only a low potential was required. In 1892, Zehnder demonstrated the first vacuum tube used in high frequency circuits in an apparatus built to exhibit to large audiences the then recently discovered Hertzian waves. It was a small gas-filled tube in which gaseous discharge was triggered by high frequency oscillations.

The first tube developed for practical radio was the famous Fleming valve of 1904. It was a diode detector based on the phenomenon of unidirectional current flow. Oxide-coated cathodes were invented the same year by Wehnelt. The triode was invented in 1908 by de Forest and was known as the audion. The effect of the grid on the electrostatic field was known to scientists from the fundamental works of Maxwell and Rieman.

Langmuir in this country in 1913 demonstrated the importance of high vacuum as did Schottky in Europe early in 1914. Langmuir and Gaede developed mercury pumps about this time capable of establishing the highest necessary vacua. As a direct result, it became possible to design and build triodes capable of operation at high voltages and capable of generation of high frequency output up to several hundred watts. This progress made radio broadcasting feasible in 1920. Another technical advance in the art of making tubes was made in 1922 by Houskeeper, who developed the important glass-to-copper seals. This made water-cooled high power tubes possible.

The foregoing represents in a general way the state of development of the vacuum tube triode in the early 1920s. The next significant advance occurred in 1932 when B. J. Thompson and G. M. Rose developed a tube of small dimensions for use at wavelengths around 50 centimeters. The previous limit was around 5 meters. This inspired new attempts to further extend the upper frequency limit of the triode. The advent of television, frequency modulation, and radar since then has created a tremendous demand for increasing the maximum frequency at which the triode can successfully amplify and oscillate. During World War II, disk seal tubes were developed for use in circuits in which the tube is an integral component. Modifications and improvements on this type of tube have brought about the modern microwave triode.

2. Summary of this thesis.

This paper considers in some detail the problems encountered in the operation of space-charge control triodes (as distinguished from electron velocity types, such as the klystron and the magnetron) at ultra high frequencies corresponding to wavelengths in the microwave region (30 cm and less).

Low frequency amplifier and oscillator theories do not apply at these high frequencies. Acceptable theories are derived as an aid to understanding the high frequency limitations. The operation of the triode is explained on the basis of various admittances involved. The failure of the triode at ultra high frequency is due to changes in these admittances at high frequency caused by three factors; transit time, lead inductance, and interelectrode capacitance. Each of these factors is considered in detail. Means of increasing the frequency range of the triode by minimizing the effects of these factors are demonstrated. The combination of tubes with small interelectrode spacings and concentric line circuitry in which the tube is an integral part of the circuit accounts for the progress in recent years in extending the range of the triodes into the microwave region. Tubes in use today belong to the disk seal family in which the electrode leads or contacts are disks or cylinders of metal brought through the tube envelope by glass seals. The familiar lighthouse tubes and the more recent pencil tube are examples of this type of triode and are described briefly at the end of this paper.

CHAPTER II

AMPLIFIER AND OSCILLATOR THEORY

1. Amplifier theory.

In order to specify completely the behaviour of a tube in a single stage linear amplifier circuit at any frequency, it is necessary to specify the short-circuit input, output, forward, and feedback admittances at this frequency. Using these admittances, a four-terminal equivalent circuit for the tube can be constructed and the performance of the tube may then be calculated on a nodal basis. These admittances may be defined as follows:

(a) The short-circuit input admittance, measured at the input terminals, is the quotient of the current at the input terminals divided by the voltage at the input terminals, with the output terminals shorted for the signal frequency.

(b) The short-circuit output admittance, measured at the output terminals, is the quotient of the current at the output terminals divided by the voltage at the output terminals, with the input terminals shorted for the signal frequency.

(c) The short-circuit forward admittance is the quotient of the current at the output terminals divided by the voltage between the input terminals, with the output terminals shorted for the signal frequency.

(d) The short-circuit feedback admittance is the quotient of the current at the input terminals divided by the voltage between the output terminals, with the input terminals shorted for the signal frequency.

In the following section on amplifier theory, the manner in which these various admittances affect the gain and stability of the amplifier is discussed in detail.

At zero or very low frequencies the input admittance is essentially the grid conductance, the output admittance is the plate conductance and the feedback and forward admittances are likewise conductances. The equivalent circuit for these low frequencies is shown in Fig. 1(a).

For frequencies of the order of a kilocycle to a megacycle per second, the effects of the input, output, and feedback capacitances are handled by adding them appropriately as shown in Fig. 1(b). At these lower frequencies the values of the conductances and capacitances can be measured, and each element in the equivalent circuit is then accurately known.

At frequencies higher than about ten megacycles per second, marked increases in the input, output, forward, and feedback admittances occur other than those caused by interelectrode capacitance. The causes for these modifications can generally be traced to either or both of the following effects:

(a) the effect of the passive coupling circuit connecting the electron stream to the externally available terminals, and

(b) the effects of finite transit time for electron travel through the interelectrode spaces.

The effects of passive coupling are indicated in Fig. 1(c) by the insertion of lead inductances and by the splitting of the low frequency capacitances into electrode and circuit components. It is thus apparent that at these high frequencies the equivalent circuit

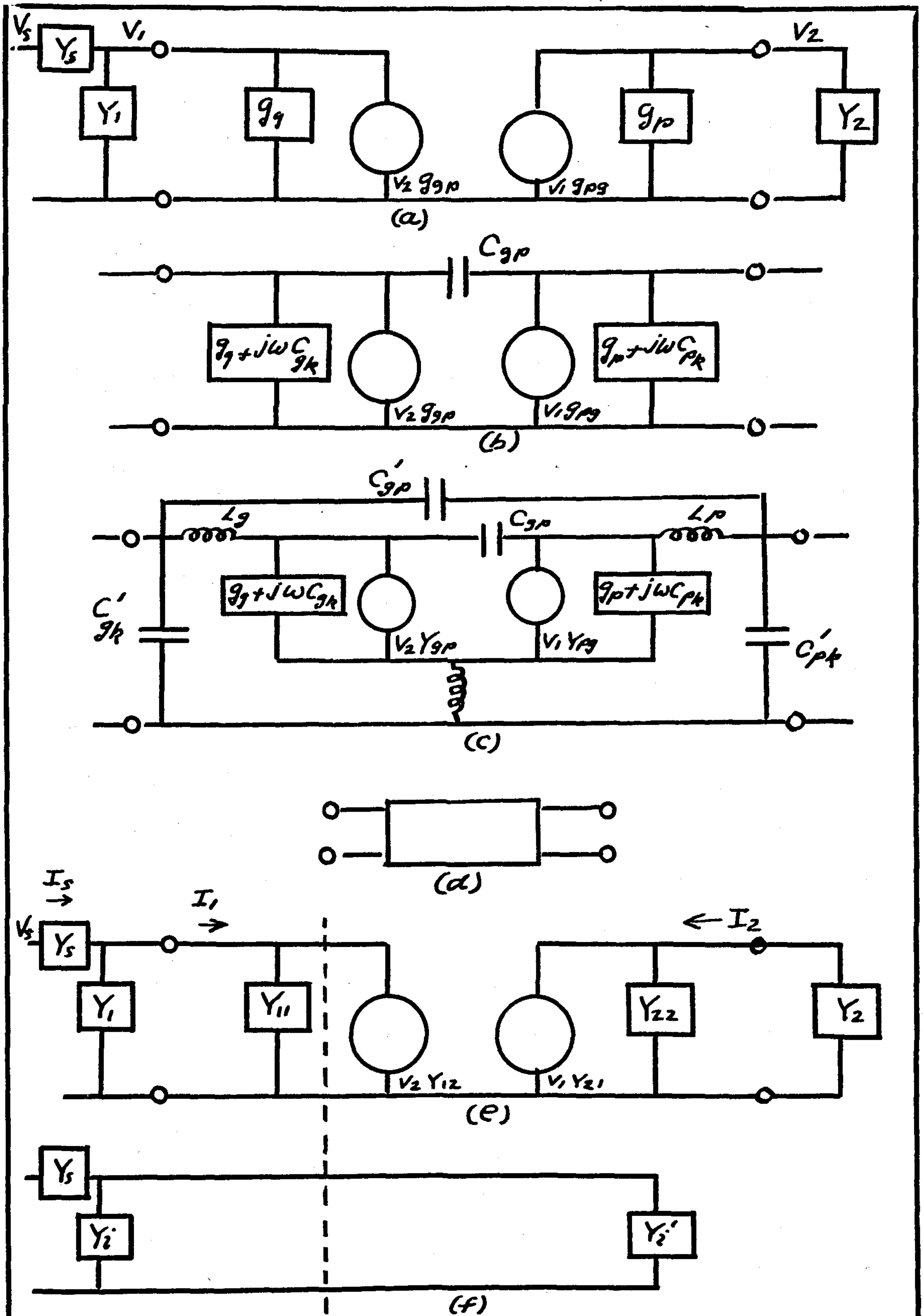
Vs
Ys
V1
V2
Y1
gg
gp
Y2
V2 ggp
V1 gpg
(a)
Cgp
gg+jωCgk
gp+jωCpk
V2 ggp
V1 gpg
(b)
C'gp
Lg
Cgp
Lp
C'gk
gg+jωCgk
gp+jωCpk
V2 Ygp
V1 Ypg
C'pk
(c)
(d)
Is
I1
I2
Vs
Ys
Y1
Y11
Y22
Y2
V2 Y12
V1 Y21
(e)
Ys
Yi
Yi'
(f)

figure 1

becomes quite complex, and the computations are correspondingly complicated. What is even worse, the measurement of the various component values within the tube becomes too difficult, if not impossible. Above one hundred megacycles per second the effects of transit time are felt to a greater extent and the problem gets even more involved. Clearly, what is desirable is a concept of the tube with its internal circuits and external sockets, leads, filters, etcetera, considered as a unit with all measurements taken from four accessible terminals.

Consider the four terminal box of Fig. 1(d). As a linear transducer, its performance can be specified by the four short-circuit admittances previously defined. Connected across the input terminals of this box is a source of voltage, V_s , through a general admittance, Y_s , which includes the internal admittance of the voltage generator. An input admittance, Y_1 , is also connected across the input terminals, in order to allow tuning in the input circuit. Similarly, there is a tunable load admittance, Y_2 , across the output terminals. Y_{11} and Y_{22} are the short-circuit input and output admittances, respectively, and Y_{12} and Y_{21} are the short-circuit feedback and forward admittances, respectively. V_1 is the voltage at the input terminals and V_2 is the voltage at the output terminals. The nodal equations for this arrangement are the familiar

$$I_1 = Y_{11} V_1 + Y_{12} V_2 \qquad (1)$$

$$I_2 = Y_{21} V_1 + Y_{22} V_2 \qquad (2)$$

The terminal conditions are

$$I_s = (V_s - V_1) Y_s = V_1 Y_1 + I_1 \tag{3}$$

$$I_2 = - Y_2 V_2 \; . \tag{4}$$

From equations (2) and (4),

$$\frac{V_2}{V_1} = - \frac{Y_{21}}{Y_{22} + Y_2} \; . \tag{5}$$

The total input admittance under operating conditions is

$$Y_T = \frac{I_1}{V_1}$$ which, from (5) and (1), becomes

$$Y_T = Y_{11} - \frac{Y_{12} Y_{21}}{Y_{22} + Y_2} \; . \tag{6}$$

From equations (1), (3), and (6),

$$\frac{V_1}{V_S} = \frac{Y_S}{Y_S + Y_1 + Y_T} \; . \tag{7}$$

The overall gain of the circuit is the product of input gain times output gain. The input gain, A_i, is given by equation (7). The output gain, A_o, is given by equation (5). Therefore the overall gain is

$$A = A_i \cdot A_o = \frac{Y_S}{Y_S + Y_1 + Y_T} \cdot \frac{-Y_{21}}{Y_{22} + Y_2} \; . \tag{8}$$

From the above equations certain conclusions may be drawn relating the various short-circuit admittances to the total operating input admittance. Thus the voltage gain from the input to the output terminals is the ratio of the short-circuit forward admittance to the sum of the short-circuit output admittance and the admittance of the load connected between the output terminals. The overall gain is a function of this intrinsic gain and also of the input admittances. Also, the added current at the input terminals due to the presence of the load is the product of the input voltage, the voltage gain, and the short-circuit feedback admittance. Finally, the total input admittance is the sum of the short-circuit input admittance and the product of the voltage gain and the short-circuit feedback admittance. The phase angle of the added component is the sum of the phase angle for the voltage gain and the phase angle for the feedback admittance.

The four terminal box of Fig. 1(d) can now be replaced by an arrangement of the above admittances as shown in Fig. 1(e). This is a completely general equivalent circuit for a single stage amplifier including the tube and the circuits attached to its elements. All voltages and admittances are considered to be complex. This network is applicable for all frequencies. As previously stated, the terminals are accessible and all parameters are assumed to be known and measurable functions of frequency. For further simplification, the network of Fig. 1(f) will be used in addition to its equivalent network, that of Fig. 1(e). In this simplified network there are defined in addition to the parameters of Fig. 1(e), the admittance, Y_1', or input admittance, which is the parallel combination of the short-circuit input admittance, Y_{11}, and the tunable admittance, Y_1, on the generator side of the input terminals. Y_2' is defined as the additional input admittance due to feedback from the output circuit to the input circuit. Finally, Y_0 is defined as the output admittance and includes the short-circuit output admittance, Y_{22}, and the tunable load admittance, Y_2. Thus, when reference is made to the tuning of Y_i or Y_o, it will be understood that the parameters actually varied are Y_1 and Y_2, respectively.

From equation (6) it is seen that the total input admittance consists of the short-circuit input admittance in parallel with another admittance which may be termed the additional input admittance. Denoted this by Y_i', then

$$Y_i' = -\frac{Y_{12}Y_{21}}{Y_{22}+Y_2} = -\frac{Y_{12}Y_{21}}{Y_o}. \tag{9}$$

This additional input admittance represents the extent of the reaction of the output circuit upon the input circuit. In other words, it is a measure of the loading and detuning of the input circuit caused by the output circuit. This quantity may also be used to determine the conditions for stability of the amplifier.

Equation (9) makes several important properties of additional input admittance apparent. First, Y_i' is zero when Y_o is infinite or when either Y_{12} or Y_{21} is zero, providing that Y_o is not zero. In general, Y_{21} and Y_o are neither zero or infinite. It can then be stated that the output circuit has neither loading nor detuning effects when Y_{12}, the short-circuit feedback admittance, is equal to zero.

Second, the conductance component of the additional input admittance is zero when the phase angle, or argument, of Y_i' is plus or minus 90 degrees. Thus the output circuit reflects no conductance across the input circuit when

$$\arg Y_i' = \pm\frac{\pi}{2}.$$

$$\arg\left(\frac{Y_{12}Y_{21}}{Y_0}\right) = \pm\frac{\pi}{2},$$

$$\arg Y_{12} + \arg Y_{21} = \pm\frac{\pi}{2} + \arg Y_0 . \qquad (10)$$

Third, the susceptance component of Y_i' is zero when arg Y_i' is zero or 180 degrees, in which case the output circuit will reflect no detuning component. Under these conditions,

$$\arg Y_{12} + \arg Y_{21} = (0 \text{ or } \pi) + \arg Y_0 . \qquad (11)$$

Fourth, the conductive component of Y_i' is a maximum or minimum when *

$$\mathcal{R}_e\left[\frac{dY_i'}{dY_0}\, dY_0\right] = 0. \qquad (12)$$

In general, the load is tunable only in its susceptance component so that the real part of dY_0 is zero and equation (12) becomes

$$\arg \frac{dY_i'}{dY_0} = 0 \text{ or } \pi. \qquad (12a)$$

Performing the indicated differentiation on equation (9) results in

$$\arg Y_{12} + \arg Y_{21} = (0 \text{ or } \pi) + 2\arg Y_0 . \qquad (13)$$

Fifth, the conduction which gives the maximum and minimum values

* Walter Van B. Roberts, (32), describes the method for making this maximization.

of the susceptance component of Y_i' is

$$\mathcal{I}_m \left[\frac{d Y_i'}{d Y_0} d Y_0 \right] = 0 . \tag{14}$$

Again assuming Y_0 to be tunable in its imaginary part only, equation (14) becomes,

$$\arg \frac{d Y_i'}{d Y_0} = \pm \frac{\pi}{2} . \tag{14a}$$

Proceeding as before,

$$\arg Y_{12} + \arg Y_{21} = \pm \frac{\pi}{2} + 2 \arg Y_0 . \tag{15}$$

From the foregoing, the conditions for stability of the amplifier may be obtained. When the amplifier is on the verge of oscillations, the total effective conductance and the total effective susceptance of the circuit as a whole are each equal to zero. Thus, for stability,

$$g_s + g_i + g_i' > 0 \tag{16}$$

$$b_s + b_i + b_i' > 0 . \tag{17}$$

Before oscillations can occur in the amplifier, both of the above conditions must fail. Therefore, if the condition expressed in inequality (16) is always fulfilled, the amplifier will be stable. If g_i has its maximum negative value as determined by equation (13) and condition (16) still holds, then the amplifier will be stable regardless of the tuning of Y_0 .

The overall voltage amplification of the stage is given by equation (8). As previously stated, the overall gain is the product of A_i, the gain of the input circuit, and A_o, the gain of the output circuit. These factors are not independent, however, since a variation in the load admittance affects not only the output voltage but also the additional input admittance, thereby causing a change in the input voltage from grid to cathode. For this reason, tuning the output admittance, Y_o for maximum absolute value of output gain, $|A_o|$, does not, in general, provide the optimum overall gain tuning.

From equation (5) it is apparent that $|A_o|$ can not be infinite as long as the real parts of Y_{22} and Y_2 are positive. The condition for the maximum value of $|A_o|$ is

$$\mathcal{R}_e\left[\frac{1}{A_o}\frac{dA_o}{dY_o}\,dY_o\right] = 0. \qquad (18)$$

If Y_o is tunable only in its susceptence component, this condition may be stated as

$$\arg\left(\frac{1}{A_o}\frac{dA_o}{dY_o}\right) = 0 \text{ or } \pi. \qquad (18a)$$

This requires that

$$\arg Y_o = 0 \text{ or } \pi. \qquad (19)$$

The π solution of (19) is not possible if g_o is positive.

From equation (8) it is apparent that A_i will be infinite (and so, therefore, will A providing A_o is not zero) when the input and output circuits are so adjusted that

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$$Y_s + Y_1 + Y_T = Y_s + Y_i + Y_i' = 0. \qquad (20)$$

This condition corresponds to the unstable state discussed in connection with inequalities (16) and (17).

It is important to know the value of the maximum overall gain and the adjustments of input and output circuits which gives this maximum, for given values of Y_s, Y_{21}, and Y_{12} in the range from the point of no feedback to the point of maximum stability. These adjustments can be calculated by maximizing $|A|$ with respect to Y_i and Y_o separately, and then combining the resultant conditions. The maximization relations are

$$R_e\left[\frac{1}{A}\frac{dA}{dY_i}\,dY_i\right] = 0. \tag{21}$$

$$R_e\left[\frac{1}{A}\frac{dA}{dY_o}\,dY_o\right] = 0. \tag{22}$$

Assuming, as before, that Y_i and Y_o are tunable in the imaginary parts only, equations (21) and (22) become, respectively,

$$\arg\left(\frac{1}{A}\frac{dA}{dY_i}\right) = 0 \text{ or } \pi \tag{21a}$$

$$\arg\left(\frac{1}{A}\frac{dA}{dY_o}\right) = 0 \text{ or } \pi. \tag{22a}$$

Performing the indicated differentiations equation (21a) becomes

$$\arg(Y_s + Y_i + Y_i') = 0 \text{ or } \pi. \tag{21b}$$

and equation (22a) becomes

$$\arg\left(Y_o - \frac{Y_{12}Y_{21}}{Y_s + Y_i}\right) = 0 \text{ or } \pi. \tag{22b}$$

From equations (21b) and (22b) it can be shown that

$$\arg Y_o = \arg(Y_s + Y_i) . \tag{23}$$

Thus, for maximum absolute value of overall gain, the tuning adjustments for the output circuit and the combined input circuit must be similar. The actual value of the adjustment, in terms of the circuit parameters, may be determined from equations (22b) and (23). This results in

$$b_o = \frac{b(g_{12}g_{21} - b_{12}b_{21}) - g(b_{12}g_{21} + b_{21}g_{12})}{g^2 + b^2} \tag{24}$$

$$g_o = g\frac{b_o}{b} . \tag{25}$$

where g and b are the conductive and susceptive components, respectively, of the combined input admittance $Y_s + Y_i$.

Equations (24) and (25) may be combined to give

$$\tau_o^3 + \frac{(1 + b_{12}b_{21} - g_{12}g_{21})\tau_o}{g_o g} + \frac{b_{12}g_{21} + b_{21}g_{12}}{g_o g} = 0 \tag{26}$$

$$\tau_o = \frac{b_o}{g_o} .$$

The solution of equation (26) provides the value of the angle to which the output circuit and combined input circuit must be tuned in order that maximum $|A|$ be obtained.

Equations (1) through (26) are a modification of the equations derived by B. Salzberg, (33), whose work is based on the more conventional equivalent circuit in which the connection between the plate and grid circuits is by means of a coupling admittance, Y_c , between the plate and the grid and the current generator, $Y_M V_1$, between the plate and ground where Y_M is the transadmittance and V_1 is the voltage at the grid. This circuit is useful in relating the maximum permissible coupling capacitance between the grid and the plate as a function of the phase angle of the transadmittance (for small phase retardations the maximum permissible coupling before oscillations occur is increased, but the improvement in stability can be shown to be at the expense of overall amplification). For the reasons outlined earlier, the more general four terminal network was considered to be more satisfactory.

The operating characteristics of the amplifier, insofar as gain and stability are concerned, are related to the various admittances in the manner developed in this section. A similar discussion of the triode as an oscillator will demonstrate other important relationships.

2. Oscillator theory.

In order to maintain oscillations in a tuned circuit, energy must be applied to the circuit to overcome the resistive losses. The system then behaves as though it had a zero or negative resistance. Therefore the condition for oscillation of a triode can be obtained by calculating the effective admittance between any pair of electrodes and inserting the condition that the real part of this admittance must be zero or negative.

Consider the circuit shown in Fig. 2(a). This circuit indicates the admittances Y_1 , Y_2 , and Y_3 between the electrodes of a triode. These admittances include the components inside the tube as well as any external components. The circuit will oscillate if the effective admittance between any two electrodes has a real component equal to or less than zero. For example, let it be required the real part of the effective admittance, Y_2', between the plate and the grid must be less than zero. It can be shown that

$$Y_2' = Y_2 + \frac{Y_1 Y_3}{Y_1 + Y_3 + Y_M} \tag{27}$$

when Y_2 is the admittance between the plate and grid

Y_1 is the admittance between the cathode and grid

Y_3 is the admittance between the plate and cathode

Y_M is the grid-to-plate transadmittance.

In general, all admittances are complex.

Thus,

$$Y = g + jb .$$

At low frequencies, where the transit time effects are negligible, the transadmittance is a real number, or

$$Y_M = g_M .$$

Then, equating the real parts of equation (27),

$$g_2' = g_2 + \frac{g_1 g_3 (g_1 + g_3 + g_M) + g_3 b_1^2 + g_1 b_3^2 - g_M b_1 b_3}{(g_1 + g_3 + g_M)^2 + (b_1 + b_3)^2} \tag{28}$$

It is required that g_2' be negative. Since g_1, g_2, g_3, and g_M are positive, it is therefore apparent from equation (28) that the product of b_1 times b_3 must be positive. In other words, b_1 and b_3 must be alike in sign, either both capacitive or both inductive. From a similar solution for g_3', it emerges that b_2 must be of opposite sign to b_1 . From these considerations it is apparent that there are two ways only of adjusting the circuits of a triode for oscillation. These two classes may be presented as follows in terms of the susceptive components of the admittances between the electrodes:

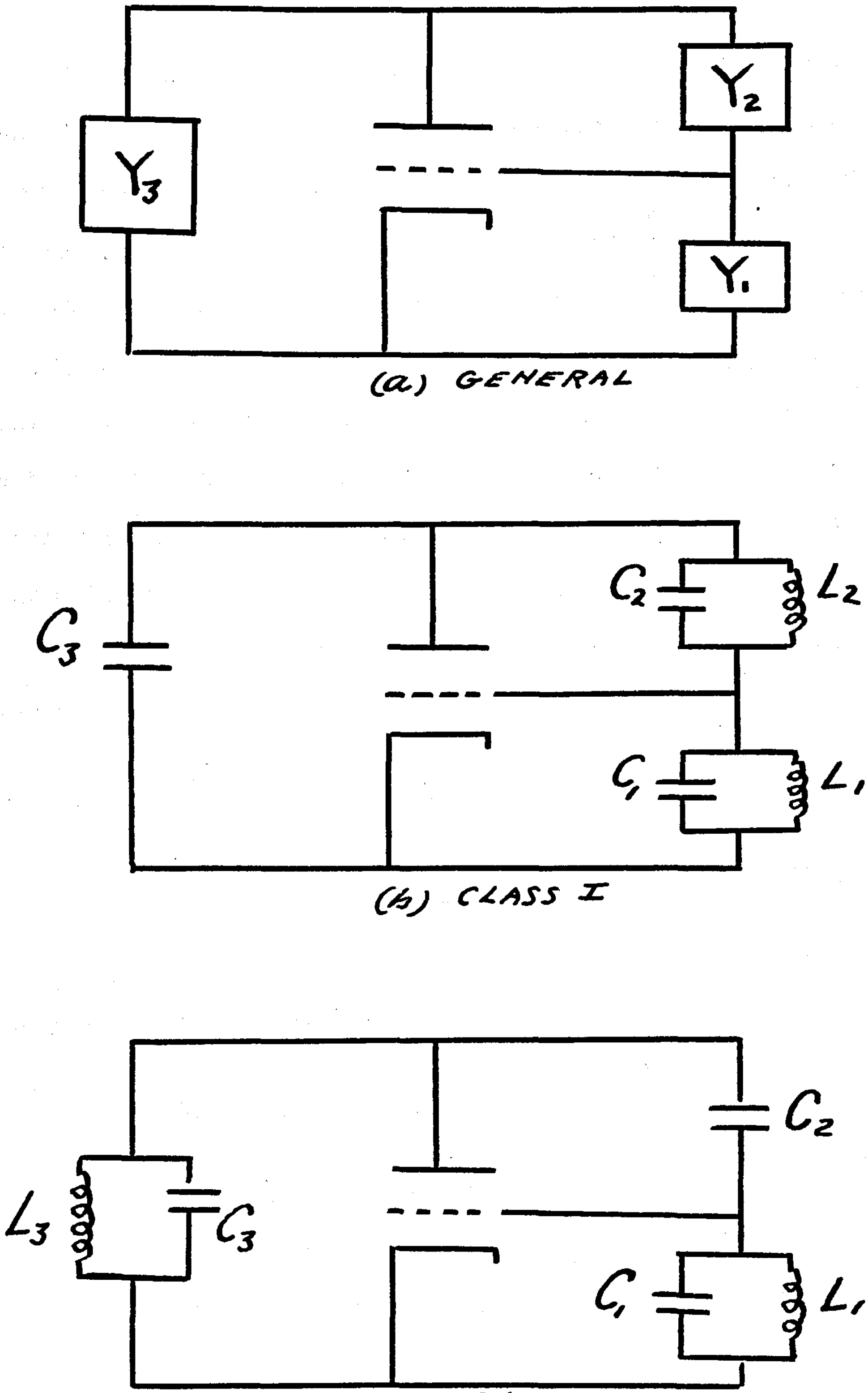
	b_1	b_2	b_3
Class I	+ (c)	-(L)	+ (c)
Class II	- (L)	+(c)	- (L)

Figures 2(b) and 2(c) are the respective circuits for Class I and Class II oscillators at ultra high frequencies, in which C_1 , C_2, and C_3 consist of interelectrode capacitance.

Consider the Class I oscillator of Fig. 2(b). The susceptance b_1 is capacitive. This capacitance in series with C_3 results in a net capacitance smaller than C_3 . The frequency of oscillation is determined by the external inductance L_2 resonated by the parallel combination of C_2 and some capacitance less than C_3 . In the Class II oscillator, Fig. 2(c), b_1 and b_3 must be inductive. The inductive susceptance b_1 in series with C_2 gives a resultant capacitance greater than C_2 (since b_3 is inductive, the combination of b_2 and b_1 must be capacitive). The resonant frequency is determined by the value of L_3 and the value of the parallel combination of C_3 and some capacitance greater than C_2 . It is thus apparent that higher frequencies are obtainable with Class I oscillators because they have smaller resonating capacitances.

Equation (28) may be written

$$g_2' = g_2 + g_0 . \qquad (29)$$



(a) GENERAL

(b) CLASS I

(c) CLASS II

figure 2

where $g_0 = \dfrac{g_1 g_3(g_1+g_3+g_M) + g_3 b_1^2 + g_1 b_3^2 - g_M b_1 b_3}{(g_1+g_3+g_M)^2 + (b_1+b_3)^2}$.

In a Class I oscillator, either b_1 or b_3 is variable and the other is fixed. It is possible to obtain g_0 as a function of b_1 for various values of b_3. It can be shown that as b_3 increases, the value of b_1 for maximum negative conductance increases so the effective susceptance, and, therefore, capacitance between the anode and the grid is increased. This means that a shorter circuit, or lower inductance, between the anode and the grid is needed for oscillations to occur at the same frequency. Thus the optimum value of susceptance may not be the value giving the maximum negative conductance but will probably be some smaller value.

In a circuit with negligible losses, g_2 and g_1 are zero. Then, since

$$g_3 = g_p = \frac{1}{r_p}$$

$$g_2' = \frac{g_p b_1^2 - g_M b_1 b_3}{(g_p+g_M)^2 + (b_1+b_3)^2} \qquad (30)$$

In order for oscillations to occur

$$g_M b_1 b_3 \geq g_p b_1^2 .$$

$$\text{but } g_M = \mu g_p$$

$$\therefore \mu b_3 \geq b_1$$

$$\text{or } \quad b_3/b_1 \geq \frac{1}{\mu} . \qquad (31)$$

These notes on the theory of the triode oscillator are derived from the paper on short wave oscillators by J. Bell, M. R. Gavin,

E. G. James, and G. W. Warren, (1).

CHAPTER III

FACTORS LIMITING THE FREQUENCY OF OPERATION OF THE TRIODE AMPLIFIER

1. Introduction.

There are three major factors that limit the upper frequency at which the conventional space charge control vacuum tube triode can amplify signals or generate oscillations. These limiting factors are

(a) Electron interelectrode transit times

(b) lead inductance

(c) interelectrode capacitance.

The manner in which these factors exert their influence will be considered in some detail in this chapter.

2. Transit time.

By the term transit time is meant that interval of time that elapses during the travel of an electron from one electrode to another in a vacuum tube. In the triode, of course, the cathode-to-grid and the grid-to-plate transit times must be considered.

The time involved in travel from one point to another is a function of the distance traveled divided by the average velocity during travel. In a vacuum tube, the distance between the electrodes is specified in the design of the tube. The velocity may be obtained as follows. The electric potential difference between two points is by definition the work expended in moving a unit charge from one point to another. Therefore the change in potential energy corresponding to the movement of an electron through a field is the product of the potential difference, E, through which it moves and the electron charge, e * Equating this to the kinetic energy gained,

$$Ee = \frac{1}{2} m v^2 \tag{32}$$

$$\text{whence} \quad v = \sqrt{\frac{2e}{m} E} \; . \tag{33}$$

where v is the velocity and m the mass of the electron.

W. R. Ferris, (14), demonstrates how to compute the transit times in a triode. Thus, to obtain the cathode-to-grid transit time, τ_1 , in a space-charge limited parallel-plane structure, the following equation applies:

$$\tau_1 = \int_0^a \frac{dx}{v} \tag{34}$$

in which τ_1 = the cathode-to-grid transit time

a = the cathode-to-grid spacing

v = the electron velocity at point x.

From equation (33), the electron velocity is

$$v = \sqrt{\frac{2e}{m} V}$$

where V = voltage at point x between cathode and grid.

Langmuir's equation for current density is

$$I = \frac{1}{9\pi} \sqrt{\frac{2e}{m}} \; \frac{V^{3/2}}{x^2} \; . \tag{35}$$

Then $$I = \frac{1}{9\pi} \sqrt{\frac{2e}{m}} \; \frac{V_g^{3/2}}{a^2} \tag{36}$$

* Dow, (9).

where V_g = grid potential (effective)

since the current density is the same at all points in a parallel plane structure. Then the voltage at any point x is

$$V = V_g \left(\frac{x}{a}\right)^{4/3} \tag{37}$$

$$\tau_1 = \frac{a^{2/3}}{\sqrt{\frac{2e}{m} V_g}} \int_0^a \frac{dx}{x^{2/3}} \tag{38}$$

$$\therefore \tau_1 = \frac{3a}{\sqrt{\frac{2e}{m} V_g}} = \frac{3a}{5.95 \times 10^7 V_g^{1/2}} \tag{39}$$

where τ_1 is in seconds, a is in centimeters and V_g is in volts.

This is one and one-half times the time required with uniformly accelerated electrons, as in the case of no space charge.

The transit time between grid and plate, τ_2, is easier to calculate because the potential distribution between the electrodes is linear. Thus,

$$\tau_2 = \frac{b}{\frac{1}{2}(v_g + v_p)} \tag{40}$$

where b = grid-to-plate spacing

v_g = velocity at the grid

v_p = velocity at the plate

Hence $$\tau_2 = \frac{2b}{5.95 \times 10^7 (V_g^{1/2} + E_b^{1/2})} \tag{41}$$

in which E_b = dc plate voltage, and is the effective plate voltage.

It is also possible to calculate the transit times for concentric cylinder electrodes. Thus,

$$\tau_1 = \int_{r_0}^{r_1} \frac{dr}{v} \tag{42}$$

in which r_1 = radius of the grid

r_0 = radius of the cathode

r = radius at a point between grid and cathode

v = electron velocity at r .

$$\text{but} \quad v = \sqrt{\frac{2e}{m} V} \; . \tag{33}$$

The current density is now given by

$$I = \frac{2}{9}\sqrt{\frac{2e}{m}}\,\frac{V^{3/2}}{r\beta^2} = \frac{2}{9}\sqrt{\frac{2e}{m}}\,\frac{V_g^{3/2}}{r_1\beta_1^2} \tag{43}$$

$$\text{where } \beta = f\left(\frac{r}{r_0}\right),$$

$$\beta_1 = f\left(\frac{r_1}{r_0}\right).$$

The function β is defined by Langmuir and Blodgett, (22), and tabulated for several values of the radius ratio. From equation (43)

$$V = V_g\left(\frac{r\beta^2}{r_1\beta_1^2}\right)^{2/3} . \tag{44}$$

$$\therefore \tau_1 = \frac{(r_1\beta_1^2)^{1/3}}{\sqrt{\frac{2e}{m}V_g}} \int_{r_0}^{r_1} (r\beta^2)^{-1/3}\,dr . \tag{45}$$

$$\text{whence } \tau_1 = \frac{K(r_1 - r_0)}{5.95 \times 10^7 V_g^{1/2}} . \tag{46}$$

Langmuir and Blodgett obtained values for K of 1.5 to 3 for cathodes internal to the grid and 3 to ∞ for cathodes external to the grid.

The grid-to-plate transit time, τ_2, is

$$\tau_2 = \frac{2 r_2}{5.95 \times 10^7 \sqrt{E'}} \frac{1}{\epsilon^{E_b/E'}} \int_{\sqrt{\frac{V_g}{E'}}}^{\sqrt{\frac{E_b}{E'}}} \epsilon^{x^2} dx . \tag{47}$$

in which r_2 = radius of the plate

and $E' = \dfrac{E_b - V_g}{\log r_2/r_1}$

For small values of r_2/r_1, on the order of less than 2, τ_2 may be found approximately by assuming the plate and grid to be parallel planes. Then,

$$\tau_2 = \frac{2(r_2 - r_1)}{5.95 \times 10^7 (V_g^{1/2} + E_b^{1/2})} . \tag{48}$$

The calculation of the effective voltage at the grid, V_g, is done by Ferris with the result

$$V_g = \frac{3}{2} \frac{I_b}{g_m} \frac{dV_g}{dE_c} \tag{49}$$

$$\text{where } \frac{dV_g}{dE_c} = \frac{1}{1 + \frac{1}{\mu} + \frac{4}{3\mu} \frac{b}{a}} \quad \text{for plane electrodes} \tag{50}$$

$$\text{and } \frac{dV_g}{dE_c} = \frac{1}{1 + \frac{1}{\mu} + \frac{2}{3\mu} \log \frac{r_2}{r_1}} \tag{51}$$

for cylindrical electrodes. I_b, μ, and g_m are determined experimentally. The effective plate voltage is simply the direct voltage applied to the plate.

3. Effects of transit time on tube admittances.

In considering vacuum tube theory at low frequencies, there is little need to take into account the effect of transit time of electrons in their travel from cathode to anode. This time is such a negligible fraction of the period of one such low frequency oscillation that, for all practical purposes, the transfer of electrons can be regarded as occuring instantaneously. This reasoning leads to a conception of plate current as being a conduction current, varying with the rate of arrival of the electrons at the plate. It might then be assumed that no current flows in the plate circuit until the electrons actually arrive at the plate. This viewpoint, acceptable enough at low frequencies, is no longer satisfactory at frequencies where the transit time is in appreciable portion of a cycle.

The current flow to an electrode is better regarded as the result of the motion of charges in the interelectrode space as explained by B. J. Thompson, (36). Consider, for example, two infinite parallel-plane electrodes, separated a distance, d, with a voltage, E, between them as shown in Fig. 3(a). Then the electric field, F, is given by the expression

$$F = \frac{E}{d} \quad . \tag{52}$$

If a small positive charge is placed between the plates very close to the positive plate, there will be an increase in the charge of the positive plate of amount $-q$ and no change in the charge on the negative plate. A force of magnitude Fq will tend to move the charge towards the negative plate. If the charge moves, the work done on it will be Fqx, the product of the field times the charge times the distance, x, through which it moves. This work is supplied by the battery and is equal to its voltage, E, times the change in charge induced on one of the plates. If q_n represents the charge induced on the negative plate, then

$$-Eq_n = Fqx \tag{53}$$

$$-Eq_n = \frac{E}{d}qx \tag{54}$$

$$\therefore \quad q_n = -\frac{qx}{d}. \tag{55}$$

This condition is illustrated in Fig. 3(b). Thus the charge induced on the negative plate is proportional to the charge in the space between the plates and to the fraction of the total distance between the plates through which the charge has moved. The total charge induced on the two electrodes is equal in magnitude and opposite in sign to the space charge.

The current flowing to the negative plate as a result of the motion of the charge q is equal to the rate of change of the charge q_n induced on it by this motion. That is,

$$i_n = \frac{dq_n}{dt} \tag{56}$$

$$i_n = -\frac{q}{d}\frac{dx}{dt} = -q\frac{v}{d}. \tag{57}$$

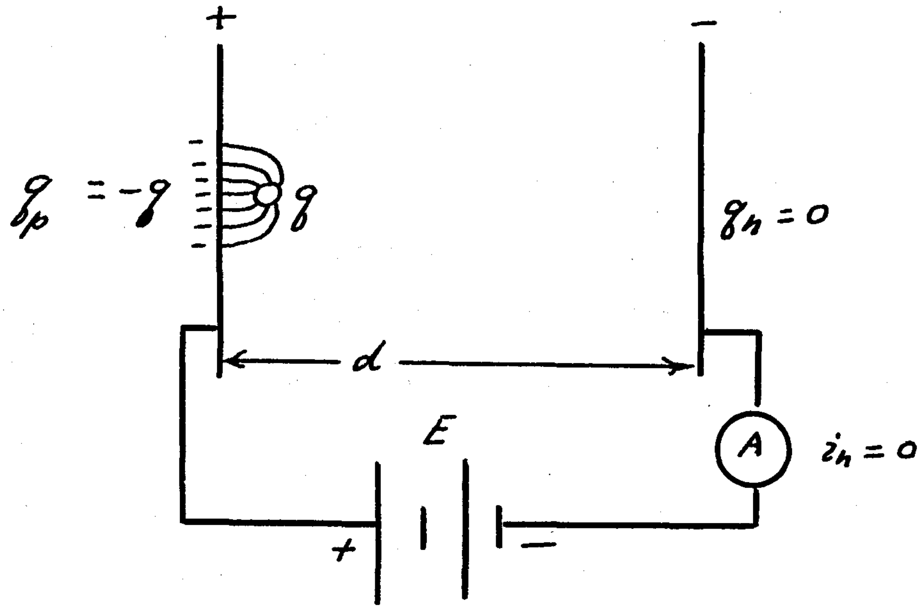


figure 3(a)

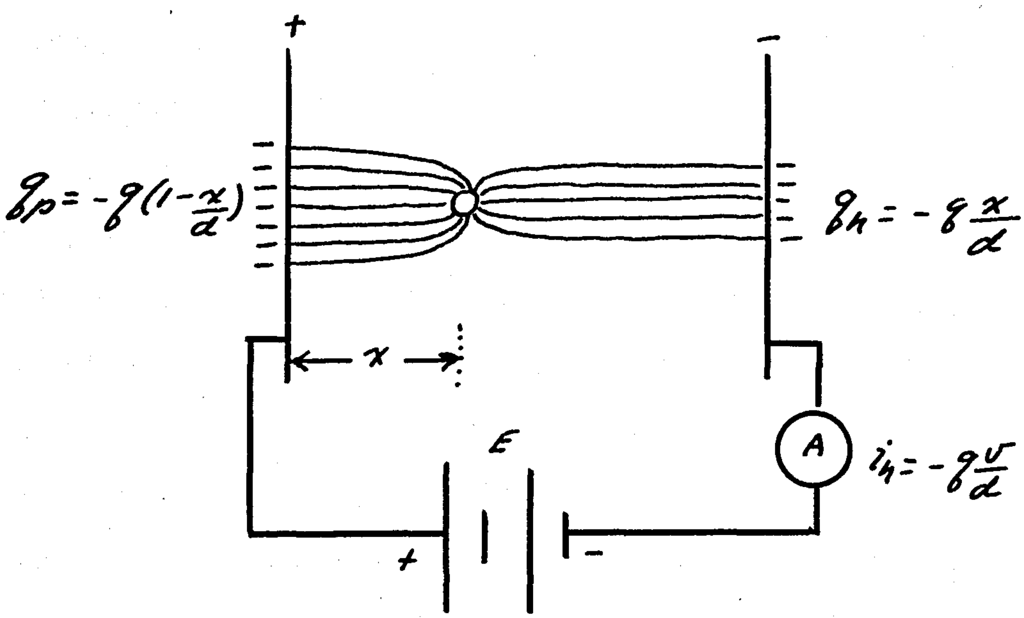


figure 3(b)

Thus, the current flowing as a result of the motion of a charge between two plates is equal to the product of the charge times the velocity divided by the distance between the plates. The current flow to the positive plate is always instantaneously equal in magnitude to the current flow to the negative plate.

W. R. Ferris, (14), applied these considerations to the triode of Fig. 4. Operated under direct current conditions and with a negatively biased grid is the triode shown in Fig. 4(a). The current induced in the grid by the motion of the electrons from the cathode towards the grid is exactly balanced by the current induced by the electrons flowing away from the grid towards the plate. Since no electrons are collected by the negative grid, there is zero current in the grid.

In Fig. 4(b) is shown the same triode with a small alternating signal impressed on the negatively biased grid. The electron convection current is a function of the instantaneous potential of the grid. Therefore, during the time that the grid potential is increasing, the convection current is increasing. Since the convection current, or flow of electrons, originates at the cathode and propagates at finite velocity across the interelectrode space, it is greatest nearer the cathode and least near the plate during the period that the grid potential is constantly increasing. In other words, there are more electrons, moving faster, between the cathode and the grid, than between the grid and the plate. Therefore, the current induced in the grid by the electrons moving towards it is greater than that induced in the other direction by the electrons moving away from it and there is a net flow of current into the grid. When the grid potential

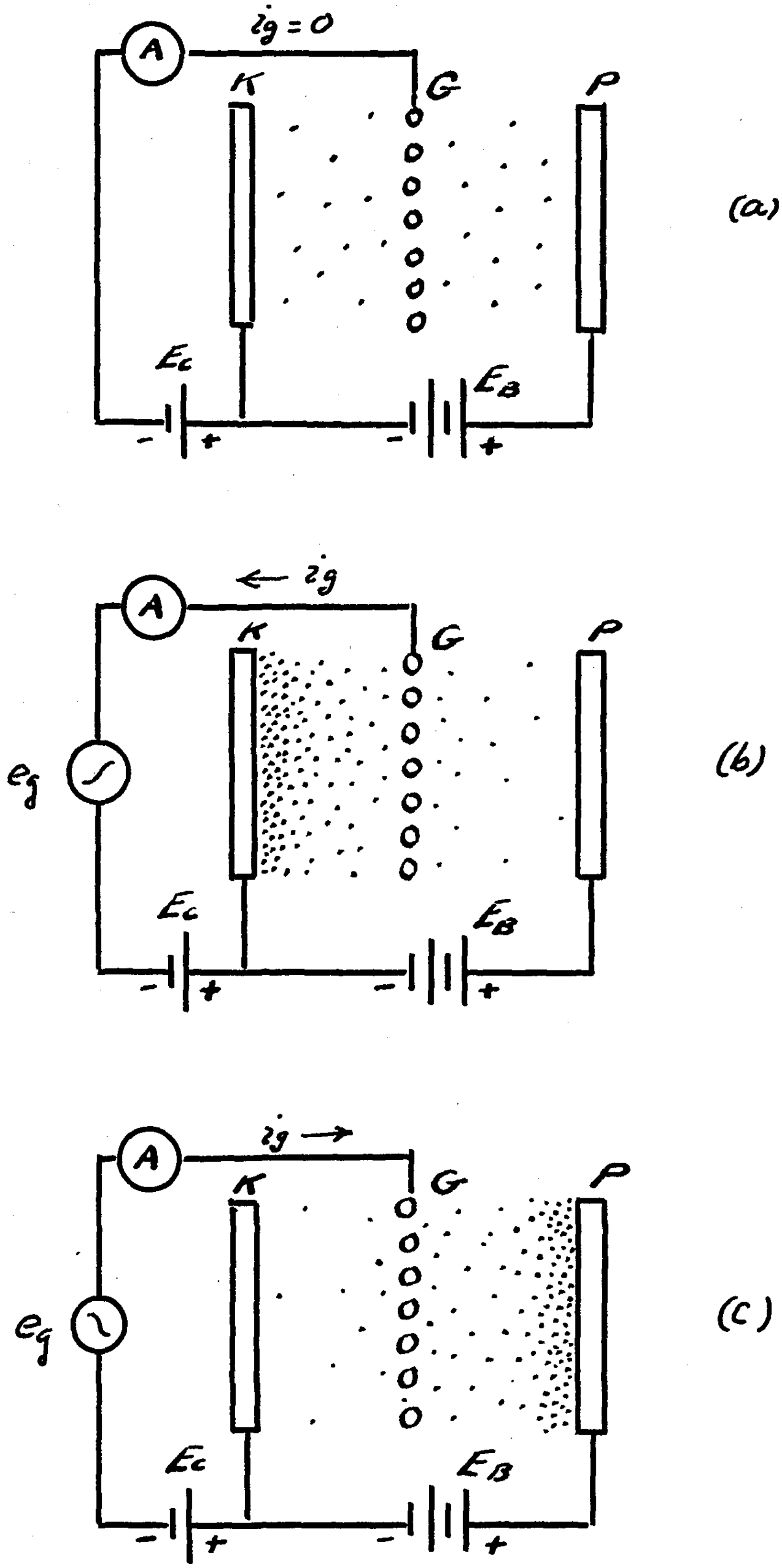


figure 4

is decreasing, as shown in Fig. 4(c), opposite conditions obtain and there is a net current flowing out of the grid.

At any instant the difference between the convection currents on the two sides of the grid measured at two arbitrary points (which can be taken very close to the grid) is proportional to the time rate at which the convection current is modulated. This is the rate of change of grid potential times the transconductance, g_M. Furthermore, the rate of change of grid voltage is proportional to the root mean square magnitude of grid voltage times the frequency of the alternating grid voltage. The difference is also proportional to space modulation of the convection current, which is a function of the transit time between the two points. Then the grid current can be expressed as

$$I_g = K_1 E_g g_M f \tau \tag{58}$$

$$\text{or } I_g = Y_g E_g \tag{59}$$

$$\text{where } Y_g = K_1 g_M f \tau$$

I_g = alternating grid current

E_g = alternating grid potential

g_M = transconductance

f = frequency of alternating grid potential

τ = transit time, cathode-to-grid

Y_g = grid admittance

K_1 = undetermined parameter

From the foregoing, it is apparent that Y_g is a capacitive susceptance.

At the moment when the grid potential reaches its maximum value, the convection current is still greatest at the cathode and least at the plate. Therefore there is still a net convection current in the grid circuit. There is a finite time, a function of transit time, τ, before this current becomes zero. This results in a phase displacement, θ , between the time of maximum grid potential and time of zero grid current. This phase shift represents a conductive component, g_g , in the grid admittance which may be defined as

$$g_g = Y_g \sin\theta \qquad (60)$$

for small θ, $\sin\theta \approx \theta$

$$\therefore\ g_g \approx Y_g\,\theta. \qquad (61)$$

$$\theta = \omega\tau \qquad (62)$$

where $\omega = 2\pi f$

$$\therefore\ g_g = K g_m f^2 \tau^2. \qquad (63)$$

In other words, the grid conductance increases as the square of the frequency and of the transit time.

This variation of conductance with frequency was predicted by B. J. Thompson while studying small "acorn" tubes. The effect was noted by H. O. Peterson in the damping of a tuned circuit at high frequencies as a result of which Thompson derived an equation in the above form for g_g . The results were verified experimentally by W. R. Ferris and D. O. Nath. North, (31), derived a trigonometric equation for grid conductance in which K is found to be a function of the transit times and velocities at the electrodes. Thus

$$K = f\left(\frac{\tau_2}{\tau_1} \cdot \frac{v_g}{v_p}\right)$$

τ_2 = grid-plate transit time

τ_1 = cathode-grid transit time

v_g = electron dc velocity at the grid

v_p = electron dc velocity at the plate .

As for the input susceptance, the investigations of many engineers, Benham, (2), Llewellyn, (26), and North, (31), to mention a few, result in two common conclusions:

(a) There is an increase in capacitance due to the presence of electrons in the cathode-grid space. This leads to a so-called "hot" capacitance greater than the cut-off or "cold" capacitance, and

(b) This increase is independent of frequency for all practical purposes. North obtained a theoretical variation in capacitance of 1.2% for a frequency variation from zero cycles per second to one hundred megacycles per second. Ferris was unable to detect experimentally any variation at all, although the difference between hot and cold capacitance was readily observed.

Transit time effects are also noted in the grid-plate space. Llewellyn, (27), describes an equivalent circuit for the triode at ultra high frequencies which takes the same form as at low frequencies in which the tube is represented as a voltage generator, μe_g, in series with the internal resistance of the tube, r_p. Now, however, μ is a complex quantity and so is the internal impedance of the tube, Z_p. In other words, a phase angle is introduced in the driving electromotive force of the tube and another phase angle is introduced in the resultant plate current. The transconductance of the tube is defined for low frequencies as

$$g_M = \frac{\mu}{r_p}.$$

At frequencies where μ is complex and r_p must be replaced by the complex Z_p, the term transadmittance replaces the term transconductance since a phase angle is involved. Transadmittance is

$$Y_M = \frac{\mu}{Z_p}.$$

The magnitude of this transadmittance decreases with increasing frequency but the effect is small.

In summary, the principal effects of increasing transit time are

(a) Increased input conductance

(b) introduction of phase angle in transadmittance

(c) decreased magnitude of transadmittance

(d) decreased magnitude of plate impedance and the introduction of a phase angle therein.

The most important of these effects is the increased input conductance, which varies as the square of the frequency and of the transit time. The introduction of a phase angle, increasing with transit time, in the transadmittance is also important. It is apparent that transit time effects cause variations in all the admittances which are involved in the operation of the tube either as an amplifier or as an oscillator.

2. The effects of transit time on amplifier and oscillator efficiency.

Because the gain of the amplifier is inversely proportional to the input conductance, it becomes exceedingly difficult to obtain usable amplification at ultra high frequencies. In addition, the gain is proportional to the magnitude of the forward admittance, which is decreased slightly due to the effects of transit time. Similarly, since the input voltage for the oscillator is derived from its plate voltage, the input loading seriously limits the maximum frequency at which the tube can oscillate.

The phase angle introduced in the transadmittance by transit time effects is serious only for the oscillator. It is for this reason that higher efficiencies are obtainable for the triode when operated as an amplifier rather than as an oscillator under similar operating conditions.

The effects of transit time are emphasized in Class C operation, because the electrons travel towards the plate in pulses rather than in the continuous stream encountered in Class A operation. For this reason, the following development is based on Class C operation.

In an oscillator the Rf grid and plate voltages are derived from the same resonant circuit. For this reason, it is very difficult to shift the phase of the plate voltage with respect to the grid voltage for even a few degrees. Due to transit time, electrons arrive after the plate voltage has swung through a minimum and starts increasing. The energy given up at the plate is the total work done on the electrons by the voltage present during flight. In this case, the voltage is approximately equal to the plate voltage at the time of arrival of the electrons at the plate. The greater the phase difference between the time of the plate voltage minimum and the time of the electron arrival at the plate, the higher are the energy losses and the lower is the efficiency. In a neutralized amplifier, on the other hand, or any amplifier in which the phase of the output voltage is independent of the phase of the input voltage, the plate voltage phase may be shifted for minimum loss (180 degrees phase difference between alternating plate voltage and plate current). Therefore the efficiency of the triode as an amplifier will be higher than as an oscillator. It will be lower, however, than under low frequency conditions because the electrons are in the grid-anode space for a larger fraction of the Rf cycle.

The efficiency of the oscillator is zero when T_1, the period of oscillation, is equal to some constant, A, times the transit time, τ, where τ is the total transit time from cathode to anode. M. R. Gavin, (16), determined that A is approximately equal to 2. In other words, oscillations cease when the transit time is about one-half the period of oscillation. This result is obtained as follows, for planar triodes:

$$\lambda = cT \tag{64}$$

$$T_{MIN} = A\tau \tag{65}$$

from eq (39) and (40), $$\tau = \frac{3a}{5.95\times10^7 V_g^{1/2}} + \frac{2b}{5.95\times10^7 (V_g^{1/2}+E_b^{1/2})} \tag{66}$$

$$\therefore \lambda_{MIN} = cA\tau = k\left(\frac{3a}{V_g^{1/2}} + \frac{2b}{V_g^{1/2}+E_b^{1/2}}\right) \tag{67}$$

where $k = \dfrac{cA}{5.95\times10^7}$.

$\lambda =$ wavelength

$c =$ velocity of light

$T =$ period of oscillation

From measurements of λ_{MIN}, a, b, V_g and E_b, k was determined to have a value of about 1000. From this,

$$A = \frac{5.95 \times 10^7 \times 10^3}{3 \times 10^9} \simeq 2 .$$

This value of A = 2 is satisfactory for cylindrical electrode triodes when the interelectrode spacing is very small so that the ratios of grid radius to cathode radius and anode radius to grid radius are each less than 2.

V_g, the effective voltage at the grid, has been shown in equation (49) to be inversely proportional to μ. Then it is apparent from equation (67) that the minimum wavelength is proportional to the square root of μ. Therefore, low μ tubes are desirable for ultra high frequency oscillation.

The data of Wagener, (38), and Haeff, (18), show efficiency to be an almost linear function of frequency in accordance with an empirical relation which can be adjusted to fit both the amplifier and the oscillator. Thus,

$$\eta = \eta_0 (1 - K_1 f) \tag{68}$$

in which $\eta =$ efficiency at frequency f

$\eta_0 =$ low frequency efficiency

$K_1 =$ constant, determined by operating and design parameters

$f =$ frequency .

Wang, (39) defines four modes of Class C operation from the low frequency case to the ultra high frequency where the efficiency becomes zero. These modes are determined by the relations between the following periods of time involved in the flow of a pulse of electrons from the emitter to the anode;

τ_1, the time when the first electron emitted at the cathode at equal zero arrives at the plate;

τ_2, the time when the cathode stops emitting electrons;

τ_3, the time when the plate receives the last electron; and

τ_4, the time when the last electron is returned to the emitter and the interelectrode space is empty.

In the low frequency case, transit times are very small and the tube operation can be understood from its static characteristics. In the second mode, or moderately high frequency case, τ_1 is still less than τ_2. The third mode occurs at higher frequencies when τ_1 is greater than τ_2. No electrons reach the plate in the fourth mode and it is the boundary between the third and fourth modes that sets the limit for pulsed operation. The mode of operation for a specified waveform depends on a single normalization constant, C, where

$$C = 2\frac{e}{m}\frac{\tau_0^2 E_{gm}}{d^2} \tag{69}$$

in which e/m = ratio of charge to mass of an electron

τ_0 = $\frac{1}{2}$ period of voltage pulse

E_{gm} = peak value of voltage pulse

d = equivalent diode spacing .

At the boundary between the third and fourth modes, C has a value of

0.7.

Lehmann and Vallarino, (25), extended these results by describing the voltage pulse in terms of anode voltage. Their work is based on operation in the second mode. In the third mode, the efficiency is less than 25% and the amplitude of oscillation is small so that Class A theory is applicable. This voltage pulse in the cathode-grid region is proportional to E_b, the dc plate voltage, assuming practical Class C operating conditions. On this basis, the properties of a Class C ultra high frequency amplifier and oscillator may be expressed in terms of the dimensionless parameter ϕ, where

$$\phi = \frac{fd}{\sqrt{E_b}} \tag{70}$$

$$\therefore K_1 f = \frac{Kfd}{\sqrt{E_b}} \tag{71}$$

$$\text{and} \quad \eta = \eta_0 \left(1 - \frac{Kfd}{\sqrt{E_b}}\right). \tag{72}$$

η_0 = low frequency efficiency,

f = frequency in megacycles,

E_b = dc anode voltage in volts,

d = effective cathode to grid spacing in inches,

K = 1.75 for oscillator
1.2 for amplifier

K was determined empirically for oxide coated cathodes.

This analysis of the work of Wagener, Haeff, Wang, and Lehmann and Vallarino was made by R. R. Law, (24).

Lehmann and Vallarino demonstrated other interesting relations.

From the condition that the efficiency, η , is a constant,

$$\phi = \frac{fd}{\sqrt{E_b}} = \text{constant}.$$

Using the symbol A as the current density per unit area and from the Child - Langmuir law,

$$A = K_2 \sqrt{\frac{e}{m}} \, \frac{V^{3/2}}{d^2} \qquad (73)$$

$$\text{from eq. (69)} \quad K_3 = \frac{m}{e} \, \frac{d^2 f^2}{V} \, . \qquad (74)$$

eliminating d from the above equations

$$V = K_4 \frac{A^2}{f^4} \, . \qquad (75)$$

which shows the maximum output plate voltage for a constant efficiency. Eliminating V from equations (73) and (74)

$$d = K_5 \frac{A}{f^3} \qquad (76)$$

which is the grid to cathode distance when the efficiency is constant. This shows how much the dimensions of the tube must be reduced for constant efficiency with increasing frequency. For practical reasons, the dimensions can be reduced only so much. The current can be shown to be proportional to a constant times a function of current density and frequency. Thus

$$I = K_6 \frac{A^3}{f^6} \, . \qquad (77)$$

W, the useful power, becomes

$$W = VI = K_7 \frac{A^5}{f^{10}} \quad . \tag{78}$$

This does not correspond to the exact conditions under which the tubes are used but it does indicate to some degree the effect of increasing frequency in decreasing power output quickly. Equations (75) through (78) indicate the desirability of having maximum possible current density, A .

5. Lead inductance and interelectrode capacitance.

In the circuit of Fig. 5(a) is shown a triode with the lead inductances and interelectrode capacitances indicated therein. An oscillator circuit may be constructed using for the tuned circuit the inductance L where

$$L = L_a + L_g \tag{79}$$

and the capacitance C where

$$C = C_{ag} + \frac{C_{ak} C_{gk}}{C_{ak} + C_{gk}} \quad . \tag{80}$$

The wavelength, λ , of oscillation is

$$\lambda = 60\pi \sqrt{LC} \tag{81}$$

when L is in uhenries and C is in uufarads.

L_a and L_g , the plate lead and grid lead inductances, respectively, can be calculated from the formula for the inductance of a straight wire

$$L = .002\, l \left(\log_e \frac{4l}{d} - 1 \right) \text{uhenries} . \tag{82}$$

in which ℓ is the length in cm

d is the diameter in cm.

From equation (81) it is apparent that L and C must be made as small as possible in order to decrease the wavelength, λ.

Lead inductance has another undesirable effect on the tube operation because it can result in an input loading which is in addition to that caused by the effects of transit time. Consider the circuit shown in Fig. 5(b). The input voltage E_1 is applied to the grid as shown. A portion of this voltage, E_{gk}, appears across the grid-to-cathode admittance, g_{gk} in parallel with C_{gk}, and the remainder, E_k, appears across the cathode lead inductance, L_k. Then

$$E_1 = E_{gk} + E_k \tag{83}$$

$$E_k = j\omega L_k i_k \tag{84}$$

$$i_k = E_{gk} g_{gk} \tag{85}$$

$$\therefore \quad E_1 = E_{gk}(1 + j\omega L_k g_{gk}) \tag{86}$$

also $$i_{gk} = j\omega C_{gk} E_{gk} \tag{87}$$

$$Y_a = \frac{i_{gk}}{E_1} = \frac{j\omega C_{gk}}{1 + j\omega L_k g_{gk}} \tag{88}$$

where Y_a is the component of input admittance due to i_{gk}.

$$Y_a = \frac{\omega^2 L_k C_{gk} g_{gk} + j\omega C_{gk}}{1 + \omega^2 L_k^2 g_{gk}^2} \tag{89}$$

$$g_a = \frac{\omega^2 L_k C_{gk} g_{gk}}{1 + \omega^2 L_k^2 g_{gk}^2}. \tag{90}$$

generally, $\omega^2 L_k^2 g_{gk}^2 \ll 1$.

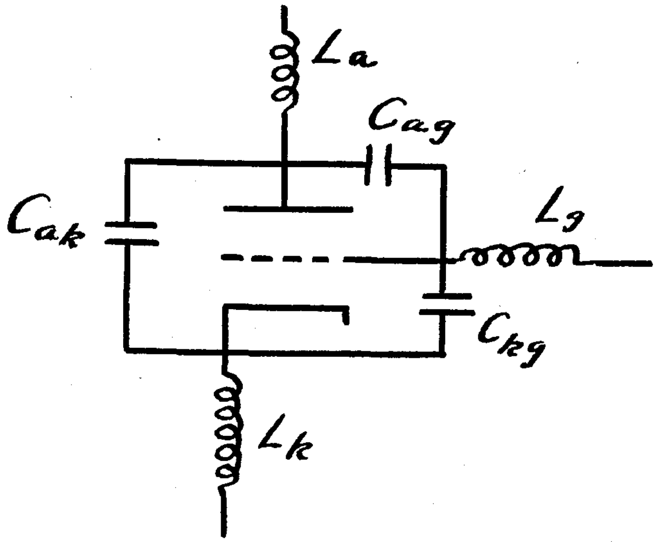


figure 5 (a)

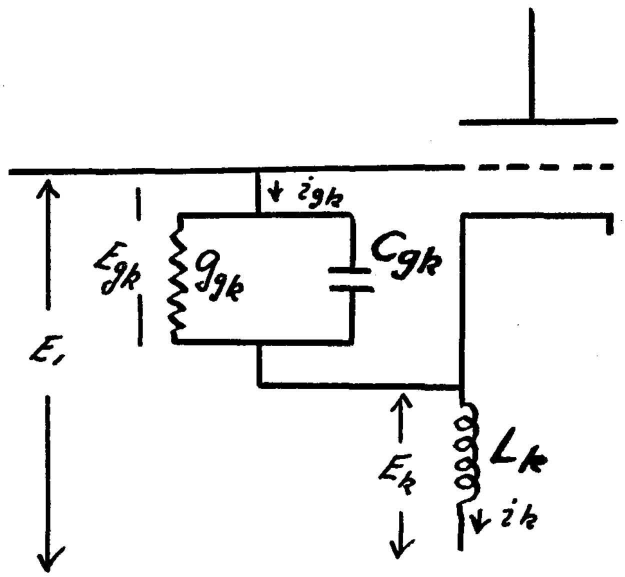


figure 5 (b)

$$\therefore \quad g_a = \omega^2 L_k C_{gk} g_{gk} \quad . \tag{91}$$

Thus there is an input conductance, g_a, which is proportional to frequency squared and to the grid-cathode transconductance. When inductance in the plate lead is considered, it can be shown that the input conductance is proportional to the square of the frequency and to the grid-to-plate transconductance, assuming that the two LC products involved are equal. Since the transit time input conductance is also proportional to the square of the frequency and to the transconductance, it is difficult to distinguish the respective components.

CHAPTER IV

INCREASING THE FREQUENCY RANGE OF THE TRIODE

1. Introduction

It has been demonstrated how transit time, lead inductance and tube capacitance affect the operation of the triode as an amplifier and as an oscillator. Obviously, to increase the upper frequency limit of the tube, it is necessary to decrease the magnitude of these three factors or otherwise compensate for their effects. It is possible to reduce transit time, lead inductance and interelectrode capacitance by reducing the physical dimensions of the tube. In addition, the effects of lead inductance can be eliminated and the effects of capacitance minimized by integrating the tube with its associated circuit.

2. Decreasing the linear dimensions of the tube.

The limiting effects of lead inductance and interelectrode capacitance have been appreciated almost since the invention of the triode. In 1916, W. C. White, (40), described a triode oscillator in which the oscillation inductances and capacitances consisted solely of these components. He obtained a wavelength of about six meters, which represented the upper frequency limit of a conventional triode using lumped constant circuits for many years. It was not until 1933, in fact, that a major development in tube design was made which increased the upper frequency limit by decreasing the linear dimensions of the tube. In that year, B. J. Thompson and George M. Rose, Jr., (37), described such a tube. Its construction was a result of the application of what B. Salzberg, (34), called "The Principle of Similitude" which is stated as follows:

If all the linear dimensions of a tube structure are divided by a constant factor, n, then the electrode currents, transconductance, amplification factor, and plate resistance will remain substantially constant, but the lead inductances and capacitances, the tube capacitances, and the time of passage of the electrons between the various electrodes will be divided by n. The direct current lead resistances will be multiplied by n, but the alternating current lead resistances will ordinarily be multiplied by something less than this factor. The allowable plate dissipation and available emission of the tube, however, will be divided by n^2 and the current densities will be multiplied by n^2. Physically the tube will be reduced in its overall dimensions by a factor n and its weight by a factor n^3.

The tube built by Rose and Thompson used planar electrodes. The reduction in dimensions over existing triodes was ten fold and so was the reduction in resultant wavelength (from 5 meters to 50 centimeters). Although this tube was designed and built more or less as a laboratory experiment without immediate commercial application, the response from the electronics world was such that further development was warranted. The so-called "acorn" tube, described by Salzburg and Burnside, (34), was the result.

3. Integrating the tube with the circuit.

There is a limit to how far the dimensions of the tube can be reduced. Reasonable plate dissipation must be allowed for, which is not consistent with very small dimensions. In addition, the grid to cathode spacing can not be reduced to any advantage beyond a point at which the effective grid voltage no longer influences the emission of electrons from the virtual cathode. It is possible to reduce the effects of lead inductance and interelectrode capacitance.

I. E. Mouromtseff and H. V. Noble, (30), designed an oscillator in 1932 in which the external circuit between the plate and the grid consisted of a length of transmission line of characteristic impedance Z_o terminated in an impedance Z_t. Refer to Fig. 6. The input impedance of the line, Z_i, is given by the familiar

$$Z_i = Z_o \frac{Z_t + jZ_o \tan \frac{2\pi l}{\lambda}}{Z_o + jZ_t \tan \frac{2\pi l}{\lambda}} . \tag{92}$$

if $Z_t = \infty$, then

$$Z_i = \frac{Z_o}{j \tan \frac{2\pi l}{\lambda}} . \tag{93}$$

if l, the length of the line, is $\frac{\lambda}{2}$, then

$$Z_i = \frac{Z_o}{0} = \infty . \tag{94}$$

This represents a parallel resonant circuit. However, there is a loading effect imposed by the tube across the input terminals of the transmission line. Even if the plate lead and grid lead inductances are eliminated by incorporating the leads in the transmission line, it is still necessary to consider the interelectrode capacitances. Mouromtseff and Noble neglected this loading, but Gavin, (16), did not. He used a shorted quarter wave section of line so that Z_t equals zero. Then

$$Z_i = jZ_o \tan \frac{2\pi l}{\lambda} . \tag{95}$$

$$Z_i = \infty \quad when \quad l = \frac{\lambda}{4}. \tag{96}$$

It is necessary to adjust the length, l, to something less than a quarter wave to resonate the input impedance of the line with the capacitive loading due to the tube. Then

$$\frac{1}{\omega C} = Z_i = Z_o \tan \frac{2\pi l}{\lambda}, \tag{97}$$

$$C = C_{ag} + \frac{C_{ak} C_{gk}}{C_{ak} + C_{gk}}, \tag{80}$$

$$\therefore \tan \frac{2\pi l}{\lambda} = \frac{1}{\omega C Z_o}. \tag{97a}$$

In order to produce ultra short waves, it is necessary that the length, l, be as long as possible for a given wavelength. This requires small values of characteristic impedance. The use of concentric transmission line is therefore suggested. The characteristic impedance of such a line is

$$Z_o = \frac{138}{\epsilon} \log_{10} \frac{D}{d} \tag{98}$$

ϵ = dielectric constant (unity for air)

D = inner diameter of the outer conductor

d = outer diameter of the inner conductor.

By designing a tube with concentric electrodes expressly for integrating the grid and plate with inner and outer conductors, respectively, and having the anode to grid diameter ratio the same as the diameter ratio of the concentric line, the plate and grid lead inductance and the plate-to-grid capacitance, C_{ag} , are no longer factors limiting the frequency of the tube. There still remains, however, the series combination of plate-to-cathode capacitance and grid-to-cathode capacitance. If the characteristic impedance of the tube elements does not match that of the line, and it generally does not, there will exist an additional component of impedance to load the input terminals of the line. In any event, what is achieved by such a design is a single electrical system having one section walled off and evacuated to house the electronic activity rather than an electron tube with an attached outside circuit.

A resonant length of shorted transmission line will develop standing waves of voltage and current when an alternating current electromotive force of frequency corresponding to $\lambda = 4\ell$ is impressed across the open end, as in Fig. 6(b). These oscillations will continue when the source of energy is removed until the energy is dissipated in the resistance of the conductors, in radiation, or in the load. In a concentric line system such as this, losses in conductor resistance and in radiation resistance are negligible. The phase relations at the ends of the concentric line system are such that the tube connected as shown in Fig. 6(c) can sustain oscillations if a dc voltage is supplied to the plate and if the grid is properly biased. The standing waves developed are the result of the transmitted and reflected waves propagated in such a manner that every impulse originated by the tube returns to it after traveling the length of the system and back, a distance 2ℓ. The time required is

$$t = \frac{2\ell}{v} . \qquad (99)$$

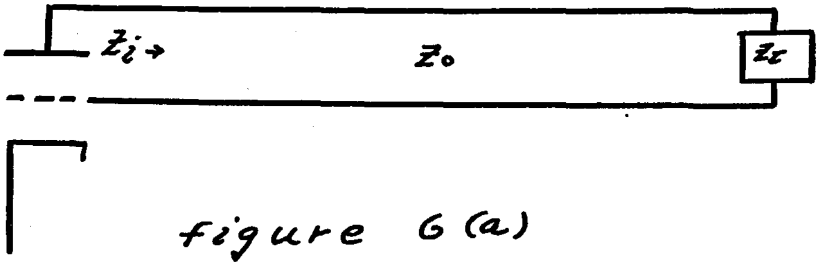


figure 6 (a)

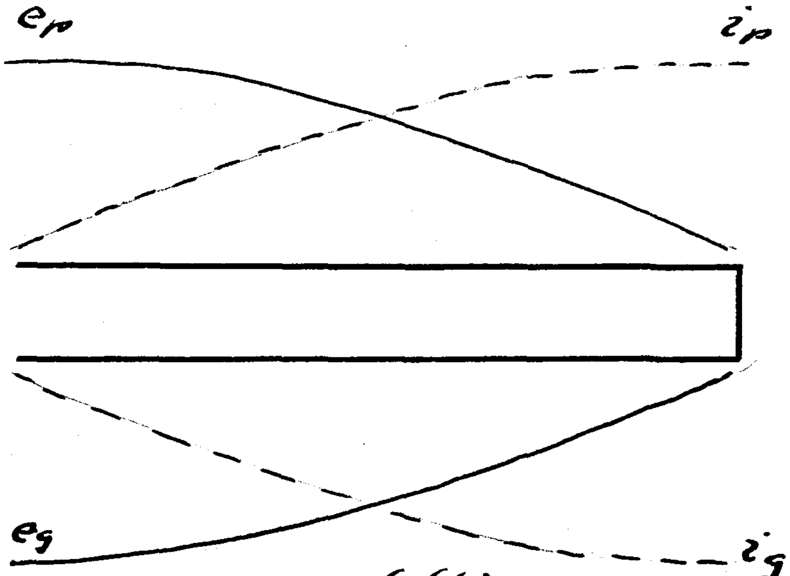


figure 6 (b)

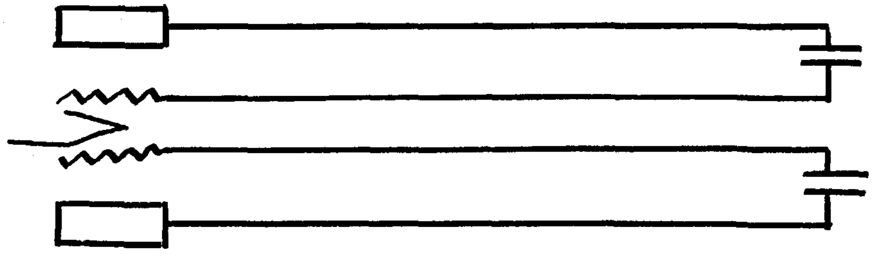


figure 6 (c)

v = the velocity of propagation of the wave and is approximately equal to c , the velocity of light.

An impulse can be produced by the tube only on a negative half cycle of plate voltage and positive swing of grid voltage. Therefore, T is the period of one half oscillation and

$$2\ell = \frac{vT}{2} = \frac{\lambda}{2} \; . \tag{100}$$

Thus the wavelength is a function of the length of the transmission line only. This means that ultra short waves can be produced in such a system. A limitation on the minimum wavelength is imposed by the necessity for not having a voltage node on the grid. Blocking condensers are used to isolate the dc voltage on the plate from the grid circuit.

A comparison between the concentric line, or standing wave, circuit and the conventional lumped constant Hartley oscillator can be made by calculating the oscillator energy in each case for the same value of maximum voltage. Consider a concentric line of length, ℓ , having distributed capacitance and inductance of C farads per unit length and L henries per unit length, respectively. Sinusoidal standing waves of voltage and current are imposed on the line. The total energy of such a system consists of the sum of the potential energy and the kinetic energy. The potential energy is a maximum and the kinetic energy zero when the current is everywhere zero and the voltage standing wave is a maximum. Then

$$W_p = \int_0^{\ell} \frac{1}{2} C v^2 \, dx \tag{101}$$

$$\text{where} \quad V = V_0 \cos \frac{\pi x}{2 l} \quad . \tag{102}$$

$$W_p = \frac{1}{2} C V_0^2 \int_0^l \cos^2 \frac{\pi x}{2 l}\, dx \tag{103}$$

$$\text{whence} \quad W_p = \frac{C l V_0^2}{4} \quad . \tag{104}$$

The kinetic energy is a maximum, and the potential energy is zero, when the voltage is zero along the line and the standing wave of current is at a maximum. Then

$$W_k = \int_0^l \frac{1}{2} L I^2 dx \tag{105}$$

$$I = I_0 \sin \frac{\pi x}{2 l} \tag{106}$$

$$\therefore W_k = \int_0^l \frac{1}{2} L I_0^2 \sin^2 \frac{\pi x}{2 l}\, dx \tag{107}$$

$$W_k = \frac{L l I_0^2}{4} \quad . \tag{109}$$

assuming zero resistive losses,

$$\frac{C l V_0^2}{4} = \frac{L l I_0^2}{4} \quad . \tag{110}$$

$$\therefore I_0 = \frac{V_0}{\sqrt{\frac{C}{L}}} = \frac{V_0}{Z_0} \tag{111}$$

$$\text{whence} \quad Z_0 = \sqrt{\frac{C}{L}} \quad . \tag{112}$$

$$I_{eff} = \frac{I_o}{\sqrt{2}} = \frac{V_o}{Z_o\sqrt{2}} . \tag{113}$$

The volt amperes of the standing wave system is

$$VA_{SW} = I_{eff}\ V_{eff} = \frac{V_o}{Z_o\sqrt{2}} \cdot \frac{V_o}{\sqrt{2}} \tag{114}$$

$$\therefore\ VA_{SW} = \frac{V_o^2}{2Z_o} . \tag{115}$$

The volt ampere product of the lumped constant Hartley oscillator having an oscillation capacitance C is

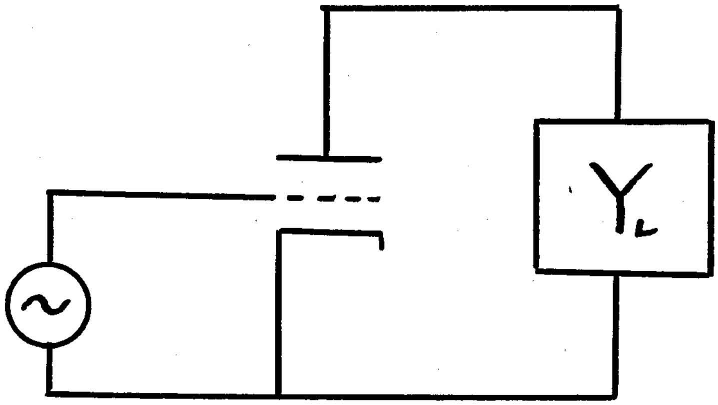
$$VA_{LC} = \pi f C V_o^2 . \tag{116}$$

Thus in the standing wave system, the oscillation energy is not a function of frequency and can be increased by decreasing Z_o, the characteristic impedance of the line. In the lumped constant circuit, on the other hand, it is necessary to increase C for a high volt ampere product. This, of course, will limit the maximum frequency obtainable.

There are other advantages gained by the use of concentric line tube-integrated circuits. Radiation losses, an important factor at frequencies involving very short wave lengths, are negligible in the concentric line system, since the standing waves are enclosed within the outer conductor. Furthermore, such circuits are particularly well suited for the grid-separation or grid return circuits. In such a circuit the output load is placed between the plate and the grid, as contrasted to the conventional arrangement where the output load is placed between the plate and the cathode. The cathode-return and grid-return circuits are shown in Fig. 7. It is readily seen that, if the cathode-grid resonator and the grid plate resonator are not coupled by external means, the only way in which energy can be exchanged between the resonators is through the electron beam and the electromagnetic coupling through the grid. Concentric line circuits lend themselves to easy tuning by means of moveable plungers in the cathode-grid resonator and the grid-plate resonator.

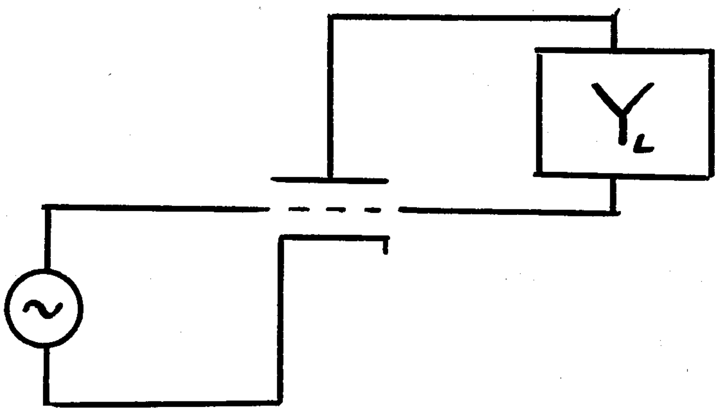
4. Other means of increasing the frequency range.

In addition to designing tubes and circuits specifically for use at ultra high frequencies it is possible to increase the upper frequency limit by adjusting the operating conditions. It was shown in Chapter III that the transit time varies inversely as the square root of the electrode voltages. It follows that by increasing these voltages the transit time can be decreased. There is a limit to how high these voltages may be increased imposed by the necessity for dissipating the resultant heat. The problem is emphasized when the tube electrodes are made smaller for high frequency use. Pulsed operation is one answer to the problem, since high voltages may be applied for short intervals without overheating the tube. Increased emission of electrons by the use of better emitting material in the cathodes would also help.



CATHODE RETURN

figure 7(a)



GRID RETURN

figure 7(b)

CHAPTER V

DISK SEAL TRIODES

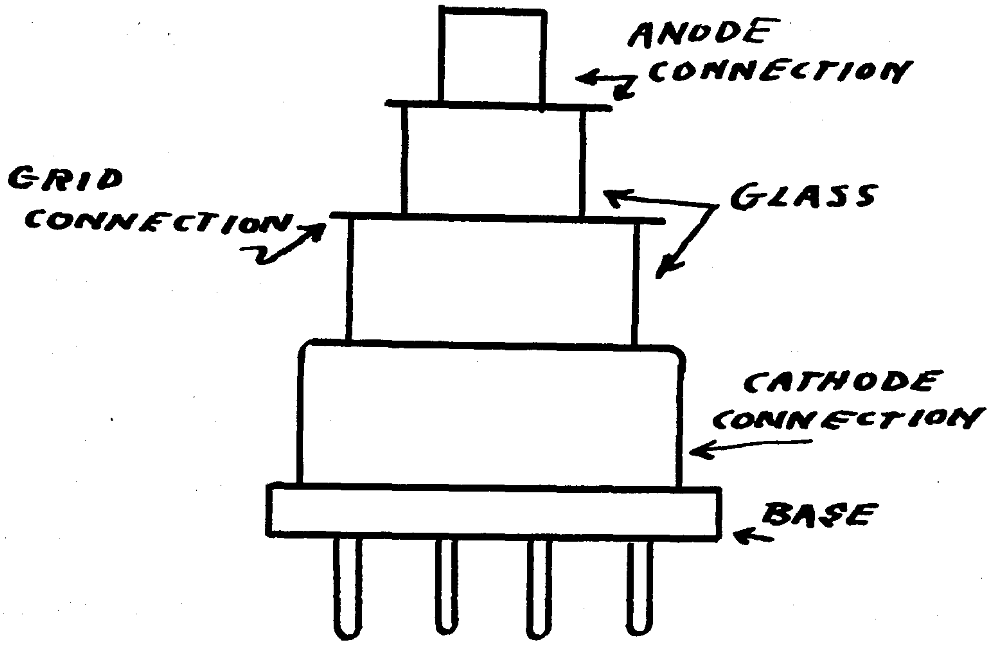
1. Introduction.

The considerations expressed in the foregoing chapters led to the development of a special tube for use at ultra high frequencies in concentric line circuits. It is known as the disk seal tube and was described in February 1945 by E. D. McArthur, (28). Electronically, the disk seal triode uses the same space-charge control principle as conventional triodes. Geometrically, it is a tube built from simple, smooth surfaced disks and cylinders into a structure which usually has circular symmetry. Philosophically, the tube is an embodiment of the principle that in the microwave region it is necessary to consider the amplifier or oscillator not as an electron tube with an attached circuit but rather as a single electrical system having one section walled off and evacuated to house the electronic activity.

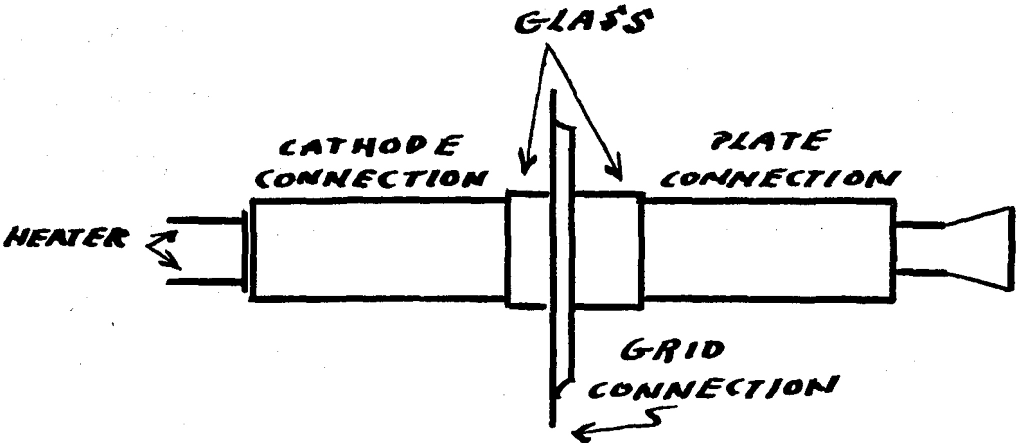
2. Planar electrode disk seal tube.

In Fig. 8(a) is shown the most common disk seal tube. It is known as the lighthouse tube because of its resemblence to that sort of structure. The anode connection is the cylinder and disk at the top. Below it and insulated by a cylindrical glass wall is the grid disk of larger diameter. Another glass section separates the grid disk from the cathode connection cylinder, which is of still larger diameter. The heater leads and the cathode bias leads are brought out through the base of the tube in pins.

This stepped construction makes the lighthouse tube well suited for use in concentric line circuits such as are illustrated in Fig. 7.



(a) LIGHT HOUSE TUBE



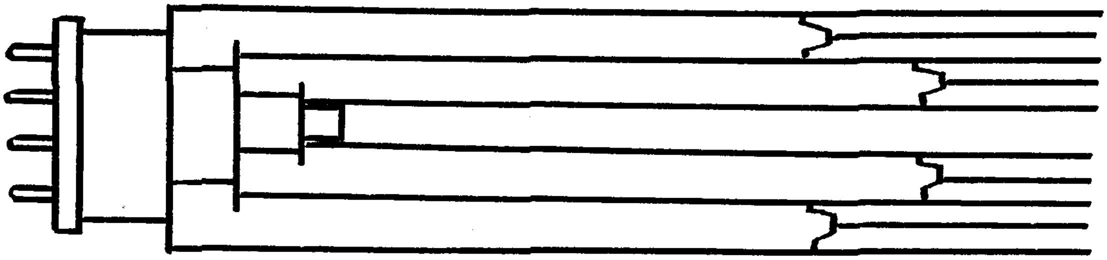
(b) PENCIL TUBE

figure 8

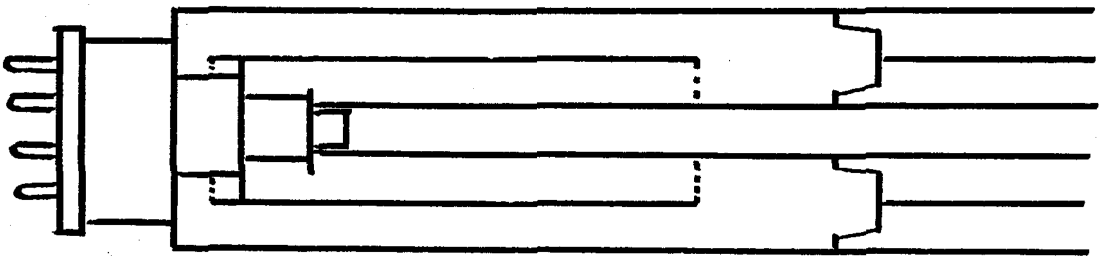
In Fig. 9(a) is shown a "folded back" grid separation circuit in which the cathode-grid cavity is folded back, or concentric with, the grid-anode cavity. Each cavity is tunable by means of a movable plunger. Radio frequency input and output are obtained by means of probes in the respective cavities. In an oscillator, feedback can be provided by probe coupling between the cavities or by some sort of cavity sharing arrangement as illustrated in Fig. 9(b). This is known as a re-entrant type circuit.

The early lighthouse tubes, such as the first 2C39s and 2C40s, appeared during World War II. They were designed for frequencies of the order of 400 megacycles. Improvements since then have extended their range to around 3000 megacycles. Some lighthouse tubes reverse the positions of the anode and cathode from that shown in Fig. 8(a) to improve the plate dissipation. Cooling fins are also common in both types.

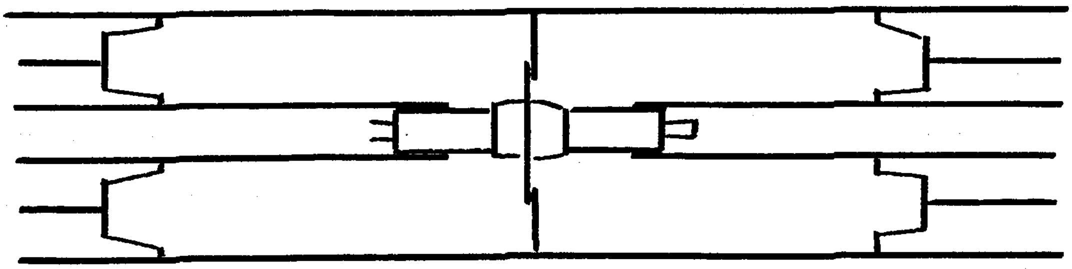
Early in 1949, J. A. Morton and R. M. Ryder described before a meeting of the New York section of the IRE a new tube of this family developed in the Bell Telephone Laboratory. It is known as the B.T.L. 1553. It is characterized by its extremely thin oxide layer (.5 mil) and extremely small cathode (from the top of this layer)-to-grid spacing (.6 mil). The grid is made of one layer of parallel wires .33 mil in diameter spaced 1000 per mils. In a triode the upper transconductance limit of approximately 11000 microhms per milliampere is reached when the cathode and grid are so close that the electron velocity due to effective grid potential is small compared to the average Maxwellian velocity of cathode emission. Ordinary microwave triodes are a factor of 20 to 25 below this limit. The 1553 has a



(a) FOLDED-BACK CIRCUIT



(b) RE-ENTRANT CIRCUIT



(c) END-TO-END CIRCUIT

figure 9

transconductance of 50,000 microhms at 25 milliamperes. It was designed to amplify at 4000 megacycles and can probably go higher, say 5000 megacycles. As a Class A amplifier it supplies a gain of 7 to 10 db with a bandwidth of 80 to 100 megacycles and power output of 1 watt at 4000 megacycles.

3. Concentric electrode tubes.

Concentric electrode tubes have many advantages over planar electrode tubes (E. G. Dorgelo, (10), electronic, mechanical, and thermal. For the same potential difference and spacing between electrodes, the grid to anode potential distribution increases more rapidly in the concentric electrode tube than in the planar. Therefore, because the electrons acquire their final velocities (a function only of the potential difference between plate and cathode) sooner, the transit time is less. In a planar tube, the current density in the grid-anode region is high. This creates a local decrease in the space potential (forming a virtual cathode if the potential drops to zero), which causes some of the electrons to return to the grid. The grid current increases at the expense of plate current. This effect is less evident in the concentric electrode tube. Here the diverging electron paths makes the current density in the grid-anode region less than in the grid-cathode region. Furthermore, all portions of the concentric electrodes are utilized uniformly over their surfaces. In planar construction, end effects introduce discontinuities.

Mechanically, the concentric construction is again better. It is very difficult to align and support planar electrodes so that they are truly parallel. If they are not absolutely parallel, the characteristics of one portion of the electrodes will vary from those at another portion. Planar tubes are therefore quite complex in internal structure. Concentric electrodes, on the other hand, are much simpler to align. Accurately turned mandrels in relatively simple jigs facilitate accurate assembly.

Thermally, the planar electrode tube is inferior. Electrode expansion under heat brings plate and cathode closer, causing the interelectrode capacitances to vary. In concentric tubes, expansion is such that the interelectrode capacitances remain relatively constant. The planar grid is subject to buckling when hot, causing non-parallelism. The concentric grid can be wound on several side rods that provide good thermal conduction and good mechanical support. In addition, it is difficult to heat efficiently a planar cathode surface.

For these reasons, a concentric electrode disk seal triode was developed at the RCA tube plant, in Harrison, New Jersey. The result is illustrated in Fig. 8(b). It is essentially a quarter-inch cylinder about 2 inches in length, for which reason it is known as the pencil tube. One end of the cylinder is the anode connection, the other is the cathode connection. The grid is supported by a grid disk about .8 inches in diameter insulated from the plate and cathode connections by glass cylinders. The grid projects from the disk into the anode, which extends from the anode connection cylinder. The cathode is connected to the cathode connection cylinder and supported by a kovar cylinder, which provides thermal insulation and electrical conductivity. It is projected through the grid disk so that its active portion is concentric with the grid and anode. The filament leads are brought through the cathode connection and kovar post to heat the active portion of the cathode. The pencil tube is small, light and compact. It is already in use in radio sonde applications where its light weight, small size and low heater power (.7 watts) are important factors. The tube is still undergoing tests, as is its suitability for various applications. Its dimensions make it especially adaptable to end-to-end grid separation circuits such as the one illustrated in Fig. 9(c). This type of plumbing is simpler then the folded back types. It is easier to locate input and output probes and the operating mechanism for the tuning plungers is less complicated.

4. Conclusion.

The space-charge control triode is still a very important component of electronic circuits. In recent years tubes that utilize transit time for their operation such as the klystron have been developed and offer many advantages for high frequency amplification. The conventional microwave triode, however, has a better bandwidth than the klystron, does not require as high voltages, and, very important, is much less expensive.

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