ELECTRON TUBES AND APPLICATION CIRCUITS

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CHAPTER 1 FUNDAMENTALS OF ELECTRON TUBES

1.1 Thermionic Emission

A metal can transmit electricity, because it contains electrons capable of moving freely, in it. In general, a substance consists of atoms, and an atom has several electrons. Electrons in a metal include those staying in the atoms they belong to, and those capable of moving freely to other atoms. The electrons capable of moving freely to serve conduction are called free-electrons or conduction-electrons. These free-electrons move with kinetic energy corresponding to temperature. However, in general, they cannot move out from the metal to surrounding space at room temperature. The reason is that attractive force like gravity acts on the surface of the metal, to keep electrons inside the substance, and that, for electrons to move out from the surface of a conductor, a work overcoming the attractive force of the surface must be done. The energy is obtained from the kinetic energy possessed by the electrons. In any known substance, the kinetic energy of electrons is not so high as to allow the emission of electrons from substances at room temperature, except the cases of photoelectric-effect and secondary-emission. However, as the temperature of a conductor rises, the kinetic energy of free-electrons increases, and at a sufficiently high temperature, a considerable quantity of electrons are emitted from the surface of the substance. This is the thermionic emission.

1.2 Triodes

A diode is a vacuum tube comprising two electrodes, viz, cathode to emit electrons by thermionic emission and plate (or anode) to surround it. In the diodes, if the plate is of positive potential, it attracts electrons and current flows, but if the plate is negative, it repels electrons, not to allow current flow.

Therefore, it is a rectifier.

A triode is made by adding a reticulate or grid-like third electrode to a diode, and the electrode is placed between the cathode and the plate, to control the electron current. This third electrode is called a control grid or generally a grid.

Examples of triode electrode structure are shown in Fig. 1.1.

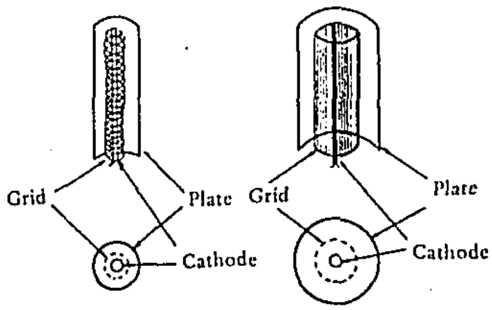


Fig. 1.1 Examples of Triode Electrode Structure

(1) Active of grid

The effect of a grid on plate current will now be considered. If voltages E_c and E_b are applied to the grid and plate respectively, charges Q as shown by the following formula are induced on the surface of the cathode.

$$Q = O_{fg}E_c + C_{fa}E_b = C_{fg}(E_c + \frac{C_{fa}}{C_{fg}} E_b) \;\cdots\cdots\; (1.1)$$

where C_{fg} and C_{fa} indicate electrostatic capacities between cathode and grid and between cathode and plate respectively. For simplification, the effect of space-charge is neglected. Formula (1.1) shows that the charges induced on cathode surface are equal to the charge induced when voltage $(E_c + \frac{C_{fa}}{C_{fg}} E_b)$ is applied to the virtual plate positioned at electrostatic capacity C_{fg}. From these relation, Fig. 1.2(a) can be taken as Fig. 1.2(b).

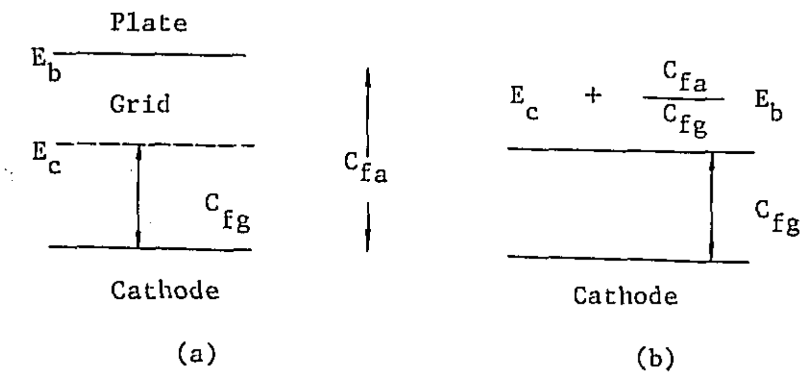


Fig. 1.2 Triode and Equivalent Diode

In general, the electric field in the vicinity of a conductor is determined by the distribution of surface-charge of the conductor, and the distribution of surface-charges is determined by all charges of the conductor. Therefore, the field strength in the vicinity of cathode is determined by Q of formula (1.1). On the other hand, since field strength determines the space-field and also the emission current, the plate current of a triode is controlled by the magnitude of Q. Therefore, plate current I_b is

$$I_b = K\left(E_c + \frac{C_{fa}}{C_{fg}} E_b\right)^{3/2} = K\left(E_c + \frac{1}{\mu} E_b\right)^{3/2} \quad .. \ (1.2)$$

In the above formula, K is a constant determined by the size of tube, and μ is given by

$$\mu = \frac{C_{fg}}{C_{fa}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \ (1.3)$$

This μ is a constant to show how many times the grid voltage affects emission current, compared to the plate voltage. This is called an amplification-factor or amplification-constant. Since the grid is closer to the cathode than the plate, C_{fg} if larger than C_{fa}, and therefore, μ is larger than 1. The μ is determined by the geometrical structure of tubes, but actually it varies to some extent, not to take a constant value, since the influence of space-charges is omitted as mentioned above.

(2) Characteristic curves of triodes

The space current of a triode, viz, all the current from the cathode, is affected by $(E_c + E_b/\mu)$, in the same way as the space-current of a diode is affected by the plate voltage, and in case the cathode voltage does not drop, it can be expressed by the following formula:

$$I_b = K\left(E_c + \frac{E_b}{\mu}\right)^{3/2} \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (1.4)$$

(Note) Diode plate current is determined by $I_b = KE_b{}^{3/2}$

In this case, it is assumed that the grid is negative and that all the space-current flows to the plate, the value in the parentheses being positive. If the value is negative, the plate current is 0.

The most important characteristics of triodes are relations of (a) constant grid voltage vs. plate voltage, and (b) constant plate voltage vs. plate current and grid voltage. These are shown in Figs. 1.3 to 1.5. Fig. 1.5 shows an example of saturation at higher plate current, due to low emission current.

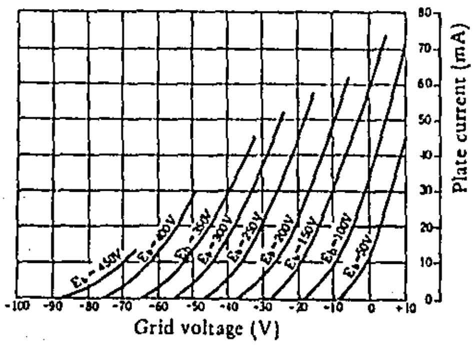


Fig. 1.3 E_G-I_p Characteristics (with E_b changed)

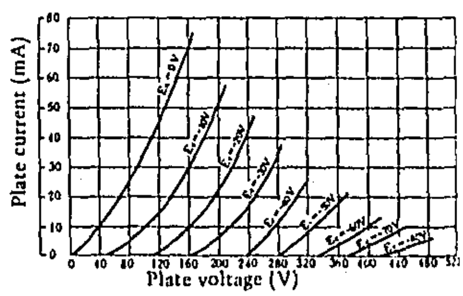


Fig. 1.4 E_p-I_p Characteristics (with E_c changed)

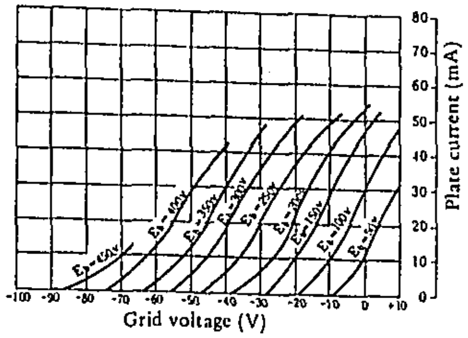


Fig. 1.5 $E_G I_p$ Characteristics with I_p saturated (with E_b Changed)

Features shown in the characteristic curves of triodes is that the plate current is determined by $(E_c + E_b/\mu)$ only. As for plate current, the state of 0 in the parentheses is established when the grid is of negative potential to neutralize the effect of the plate to the cathode. This state is known as cutoff.

$$\text{Cut off grid voltage (bias)} = -\frac{E_b}{\mu} \quad \cdots\cdots \quad (1.5)$$

(3) Grid current

When the grid voltage is negative, the emission current should exclusively flow to the plate, and not to the grid, but slight grid current flows. The main causes are initial current, ion current, leakage current, etc.

The initial current occurs because the thermions emitted from the cathode have an initial velocity. It can be suppressed to a negligible value if the grid voltage is set at -5V to -6V. Ion current occurs, when accelerated electrons collide with the residual gas in vacuum, to ionize it, and thus the generated cations flow into the (-) grid. Leakage current occurs, mainly because a metallic thin film, etc. adheres to the inside wall of tube.

The grid voltage is, even if positive, at lower potential than the plate voltage. In this state, the grid current occupies only a very small portion of all space-current. If the grid voltage becomes higher than the plate voltage, the grid current becomes larger due to the emission of secondary-electron from the plate.

Currently used triodes include miniature ones to those capable of giving several hundreds of kW output power. A tube with plate-loss up to about 1 kW are sealed in glass or ceramic, using oxcide-coated cathode or thoriated-tungsten filament. For larger tubes, tungsten filament and plate made of copper, forming part of container, and being cooled by water or evaporation cooling or forced aeration, are used.

1.3 Constant of Triodes

Characteristics of triodes can be expressed by a set of characteristic curves, but the operation in the vicinity of a specific operating point can be expressed by amplification-factor μ, plate-resistance r_p and mutual-conductance gm. By using these constants, the operation of a triode in ordinary state can quantitatively be calculated without using characteristic curves.

(1) Amplification factor

As shown in formula (1.3), amplification factor μ is determined by the geometrical arrangement of grid, plate and cathode. Amplification factor relates to the structure of grid, and the more perfectly the grid isolates the cathode from the plate, the larger the amplification factor. Therefore, the thicker the grid lines or the more narrow the grid line intervals, the larger the amplification factor. Also by increasing the distance between the grid and plate, the amplification factor becomes large. The amplification factors of ordinary triodes are between 3 to 100. Amplification factor is defined by the following formula.

$$\text{Amplification factor } \mu = -\frac{\Delta E_b}{\Delta E_c} \quad I_b \text{ constant} \quad \ldots \quad (1.6)$$

where E_c is grid voltage, E_b is plate voltage, and I_b is plate current. This formula has been obtained by differentiating formula (1.4) by E_c with I_b constant. In other words, formula (1.6) has been established on the consideration of how E_b should be changed to keep I_b constant which usually changes according to the change of E_c. Substantially, the situation is same as formula (1.3), but in the form of formula (1.6), μ is not always required to be constant. The way that μ changes according to the use of condition is shown in Fig. 1.6. Compared to other constants, it can be seen that μ changes the least.

(2) Plate resistance

This is also called dynamic internal resistance, and refers to the resistance shown by plate circuit, to a slight increase of plate voltage. If the plate voltage increment ΔE_b causes plate current increment ΔI_b, then the plate resistance r_p is expressed by the following formula.

$$\text{Plate resistance } r_p = \frac{\Delta E_b}{\Delta I_b} \quad E_c \text{ constant} \quad \ldots\ldots \quad (1.7)$$

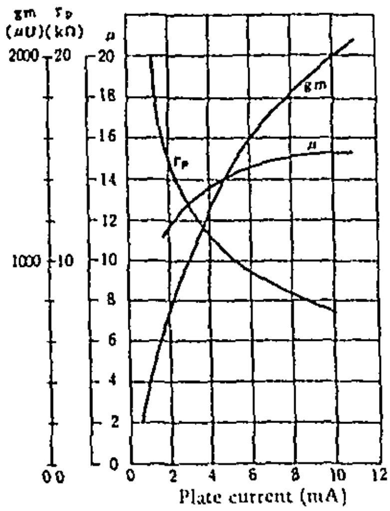


Fig. 1.6 Changes of μ, gm and r_p of triodes

The plate resistance is a reciprocal of the gradient of E_p-I_p characteristic in Fig. 1.4, and is determined by the grid and plate voltages of an operating point. The plate resistance is determined by the gradient of E_p-I_p characteristic, and at the largest gradient, the plate resistance becomes minimum.

The plate resistance of triode decreases with the increase of plate current. The extent of change is shown in Fig. 1.6. The plate resistance becomes small, if the area of cathode increases or if the distance between cathode and another electrode becomes small. With tubes different in grid structure only, the smaller the amplification factor μ, the smaller the plate resistance. The reason is that a constant plate voltage makes the intensity of electrostatic field generated in the vicinity of cathode inversely proportional to the amplification factor μ.

(3) Mutual conductance

The mutual conductance (or transconductance) gm is defined as a rate of plate-current change to grid-voltage change. If the grid-voltage changes by ΔE_c, and the plate current change in this case is ΔI_b, then the mutual-conductance gm is expressed by the following formula.

$$gm = \frac{\Delta I_b}{\Delta E_c} \quad \cdots\cdots\cdots\cdots\cdots \quad (1.8)$$

E_b constant

From formula (1.6) and (1.7),

$$gm = \frac{\Delta I_b}{\Delta E_c} = \frac{\mu}{r_p} \quad \cdots\cdots\cdots\cdots\cdots \quad (1.9)$$

This mutual conductance gm is an approximate standard in designing a tube. This is mainly determined by plate current, and becomes large with the increase of plate current, being proportional to its cube root. As obvious from Fig. 1.6, gm changes in inverse proportion to plate resistance r_p.

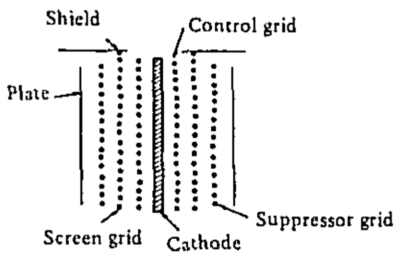
1.4 Pentodes

A pentode is a tube with a cathode, a plate and 3 grids, and the three grids are concentrically arranged between cathode and plate as shown in Fig. 1.7. The innermost grid is control grid, 2nd grid is screen-grid, and third grid is suppressor-grid.

In normal operation, the control-grid is negative to the cathode, the screen-grid is positive, and the suppressor-grid is at zero potential.

The second and third grids serve to change the relations of voltages and currents, to make tube operation suit various purposes, and electrostatically shield the first grid from the plate. This is important, in the case of use for a high-frequency amplifier.

Fig. 1.7 Arrangement of Pentode Electrodes



(1) Relations of voltage and current of pentodes

The relation of voltage and current of pentodes can be understood, if the potential distribution between plate and cathode is taken into consideration. Fig. 1.8 (a) shows the potential distribution in the case sufficient space-charges exist very close to the cathode. Under the limit of space-charge, the number of electrons drawn from the cathode is determined by the electrostatic field on the cathode surface, as in the case of triodes. The electrostatic field of pentodes is determined by the potentials of control-grid and screen-grid, and the structure and size of tube, but not being affected by the potential of plate. The reason is that the screen grid and the suppressor grid shield the cathode very efficiently from the electrostatic effect generated by the plate, and that, as a result, the number of electrons drawn out as space-charge is almost irrespective of the plate voltage.

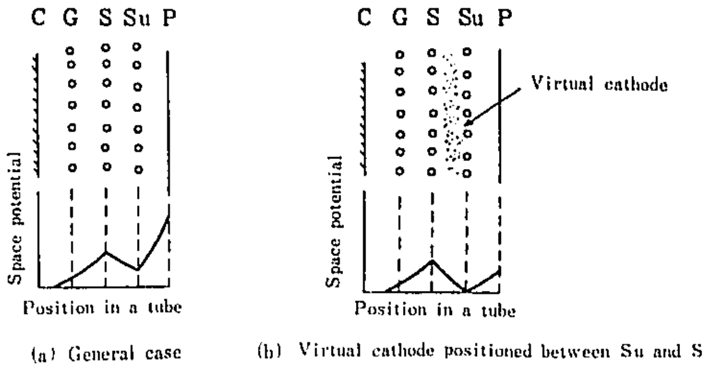


Fig. 1.8 Potential Distribution of Pentode

The electrostatic field on the cathode surface of pentodes is proportional to $(E_c + E_{c2}/\mu s)$, where E_c and E_{c2} are voltages of control-grid and screen-grid, and μs is a constant similar to amplification factor μ of triodes. The total space-current is the sum of plate current I_b and screen-grid current I_c , and takes the same form as formula (1.2).

$$\text{Total space-current} = I_b + I_{c2} = K_1 \left(E_c + \frac{E_{c2}}{\mu s}\right)^{3/2} \qquad (1.10)$$

where K_1 is a constant determined by the structure. The total current is zero, if $E_c = -E_{c2}/\mu s$, and μs is cut-off amplification factor.

The electrons of formula (1.10) pass through the control-grid, and are accelerated to high velocity by the screen-grid. The electrons emitted straight to the screen-grid only are caught by the screen-grid, and the other electrons run to the suppressor-grid. With the approach to the suppressor-grid, the velocity decreases. If the plate potential is high, the electrons passing through the suppressor-grid reach the plate.

If the plate voltage is very low, the space-charge in the vicinity of suppressor-grid take the form shown in Fig. 1.8 (b). If the space-charge is strong enough to return part of arriving electrons to the reverse direction from the suppressor-grid, they are called a virtual cathode. In this condition, the number of electrons arriving at the plate is a function of suppressor-grid and plate voltages, and the total space current becomes smaller than the value of formula (1.10).

(2) Characteristic curves of pentodes

Typical examples of characteristic curves of pentodes are shown in Figs. 1.9 and 1.10. The total space current is changed by screen grid and control grid voltages, in quite the same way as the plate current of triode changed by the control-grid and plate voltages. This can be understood also from formula (1.10).

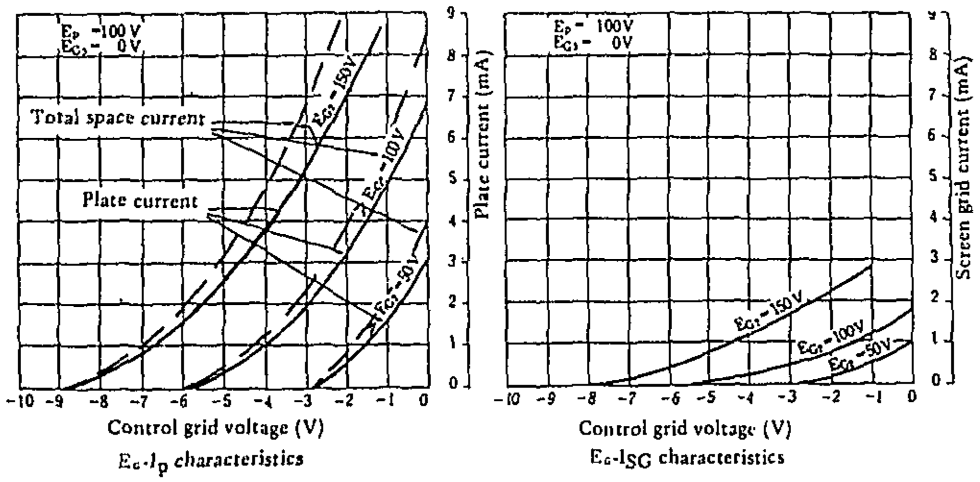


Fig. 1.9 Current Characteristics with E_G Changed in Pentodes

The distribution of total space current between plate and screen-grid can be regarded to be approximately irrespective of the plate voltage, unless the plate voltage is extremely low. Therefore, unless the plate voltage is low, the plate current and the screen-grid current are changed by the control-grid and screen-grid voltages, in the same way as the total space current.

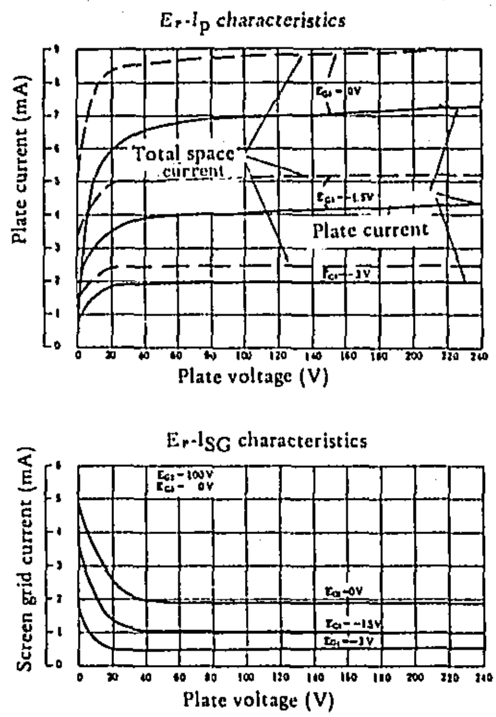
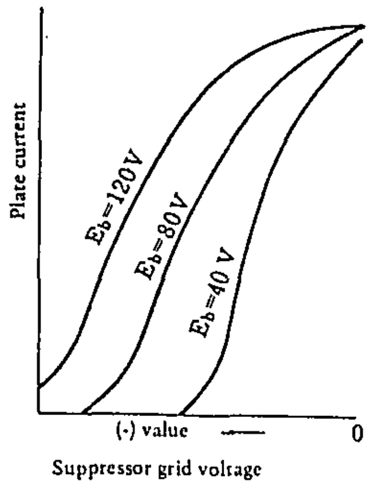


Fig. 1.10 Current Characteristics with E_p Changed in Pentodes

The plate potential to generate the virtual cathode is relating to the potential of suppressor-grid. If the suppressor grid is at a negative potential, the plate voltage must be raised, to avoid the influence of virtual cathode. Therefore, by changing the suppressor grid voltage, the plate current can be controlled as shown in Fig. 1.11. This action is used in a modulator.

1.5 Screen Grid Tubes (tetrodes)

A screen grid tube is a tetrode with suppressor grid removed from a pentode. Because the suppressor grid is removed, secondary electrodes flow between screen-grid and plate. If the potential of plate is lower than that of screen-grid, the primary electrons from the cathode produce secondary electrons on the surface of plate by impulse. The secondary electrons are attracted by the screen-grid rather than by the plate. Similarly, if the plate is higher than the screen-grid in potential, the secondary electrons generated in the plate are returned again to the plate, but the secondary electrons generated in the screen-grid, too, are attracted by the plate.



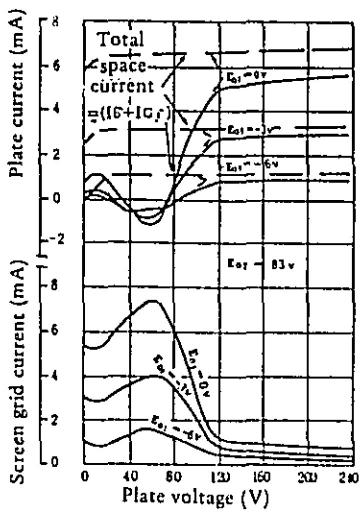


Fig. 1.11 $E_{SUG}-I_p$ Characteristics in tetrodes

Fig. 1.12 $E_p-I_p \cdot I_{SG}$ Characteristics in tetrodes

This phenomenon is remarkable particularly with the secondary electrons emitted to plate by screen grid. Because of this action, the relation of voltage and current of tetrodes are different from those of pentodes.

The secondary electron emission is usually large at the voltages of 25V to 75V, and it is not seldom that one primary electron produces one or two secondary electrons. However, if the surface of electrode is treated not to allow easy emission of secondary electrons, normally 0.1 to 0.2 electrons are produced.

The effect of secondary electron emission on a screen-grid tube is easy to understand, by watching how the plate and screen-grid currents change, according to the plate voltage, with the screen-grid and control-grid voltages kept constant. Fig. 1.12 shows these relations.

If the plate is sufficiently higher than the screen-grid, the plate current is almost irrespective of the plate voltage, as in the case of pentodes. If the plate is a little higher than the screen-grid in potential, the plate receives not only the primary electrons emitted from the cathode, but also the secondary electrons generated in the screen-grid. The number of these secondary electrons is relatively small. The number of primary electrons hitting the screen grid is only partial, and they go to the cathode side of the screen grid, not being affected by the plate directly.

If the plate is lower than the screen-grid, the secondary electrons generated on the plate surface are attracted by the screen-grid rather than the plate. Therefore, the plate current is expressed by the difference between the number of primary electrons and the number of secondary electrons. In this case, the plate current flows reversely, to be a negative value. If the plate voltage drops below this point, the plate current decreases with the voltage.

In the range the plate current decreases with the increase of plate voltage, negative resistance is shown, and the operation of screen-grid tube in this range is called dynatron.

Screen-grid tubes are popularly used as large output power amplifiers. Arrangement should be made not to have dynatron characteristics in the operation range.

1.6 Beam Power Tubes

If the space-charge density between screen-grid and plate is large in a screen-grid tetrode, a trough of potential is caused here, allowing to suppress the secondary electrons from the plate. That is, the same operation as in a pentode occurs. In the beam power tube, the suppressor-grid is not used, but the potential trough by the action of electron beams themselves is used.

To make the potential trough, the charge density must be large and the distance between screen-grid and plate must be sufficiently large. To obtain large charge density, in the case of beam power tube, the meshes of control grid are adjusted to those of screen grid, to avoid the dispersion of electrons, and electrons are gathered in beams by a beam forming plate. These relations are shown in Fig. 1.13. This structure serves to put in order the electrons arriving at the plate, and therefore the rises of plate characteristic curves are sudden, with remarkably sharp shoulders, as shown in Fig. 1.14. For this reason, undistorted output can be obtained.

Since the distance between screen-grid and plate must be large to a certain extent, to make a potential trough, beam power tubes are not adopted for large tubes, but tetrodes are used.

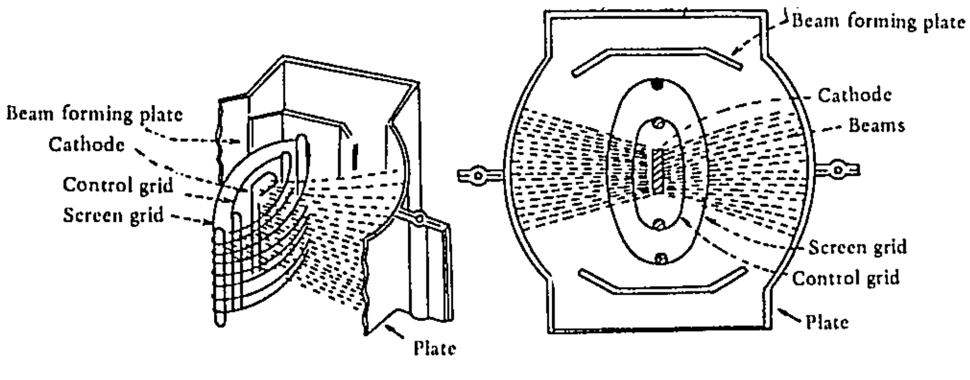


Fig. 1.13 Structure of Beam Power Tube

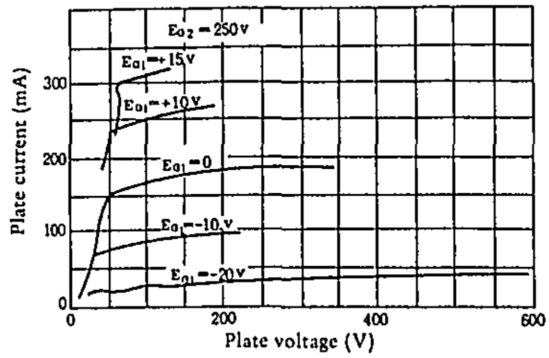


Fig. 1.14 E_p-I_p Characteristics of Beam Output Tubes

CHAPTER 2. VACUUM TUBE AMPLIFIERS WITH UNTUNED LOADS

2.1 Tube Amplifiers

If AC signal is superimposed, with negative potential applied to the control-grid of a tube, a current proportional to the AC signal flows in the plate circuit. If a load impedance is inserted in the plate circuit, an amplified AC voltage is generated across the load.

(1) Classification of amplifiers

Amplifiers can be classified into audio frequency amplifiers, to amplofy low frequencies, radio frequency amplifiers, video frequency amplifiers, etc, according to characteristics and purposes.

Amplifiers for wide frequency ranges are called wide-band amplifiers, and those for narrow frequency ranges are called narrow-band amplifiers. They are also classified into untuned amplifiers and tuned amplifiers.

Power amplifiers are classified into Class A amplifiers, Class B amplifiers and Class C amplifiers, according to the state of

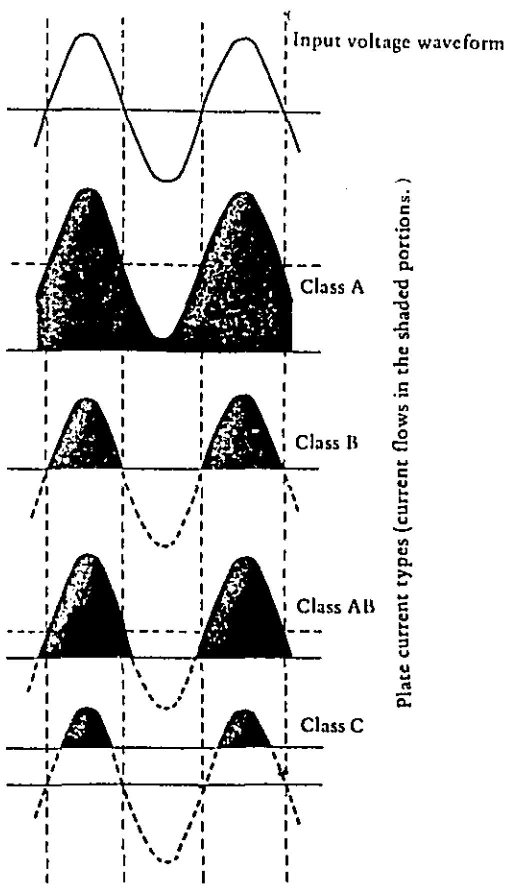


Fig. 2.1 Plate Current Waveforms for the Respective Operations

operation. This classification is based on the flowing types of plate current in one input cycle. Class A refers to the case where current flows throughout one cycle. Class B refers to the case where current flows throughout a half cycle only. Class AB refers to the case where current flows in the range less than one cycle and more than a half cycle. Class C refers to the case where current flows in the range less than a half cycle. Fig. 2.1 shows the plate current wave-forms for the respective operations.

If necessary, suffix 1 is attached in the case of no grid current flow, and suffix 2 is attached in the case of grid current flow, according to the respective operation states. For example, they are expressed like AB_1 or AB_2.

Class B amplifiers with tuned load are particularly called linear amplifiers, being used for amplification of modulated waves.

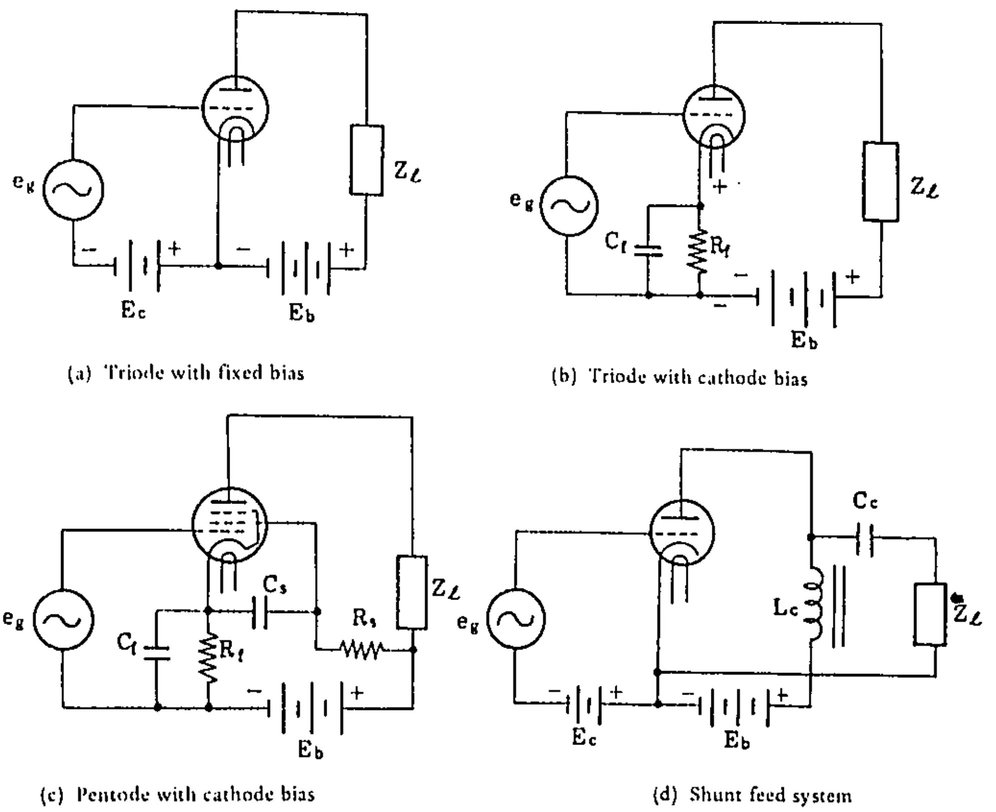
(2) Basic circuits

Fig. 2.2 shows basic circuits, Fig. (a) shows the case of a triode, in which the grid is kept at potential, negative to cathode by grid bias E_c. Signal e_g is superimposed on it. To the plate circuit, supply voltage E_b is applied through load Z_ℓ. In a general amplifier, bias is made by the plate current and the voltage drop of bias resistance R_f, as shown in (b). C_f is a capacitor for bypass, requiring a sufficiently large capacity to short R_f for the frequencies used.

Fig. (c) is the circuit of a pentode. 5 series resistance R_s is used to give proper positive potential to the screen grid, and bypass capacity C_s is inserted to make the AC potential of screen grid equal to that of cathode.

In these circuits, load Z_ℓ is directly applied to the plate, and therefore, DC component and AC component flow in superimposed state to Z_ℓ. In the case of (d), unlike the above circuits, DC component flows through L_c and AC component flows through Z_ℓ. The L_c must be small in DC resistance and large in AC impedance.

The C_c must be infinite in DC resistance and small in AC impedance.



(a) Triode with fixed bias

(b) Triode with cathode bias

(c) Pentode with cathode bias

(d) Shunt feed system

Fig. 2.2 Basic Circuits of Amplifiers

This system is called shunt-feed system, and the circuits (a) to (c) are called series-feed system.

(3) Equivalent circuits

Each of these circuits can be expressed by one equivalent circuit. The case of Fig. 2.2 (a) will be considered. When input e_g is given in this circuit, the AC component of plate current is assumed to be i_p. The AC component i_p causes the voltage drop of $-Z_\ell i_p$ across the load Z_ℓ. Therefore, in light of the tube as a whole, voltage e_g is applied to the grid and voltage $-Z_\ell i_p$ is applied to the plate. According to the definition of amplification factor μ, giving e_g to the grid is equivalent to giving $-\mu e_g$ to the plate. The total voltage of plate circuit is $-(\mu e_g + Z_\ell i_p)$.

If the input signal is not too large, the characteristic can be regarded to be linear, hence

$$i_p = -(\mu e_g + Z_\ell i_p)/r_p \quad \cdots\cdots\cdots\cdots \quad (2.1)$$

where r_p is plate resistance. Solving formula (2.1),

$$i_p = -\mu e_g/r_p + Z_\ell \quad \cdots\cdots\cdots\cdots \quad (2.2)$$

Voltage E_ℓ across the load Z is

$$E_\ell = i_p Z_\ell = -\mu e_g Z_\ell/r_p + Z_\ell \quad \cdots\cdots\cdots\cdots \quad (2.3)$$

Considering mutual conductance gm = μ/r_p, and substituting it for the above formula,

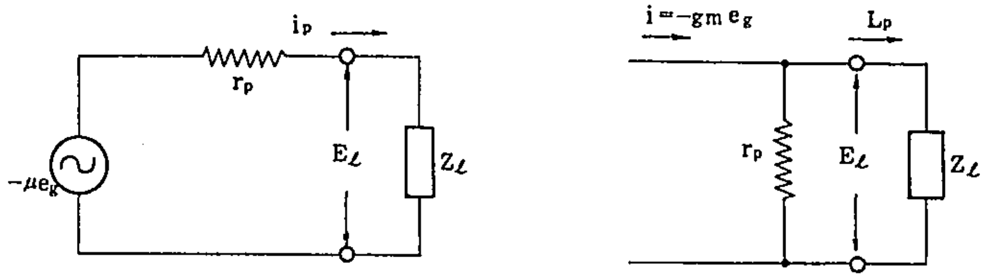
$$i_p = -gm\ e_g \frac{r_p}{r_p + Z_\ell} \quad \cdots\cdots\cdots\cdots \quad (2.4)$$

$$E_\ell = -gm\ e_g \frac{r_p Z_\ell}{r_p + Z_\ell} \quad \cdots\cdots\cdots\cdots \quad (2.5)$$

Using the relation of formula (2.3), the equivalent circuit of Fig. 2.3 (a) can be obtained. It can be replaced by a generator with electromotive force of $-\mu e_g$ and internal resistance of r_p. On the other hand, according to formula (2.5), the circuit is equivalent to the case of flowing current $-gme_g$ to the parallel coupled impedance of r_p and Z_ℓ. The former is called constant-voltage generator type, and the latter is called constant current generator type. Either can be used, but the former is mostly used when the plate resistance is almost same as the load resistance as in the case of a triode, while the latter is mostly used when the plate resistance is high as in the case of a multielectrode tube.

(4) Waveform distortion

The non-linearity in an amplifier is expressed by the ratio of the effective value of higher harmonics resultant voltage to the effective value of fundamental wave component contained in the output waveform delivered with sine wave input applied to the amplifier. This is called distortion factor. If E_1 is the effective



(a) Constant-voltage generator type (b) Constant-current generator type

Fig. 2.3 Equivalent Circuits of Amplifier

value of fundamental wave component, and E_2, E_3, ... are the effective values of higher harmonic components, then

$$\text{Distortion factor} = \frac{\sqrt{E_2{}^2 + E_3{}^2 + \cdots\cdots}}{E_1} \times 100 \; \ldots \quad (2.6)$$

Allowable distortion factor is smaller, if higher harmonics of higher order are contained. For example, the tone quality with 3% of 3rd harmonic corresponds to that with 6% of 2nd harmonic. When nonlinear distortion is caused at a point, it can be corrected, by adding a nonlinear impedance with curvature to erase it.

Frequency distortion is expressed by the frequency response given when frequency is changed with a signal applied to the input of amplifier. Phase distortion is expressed by the angle of phase shift, or by the delay time corresponding to it. If delay time is r, then the phase shift expressed by radian is ωr.

Usually, frequency distortion and phase distortion exist together in an amplifier. If there is frequency distortion, there is phase distortion, and if there is phase distortion, there is frequency distortion. The existence of the distortions allows the correction by an equalizer. Phase distortion comes into question with video frequency amplifiers, and it is not so important with audio frequency amplifiers. The reason is that human hearing sense is sensitive to the magnitude of frequency components, and not to the phase relation among component frequencies.

(5) Frequency components involved in nonlinear distortion

If an amplifier contains nonlinear distortion, new frequency components not contained in the input are generated at the output of the amplifier. The case in which nonlinear distortion is caused by the nonlinearity of tube characteristics will be described. If the characteristic is linear as shown in Fig. 2.4 (a), the input of f_a gives f_a only at the output. If the characteristic has a constant curvature as shown in Fig. (b), the output waveform is distorted with one crest collapsing, being vertically asymmetrical. In this case, 2nd harmonic $2f_a$, and DC component caused by rectification appear at the output, in addition to the fundamental wave f_a. If the curvature is not constant as shown in (c), the output waveform is distorted with both the crests collapsing. In this case, 2nd harmonic $2f_a$, DC component caused by rectification, and 3rd harmonic $3f_a$ appear at the output, in addition to the fundamental wave f_a.

In the above cases, the input voltage is always of single frequency. If the input includes two or more frequency components,

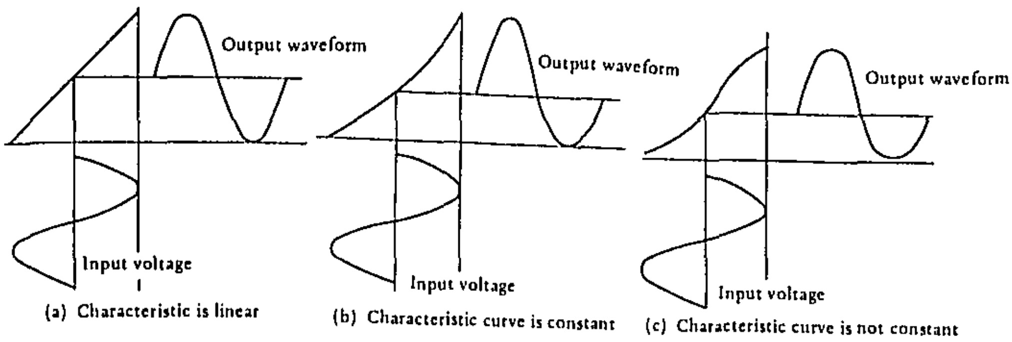


Fig. 2.4 Examples of Nonlinear Distortion

the situation is very complicated. In this case, modulation occurs among the respective frequencies, and many frequency combinations are caused. The case in which two frequencies of f_a and f_b are applied to the input will now be considered. In the case of (a), f_a and f_b only appear at the output, no intermodulation being caused.

In the case of (b), two fundamental wave components f_a and f_b, 2nd harmonic components shown by $2f_a$ and $2f_b$, and DC component involved in rectification appear at the output. Furthermore, in this case, two combinations of frequency components, viz. $(f_a + f_b)$ and $(f_a - f_b)$ by intermodulation appear.

In the case of (c), f_a, f_b, $2f_a$, $2f_b$, DC component, $(f_a + f_b)$, $(f_a - f_b)$, and furthermore, 3rd harmonic components $3f_a$ and $3f_b$ and complicated frequency combinations of $(f_a + 2f_b)$, $(f_a - 2f_b)$, $(2f_a + f_b)$ and $(2f_a - f_b)$ by intermodulation involved in 3rd harmonic appear as the output waveforms. Fundamental wave components f_a and f_b have special nature in this case, and their magnitudes have mutual relations with the magnitude of the other input voltage. The description in this paragraph is a cause of so-called cross modulation.

2.2 Tube Dynamic Characteristics

When an amplifier operates at large output, with characteristic surmised nonlinear, the operation can be obtained by drawing dynamic characteristic curves. Not only fundamental wave output but also higher harmonics outputs can be obtained.

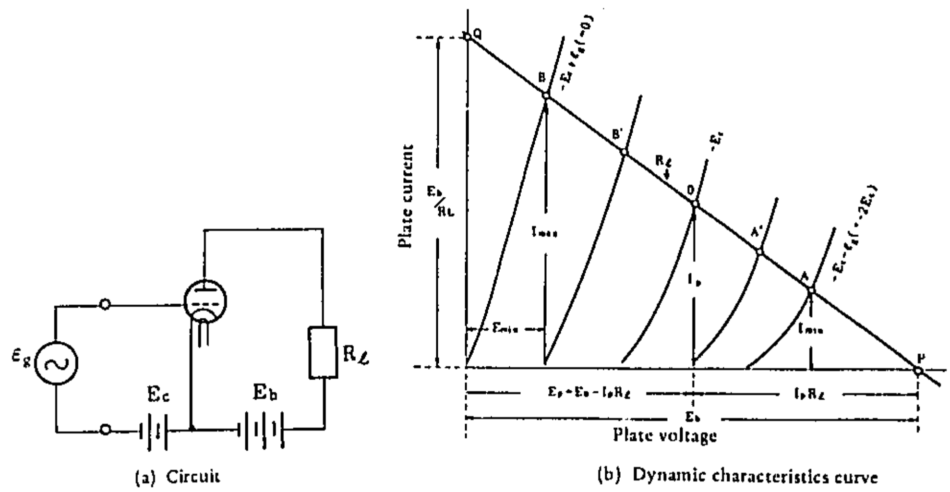


Fig. 2.5 Dynamic Characteristics of a Triode in Series Feed

(1) Triode

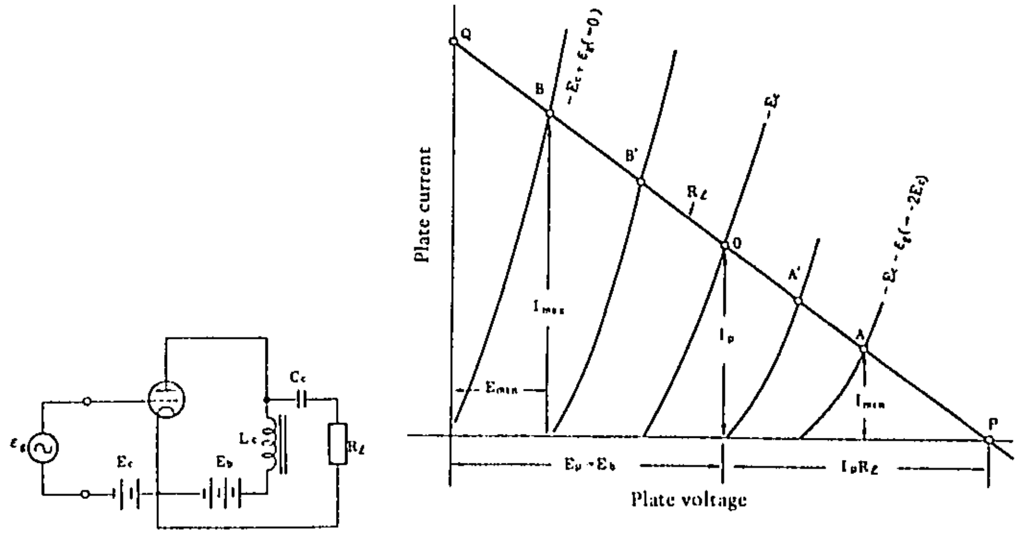
Fig. 2.5 (a) shows the circuit of a triode in series feed, and (b) shows its dynamic characteristics curve. To obtain dynamic curves, a load line is drawn on static characteristic curves. Since the plate voltage is E_b for zero current flowing through R_ℓ, point P is obtained. Then, if current flows through R_ℓ and voltage drop equal to E_b is caused across R_ℓ, then the plate voltage is zero and the plate current is equal to E_b/R_ℓ, point Q being obtained. Line P-Q is the load line. The gradient is equal to $1/R_\ell$. In the graph, A, A', O, B', and B show the operating points for respective grid voltages. If grid bias voltage is E_c, the reference point for operation with no input signal is point O. The relation among plate current I_p, grid voltage $-E_c$ and palte voltage E_p for point O is $E_p = E_b - I_p R_\ell$.

If input ε_g is applied to the grid, the operation range is from point A to point B. In Fig. 2.5, $E_c = \varepsilon_g$. Point B provides maximum plate current, minimum plate voltage and zero grid voltage. At this time, the amplitudes of the respective components in the output waveform are as shown in formula (2.7).

$$\left.\begin{aligned} &\text{Fundamental wave} = A_1 = \frac{I_{max} - I_{min}}{2} \\ &\text{2nd harmonic} = A_2 = \frac{I_{max} + I_{min} - 2I_p}{2} \\ &\text{Component by rectification} = A_0 = A_2 \end{aligned}\right\} \quad \cdots\cdots (2.7)$$

where I_{max} and I_{min} are respectively maximum and minimum plate currents.

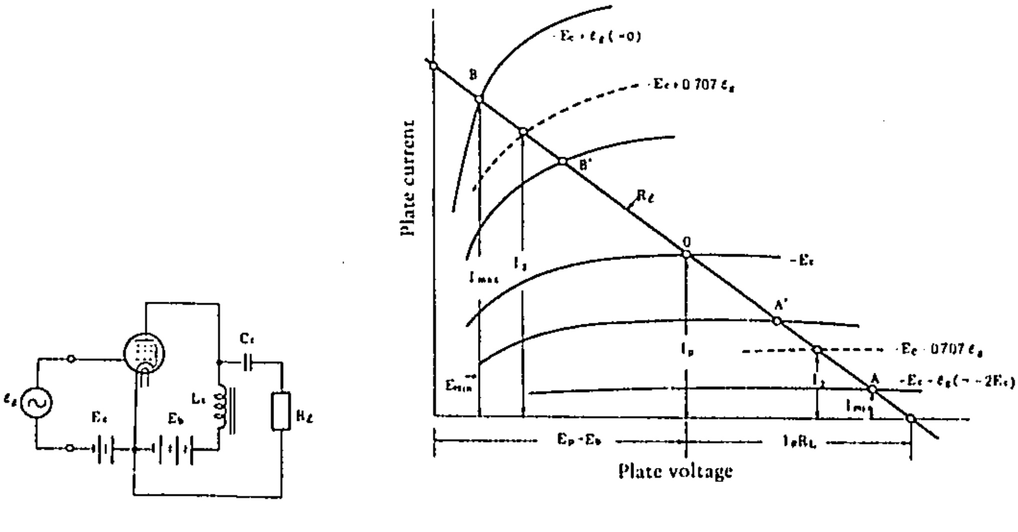
Also in the case of shunt feed as shown in Fig. 2.6, the dynamic characteristics can be considered in the same way as done in Fig. 2.5. If the DC resistance of choke coil L_c is zero, then $E_p = E_b$ for the reference point for operation. Point O is given on the bias voltage E_c curve. A line with $1/R_\ell$ gradient passing through point O, viz. P, O, Q, is the load line. The analysis for the rest is made in the same way as done in the previous paragraph.



(a) Circuit

(b) Dynamic characteristics curve

Fig. 2.6 Dynamic Characteristics of a Triode in Shunt Feed



(a) Circuit

(b) Dynamic characteristics curve

Fig. 2.7 Dynamic Characteristics of a Pentode in Shunt Feed

(2) Multielectrode tube

Also in the case of a multielectrode tube, dynamic characteristics curve are obtained in the same way. The case of pentode is shown for example in Fig. 2.7. In the case of multielectrode tube, the static characteristic curves are complicated in curvature, and the output contains considerable 3rd harmonic component. The amplitudes of the respective components are calculated by the following formula.

$$\text{Fundamental wave component} = A_1 = \frac{\sqrt{2}(I_2 - I_3) + I_{max} - I_{min}}{4}$$

$$\text{2nd harmonic component} = A_2 = \frac{I_{max} + I_{min} - 2I_p}{4}$$

$$\text{3rd harmonic component} = A_3 = \frac{I_{max} - I_{min} - 2A_1}{2} \quad \ldots\ldots \quad (2.8)$$

$$\text{4th harmonic component} = A_4 = \frac{2A_0 - I_2 - I_3 + 2I_p}{2}$$

$$\text{Component by rectification} = A_0 = \frac{\frac{1}{2}(I_{max} + I_{min}) + I_2 + I_3 - 3I_p}{4}$$

where I_{max} and I_{min} are maximum and minimum plate currents obtained with a predetermined input applied, and I_2 and I_3 are maximum and minimum plate currents for 0.707 times of predetermined input value.

(3) Load resistances and output waveforms

Fig. 2.8 shows the output waveforms affected by load resistances. Fig. (a) shows the case of triode, and (b) shows the case of pentode. In the case of triode, too high load resistance does not matter if attention is paid to nonlinearity only, but too low resistance causes a trouble. In the case of pentode, both too high and too low resistances cause large nonlinear distortion.

(4) Grid bias voltages and output waveforms

Fig. 2.9 shows how output waveform is affected by the magnitude of grid bias voltage, in the case of triode. By too high grid bias, the lower side of output waveform collapses at the negative peak of input signal. On the contrary, by too low grid bias, grid current flows at the positive peak of input signal, causing collapse. Even if the grid bias is too high, increasing the load resistance allows to improve nonlinear distortion.

2.3 Output Transformers for Power Amplifiers

As for the load of a power amplifier, a proper value for tube is selected, and the impedance is converted by an output transformer. Fig. 2.10 shows a basic circuit, and Fig. 2.11 shows an example of frequency characteristic.

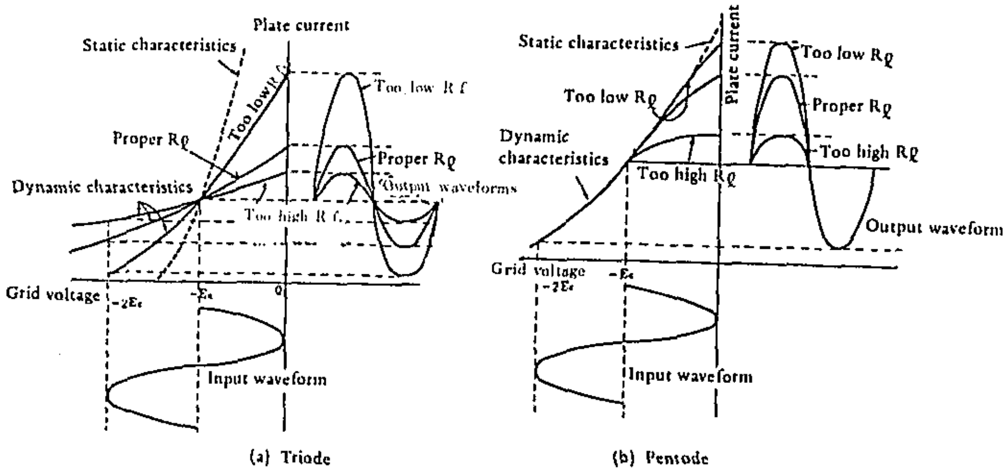
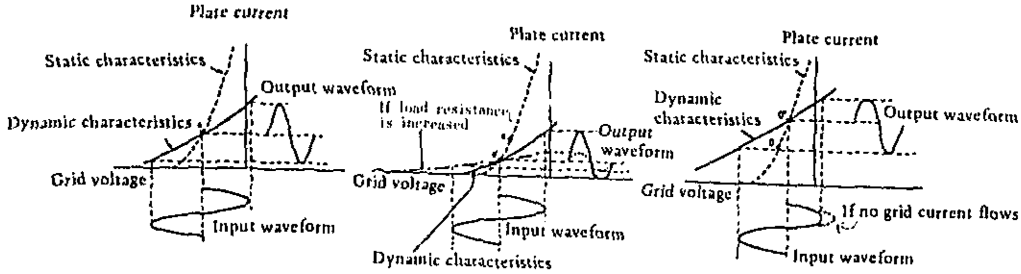


Fig. 2.8 Load Resistances and Output Waveforms



(a) Proper grid bias (b) Too high grid bias (c) Too low grid bias

Fig. 2.9 Grid Bias Voltages and Output Waveforms

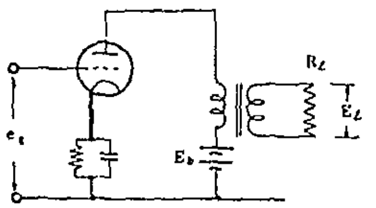


Fig. 2.10 Power Amplifier with Output Transformer Used

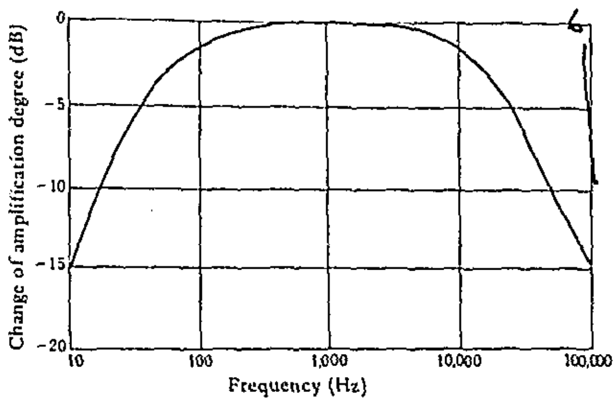


Fig. 2.11 Example of Frequency Characteristic

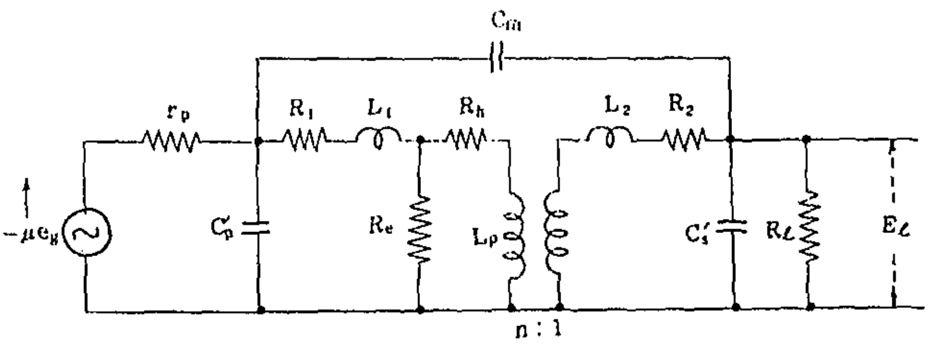
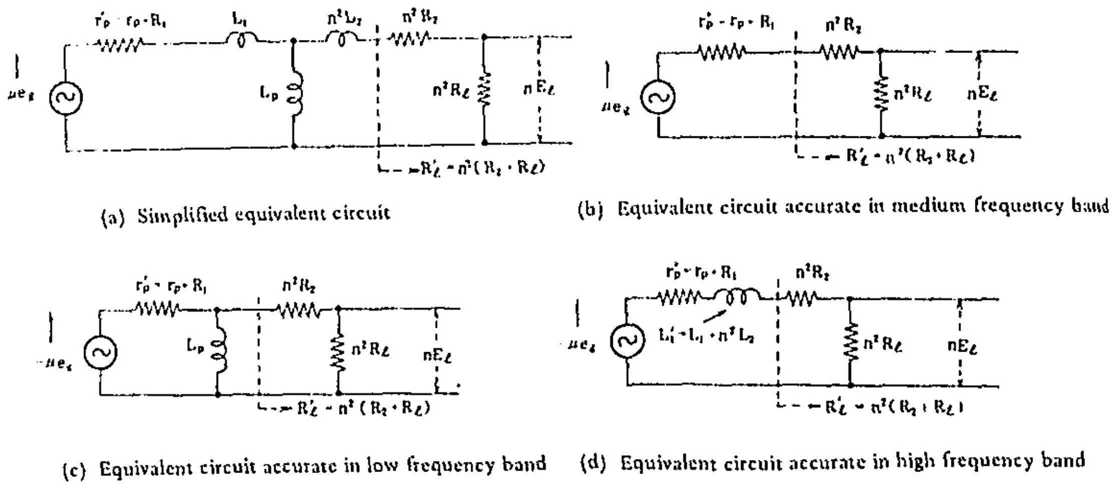


Fig. 2.12 Accurate Equivalent Circuit



(a) Simplified equivalent circuit

(b) Equivalent circuit accurate in medium frequency band

(c) Equivalent circuit accurate in low frequency band

(d) Equivalent circuit accurate in high frequency band

Fig. 2.13 Equivalent Circuits for Triode

(1) Analyses of frequency characteristic by equivalent circuits

(i) Triode

An accurate equivalent circuit is shown in Fig. 2.12. In the case of a triode, since plate resistance r_p and load resistance are relatively small, the influences of primary parallel capacity C_p', secondary parallel capacity C_s', interwinding capacity C_m and eddy current loss resistance R_e can be neglected. Hysteresis resistance loss R_h is small and can be omitted. Equivalent circuits in this case are as shown in Fig. 2.13. In this case, the frequency band is divided into 3 ranges, medium range, low range and high range.

(a) Medium range

In the medium range, L_p reactance is large compared to r_p', and is neglected. Therefore, the equivalent circuit is as shown in Fig. 2.13 (b), and the amplification degree A_m is

$$A_m = \frac{n^2 . R_\ell}{r_p' + R_\ell'} . \frac{\mu}{n} = \frac{\mu n R_\ell}{r_p' + R_\ell'} \quad \cdots\cdots\cdots\cdots\cdots \quad (2.9)$$

where R_ℓ' is an equivalent load resistance in terms of primary side, side, being a value with secondary winding resistance taken into consideration, viz. $R_\ell' = n^2\ (R_2 + R_\ell)$.

(b) Low range

In the low range, L_p reactance must be taken into account. Therefore, the equivalent circuit is as shown in Fig. 2.13 (c). Substituting for formula (2.9), the low range amplification degree A_ℓ is obtained.

$$\frac{A_\ell}{A_m} = \frac{1}{1 - j\frac{R}{X}} \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (2.10)$$

$$\text{or,}\ \frac{A_\ell}{A_m} = \frac{1}{\sqrt{1 + (\frac{R}{X})^2}}$$

$$\text{where } R = \frac{r_p'\ R_\ell'}{r_p' + R_\ell'} \qquad X = \omega L_p$$

(c) High range

In the high range, L_p reactance is large, and parallel effect can be neglected. But since the influence of leakage inductance cannot be neglected, the equivalent circuit is as shown in Fig. 2.13 (d). Similarly, substituting for formula (2.9),

$$\frac{A_h}{A_m} = \frac{1}{1 + j\frac{X'}{R'}} \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (2.11)$$

$$\text{or}\ \frac{A_h}{A_m}\ \frac{1}{\sqrt{1 + (\frac{X'}{R'})^2}}$$

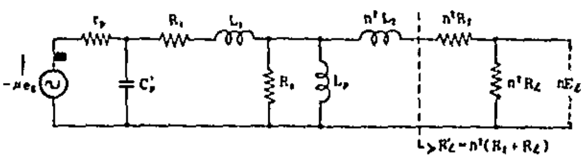
$$\text{where } R' = r_p' + R_\ell' \qquad X' = \omega L_I' = (L_1 + n^2 L_2)$$

L_1' is total leakage inductance in terms of primary side.

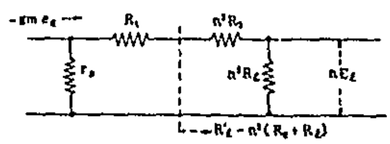
(ii) Multielectrode tube

In the case of multielectrode tube, plate resistance r_p is very large, and the influences of C_p' and R_e cannot be neglected. Therefore, the equivalent circuit is as shown in Fig. 2.14 (a). In the medium range, the reactance of leakage inductance is small, and its influence can be neglected. The reactances of C_p' and L_p are large, and their parallel effects can be neglected. At this time, the influence of R_e can also be neglected, compared to R_ℓ'. Thus, the equivalent circuit of (b) is obtained. If r_p is sufficiently large compared to other various resistances, the amplification degree A_m is,

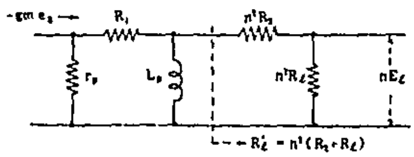
$$A_m = gm\, n^2\, R_\ell \quad \cdots\cdots\cdots\cdots \quad (2.12)$$



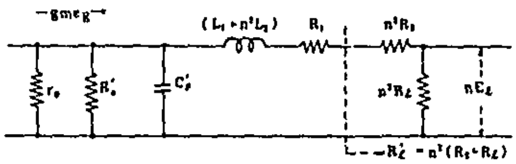
(a) Simplified equivalent circuit



(b) Equivalent circuit accurate in medium frequency band



(c) Equivalent circuit accurate in low frequency band



(d) Equivalent circuit accurate in high frequency band

Fig. 2.14 Equivalent Circuits for Multielectrode Tube

In the low range, since the parallel effect of L_p must be taken into account, the equivalent circuit is as shown in (c), and the amplification degree $A\ell$ is,

$$\frac{A_\ell}{A_m} = \frac{1}{1 - j\frac{R_\ell}{X_p}}$$

or .... (2.13)

$$\frac{A_\ell}{A_m} = \frac{1}{\sqrt{1 + (\frac{R'_\ell}{X_p})^2}}$$

where $R'_\ell = n^2(R_2 + R_\ell)$ $X_p = \omega L_p$

The frequency characteristics in this case are same as in the case of triode, and R'_ℓ can be considered instead of $r_p' + R'_\ell$ for triode.

In the high range, the influence of C_p' cannot be neglected, and furthermore, in relation to leakage inductance and the parallel resonance of C_p', R_e cannot be neglected. The equivalent circuit is as shown in (d). It is a parallel tuning circuit with C_p' and $n^2L_2 + L_1$ as tuning elements, and peaks are provided by high frequency.

(iii) Shunt feed type output transformer

If an output transformer is used in shunt feed, plate current does not flow at the primary side of the transformer, making primary inductance large, to improve low range characteristic.

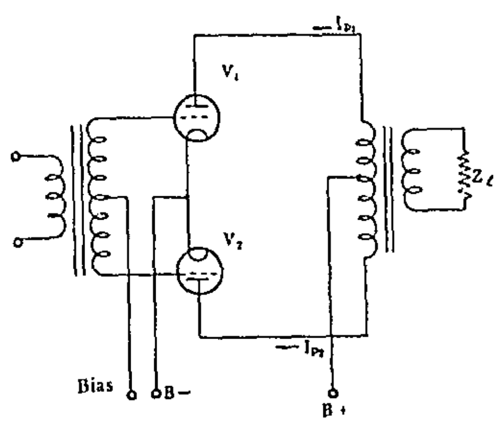


Fig. 2.15 Basic Circuit of Push-pull Amplifier

2.4 Push-pull Amplifiers

(1) Features of push-pull amplifiers

A push-pull amplifier is provided with two tubes V_1 and V_2 connected symmetrically, and inputs with same magnitude but different in phase are applied to the respective grids. The outputs are synthesized by a transformer with middle point. The tube operations can be used in any of Class A, Class B and Class AB.

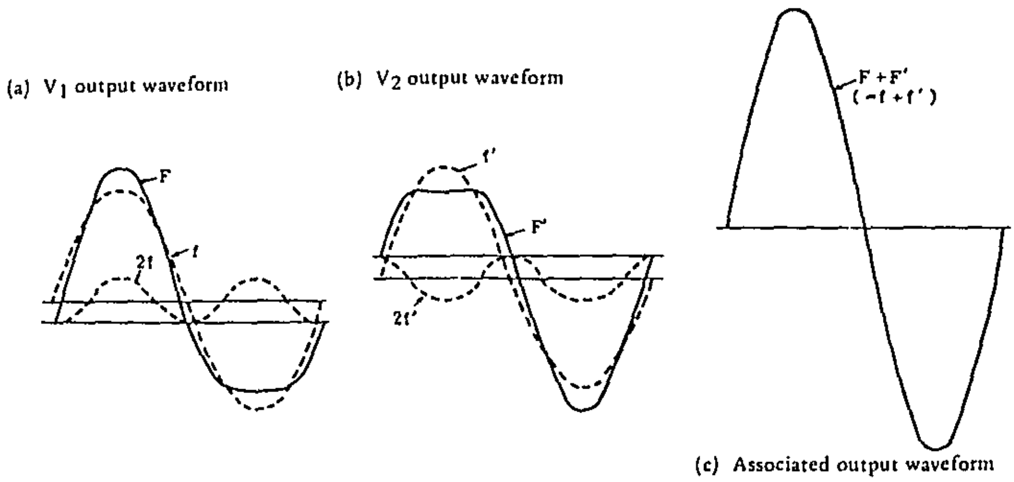


Fig. 2.16 Output Waveform Association of Push-pull Amplifier

When I_{p1} and I_{p2} flow at the primary side of the output transformer of a push-pull amplifier, the magnetization actions of both cancel each other, to nullify synthesized DC magnetization. The non-existence of DC magnetization serves to improve output transformer characteristics.

If there is distortion at the output for input voltage, higher harmonics of even-numbered orders in output current and intermodulation components generated involved in them erase each other, and do not appear at the output. Fundamental wave, higher harmonics of odd-numbered orders, and intermodulation components generated involved in them only appear at the output.

Fig. 2.16 shows the output waveform association of push-pull amplifier. In the drawing, F and F' show wave forms with the 2nd harmonics of input voltage contained in V_1 and V_2 output waveforms; f and f' show fundamental wave components; and 2_f and $2_{f'}$ show 2nd harmonics.

In the case of triode, since distortion is mainly caused by 2nd harmonics, the use of push-pull circuit allows to reduce the associated distortion considerably.

(2) Class B push-pull amplifier

The operation of Class B push-pull amplifier is as shown in Fig. 2.17, and the grid bias voltage is selected at the plate current cut-off, point P being the reference point for operation. If inputs are applied to the grids, the plate currents flow in positive half cycle, and are cut-off in negative half cycle. The operating point of positive maximum input is A, and at this time, the plate currents become maximum I_{max}, the plate voltages being

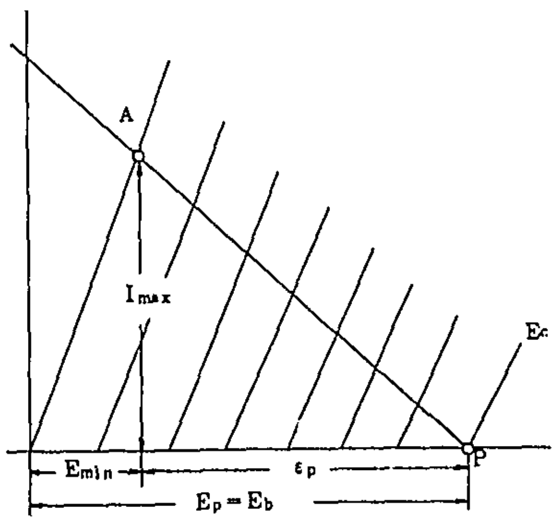


Fig. 2.17 Operation of Class B Push-pull Amplifier

minimum E_{min}. The line connecting point A with point P indicates the load resistance for the half cycle with plate currents flowing, being 1/4 of the load resistance R_ℓ between both the plates. If the inputs to the grids are negative, the load resistance in the half cycle is infinite. An averaged load resistance for one cycle is 1/2 R_ℓ.

In Class B push-pull amplifier, the plate current waveform of each tube is half-wave rectified and contains many higher harmonics of even-numbered orders which are, however, erased by the output transformer, not appearing at the output. Since the plate current flows for only half cycle of input, the plate efficiency is good, and since a positive input is applied to the grid mostly, large output can be obtained.

The plate current of each tube takes half-wave rectified waveform of sine wave with I_{max} as the peak value. The amplitude $\mathcal{I}_p$ of fundamental wave among it is given by

$$\mathcal{I}_p = \frac{I_{max}}{2}$$

If the load resistance between both the plates is R_ℓ, the primary voltage of output transformer is

$$\mathcal{I}_p R_\ell = \frac{I_{max}}{2} R_\ell$$

and the fundamental wave voltage ε_p between plate and cathode of each tube is

$$\varepsilon_p = \frac{\mathcal{I}_p R_\ell}{2} = \frac{I_{max}}{4} R_\ell$$

If the plate voltage is E_p, E_{min} is

$$E_{min} = E_p - \varepsilon_p = E_p - \frac{I_{max}}{4} R_\ell \quad \ldots\ldots\ldots\ldots\ldots \quad (2.14)$$

Hence, R_ℓ is

$$R_\ell = \frac{4\ E_p}{I_{max}}\left(1 - \frac{E_{min}}{E_p}\right) \quad \cdots\cdots\cdots\cdots\cdots \quad (2.15)$$

Output W_0 is

$$W_0 = \frac{(\mathcal{I}_p R_\ell)^2}{2 R_\ell} = \frac{I_{max}{}^2}{8} \quad R_\ell = \frac{I_{max}}{2}\left(1 - \frac{E_{min}}{E_p}\right) E_p \ldots (2.16)$$

The DC plate current I_p of each tube is

$$I_p = \frac{I_{max}}{\pi}$$

If there are two tubes,

$$2I_p = \frac{2I_{max}}{\pi}$$

Therefore, DC input is

$$E_p I_p = \frac{2I_{max}}{\pi} E_p$$

Efficiency η is obtained by dividend W_0.

$$\eta = \frac{W_0}{E_p I_p} = \frac{\pi}{4}\left(1 - \frac{E_{min}}{E_p}\right) \quad \cdots\cdots\cdots\cdots \quad (2.17)$$

In light of efficiency, E_{min} is required to be small. On the other hand, in light of output, I_{max} should be made large. The plate efficiency of Class B amplifier is maximum $\frac{\pi}{4} = 78.5\%$, and is actually from 50 to 65%.

2.5 Negative Feedback Amplifiers

(1) Basic characteristics

The circuit diagram of a negative feedback amplifier is shown in Fig. 2.18. Partially divided output E_3 of output voltage E_2 of amplifier A is fed back to the input, to be combined with input voltage E_0. The input voltage E_1 in this case is the composite voltage of E_0 and E_3 , and the phases are selected to keep E_1 smaller than E . The negative feedback allows to improve the stability of amplifier, nonlinear distortion, frequency distortion, phase distortion, internal noise, etc.

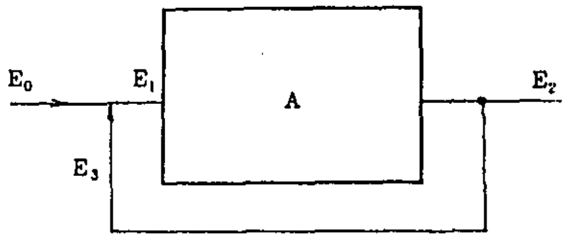


Fig. 2.18 Negative Feedback Circuit

There is the following relation among the respective voltages.

E = E + E

If amplification degree is E /E = A and amount of feedback is E /E = , then the effective amplification degree A' of negative feedback amplifier is

$$A' = \frac{E_2}{E_0} = \frac{E_2}{E_1 + E_3} = \frac{\frac{E_2}{E_1}}{1 + \frac{E_3}{E_1}} = \frac{\frac{E_2}{E_1}}{1 + \frac{E_2}{E_1} \cdot \frac{E_3}{E_2}} = \frac{A}{1 + A\beta} \quad ..(2.18)$$

where $A\beta$ is a ratio of E_3 to E_1, and is called feedback factor, and $(1 + A\beta)$ is the drop of effective amplification degree by feedback, being generally expressed as the amount of feedback in dB.

(2) Stabilization of amplification degree

If the feedback factor $A\beta$ is far larger than 1, then

$$A' = \frac{1}{\beta} \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (2.19)$$

That is, the effective amplification degree is determined by β, and is irrespective of the original amplification degree A. In general, since β is determined by circuit constants and is stable, the application of negative feedback makes the amplification degree very stable.

(3) Decrease of frequency distortion and phase distortion

The application of negative feedback decreases the nonlinear distortion of amplifier. It will now be assumed that higher harmonic electromotive force E_d is generated inside an amplifier, and that the amplification degree from that point to the output terminal is a. In the meantime, if the higher harmonic output voltage is D, the higher harmonic voltage fed back to the input terminal of the amplifier is βD. Therefore, D is

$$D = E_{da} - D\beta A = \frac{E_{da}}{1 + A\beta} \quad \cdots\cdots\cdots\cdots\cdots \quad (2.20)$$

In this case, if the higher harmonic output voltage with negative feedback removed is d, then, substituting $\beta = 0$ for the above formula,

$$d = E_{da}$$

Substituting it for the above formula of D,

$$D = \frac{d}{1 + A\beta} \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (2.21)$$

Thus, the nonlinear distortion generated inside the amplifier is decreased by negative feedback, and the degree of decrease is equal to the amount of feedback.

(4) Decrease of noise

The noise generated inside an amplifier is also decreased by negative feedback. The analysis in this case is same as in the case of nonlinear distortion. If the noise output with negative feedback is N and that without negative feedback is n, then,

$$\frac{N}{n} = \frac{1}{1 + A\beta} \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (2.22)$$

If the amplifier output E_2 is constant, the signal-to-noise ratio is improved by the amount of feedback.

(5) Oscillation in negative feedback circuit

Since an amplifier has many circuits containing reactance, phase distortion cannot be avoided. The degree is larger if the number of amplification stages increase. When phase shift is 180 degrees, the feedback increases the effective amplification degree. If the feedback voltage is equal to true input voltage, namely if $A\beta = 1$, the effective amplification degree becomes infinite, causing the state of oscillation.

The phase shift per one stage of resistance-coupled amplifier is ± 90 degrees with extremely high frequency and low frequency. If 2 stages of the amplifier are used, the maximum phase shift is ± 180 degrees, and if 3 stages, ± 270 degrees. To prevent oscillation, the overall phase shift must be 180 degrees or less, or AB for 180 degrees reached must be 1 or less.

To make the amount of feedback large in negative feedback amplifier, it is advisable to keep the amplification degree in the amplified frequency band as constant as possible and to cause attenuation as soon as possible in the ranges outside the frequency band. In this case, it is desirable that there is no phase shift involved in the attenuation, and here the relation between frequency characteristic and phase shift becomes important.

(6) Current feedback

Circuit examples of current feedback system are shown in Fig. 2.19. The amount of feedback is determined by the plate output voltage and R_s. Frequency characteristic is improved. The distortion caused by tubes is improved, but the distortion of output transformers cannot be improved.

With regard to noise, in the case of (a), (c), (d) and (e), since the noise component of output voltage is equal to the noise component of feedback voltage, the noise of output voltage is improved by negative feedback. In the case of (b), since the noise component flowing through R is different from the noise component flowing through R_S, the noise output is not improved by negative feedback.

(7) Voltage feedback

The voltage feedback includes the feedback from the primary side of output transformer, and the feedback from the secondary side. Fig. 2.20 shows circuit examples of voltage feedback system.

Frequency characteristic is not improved in the circuits of (a), (b), (c), (e) and (f) in which feedback is made from the primary side of output transformer. Distortion is improved in all the circuits, because of voltage feedback.

As for noise, in the circuits of (b), (d), (e) and (f), since the noise component of output voltage coincides with the noise output of feedback component, the noise output is improved by negative feedback. In (a) and (c) circuits, since there is no coincidence between them, it is not improved.

(8) Cathode follower

The circuit of Fig. 2.20 (e) is called cathode follower, since a load is connected to the cathode in the state of $\beta = 1$. If $\beta = 1$, the effective amplification degree A' of cathode follower circuit is

$$A' = \frac{A}{1 + A\beta} = \frac{A}{1 + A} \quad \dots\dots\dots\dots\dots\dots \quad (2.23)$$

A is an amplification degree for the case the load is connected to the plate as usual, and if A is sufficiently large compared to 1, then, from formula (2.23)

$$A \doteqdot 1 \qquad \cdots\cdots\cdots\cdots \qquad (2.24)$$

If $\beta = 1$, then input impedance $R_g{}'$ is

$$R_g{}' = R_g(1 + A) \qquad \cdots\cdots\cdots\cdots \qquad (2.25)$$

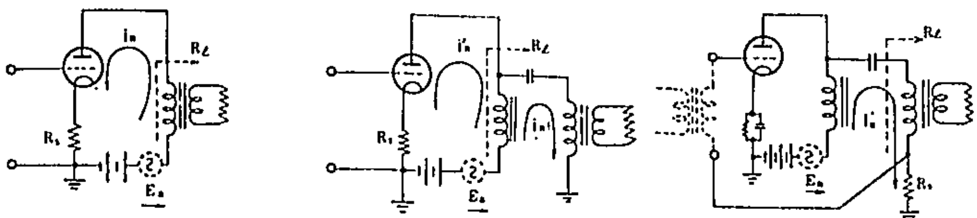
If $A' = 1$ and $\beta = 1$, then equivalent plate resistance $R_p{}'$ is

$$R_p{}' = \frac{r_p}{1 + \mu \cdot A' \cdot \beta} = \frac{r_p}{1 + \mu} = \frac{1}{\frac{1}{r_p} + gm} \qquad \cdots \qquad (2.26)$$

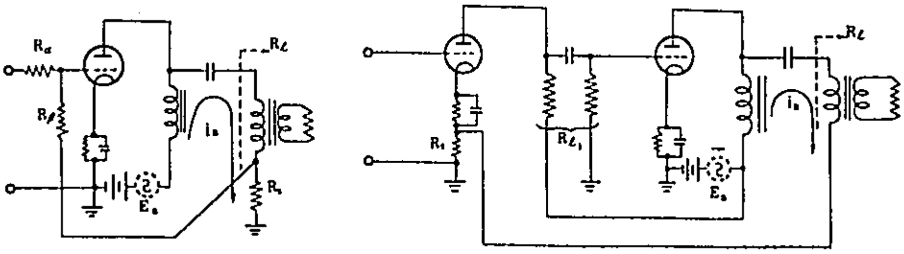
Generally $gm \gg \frac{1}{r_p}$, hence

$$R_p{}' = \frac{1}{gm}$$

A cathode follower is characterized by high input impedance and low output impedance. However, on the other hand, it has a fatal disadvantage that the required input voltage becomes conspicuously large, since the amplification degree becomes 1 or less.

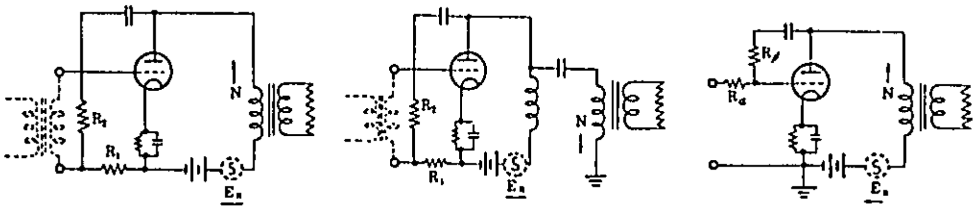


(a) Series type (b) Series type (c) Series type

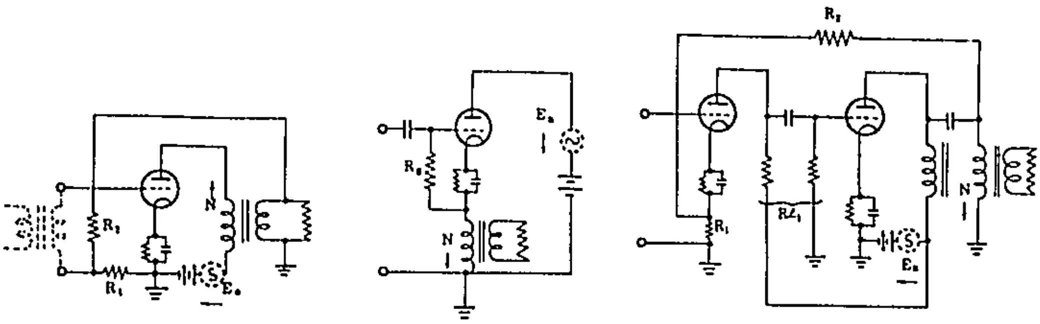


(d) Parallel type (e) Series type (2-stage amplification)

Fig. 2.19 Current Feedback Circuits



(a) Series type (b) Series type (c) Parallel type



(d) Series type (e) Series type (cathode follower) (f) Series type (2-stage amplification)

Fig. 2.20 Voltage Feedback Circuits

CHAPTER 3. VACUUM TUBE AMPLIFIERS WITH TUNED LOADS

3.1 Impedance of Single-Tuned Circuit

The parallel impedance Z of single-tuned circuit shown in Fig. 3.1 is

$$Z = \frac{Z_c \cdot Z_L}{Z_L + Z_c} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (3.1)$$

where $Z_L = R_L + j\omega L$ = Impedance of inductive shunt

$Z_c = R_c - j\frac{1}{\omega c}$ = Impedance of capacitive shunt

$f_0 = \frac{1}{2\pi\ \sqrt{L_c}}$ = Resonance frequency

$Z_s = Z_c + Z_L$ = Series impedance of circuit

$R_s = R_c + R_L$ = Total series resistance of circuit

$\omega = 2\pi f$ = 2π frequency

$\omega_0 = 2\pi f_0$ = Resonance frequency

$Q = \frac{\omega_L}{R_s}$ = Q of circuit

Q_0 = Q at resonance

Inductive shunt current

$$= \frac{e}{Z_L} = \frac{e}{R_L + j\omega L}$$

Capacitive shunt current

$$= \frac{e}{Z_c} = \frac{e}{R_c - j\frac{1}{\omega c}}$$

If the resistance components of Z_L and Z_c are omitted, then

$$Z_L = j\omega L, \quad Z_c = -j\frac{1}{\omega c}$$

Hence

$$Z_L \cdot Z_c = \omega L \cdot \frac{1}{\omega c} = \frac{L}{C}$$

At the time of resonance,

$$\omega_0 L = \frac{1}{\omega_0 c}$$

hence

$$Z_L \cdot Z_c = \frac{\omega_0 L}{\omega_0 c} = (\omega_0 L)^2$$

Substituting this for formula (3.1),

$$Z = \frac{(\omega_0 L)^2}{Z_s}$$

Since $Z_s = R_s$ at the time of resonance,

$$Z = \frac{(\omega_0 L)^2}{R_s} = (\omega_0 L)\ Q_0 = \left(\frac{1}{\omega_0 c}\right) Q_0 \quad \cdots\cdots\cdots \quad (3.2)$$

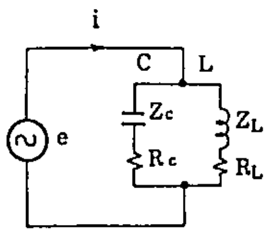


Fig. 3.1 Single tuned circuit

3.2 Voltage and Current of Double-Tuned Circuit

A circuit tuned to the same frequency on the primary side and the secondary side as shown in Fig. 3.2 is called double-tuned circuit. The double-tuned circuit is coupled by mutual inductance M.

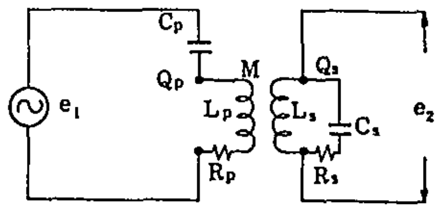


Fig. 3.2 Double tuned circuit

The coupling coefficient K between primary and secondary sides is expressed by $K = \frac{M}{\sqrt{L_p \; L_s}}$, and the current variation of primary and secondary circuits with K changed are as shown in Fig. 3.3. There are relations as described below.

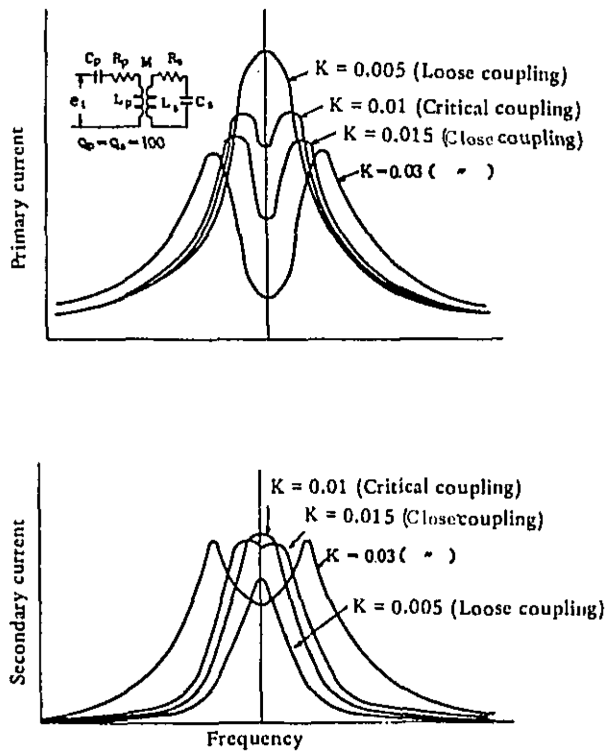


Fig. 3.3 Currents of double-tuned circuit

(1) If coupling coefficient is small:

If the coupling coefficient is small, the curve of primary current with frequency varied is almost similar to the series resonance curve in the case primary circuit is only considered. In this case, the secondary current is small, and varies fairly sharply with variation of frequency, compared to the curve of the secondary circuit only.

(2) If coupling coefficient increases to some extent:

In this case, compared to the case of (1), the primary current at the time of resonance decreases, and the primary current at slightly detuned frequencies increases on the contrary. Therefore, the curve of primary current becomes wide. At the same time, the curve of secondary current becomes high in crest, and wide.

This trend continues with the increase of coupling coefficient, until the resistance component coupled from secondary circuit to primary circuit is equal to the resistance of primary circuit at the resonance frequency, viz. until $R_p = \frac{(\omega M)^2}{R_s}$ is reached. This point is called the critical coupling K_c, and the resistance $\frac{(\omega M)^2}{R_s}$ coupled from the secondary circuit to the primary circuit is called the coupled-impedance.

(3) If critical coupling is established:

At critical coupling, the secondary current becomes maximum. The curve of secondary current becomes little wider than the resonance curve of secondary circuit only. On the contrary, the primary current becomes large at frequencies slightly deviated from the resonance frequency, and therefore the curve has two crests.

(4) If coupling is over the critical value (close coupling):

In this case, the double humped appearance of primary current becomes more distinct, and the clearance between the humps becomes wide. Though the hump of primary current becomes low, the crests of secondary current are almost constant, irrespective of the degree of coupling, as far as it is larger than the critical value.

In the state of critical coupling, maximum energy can be transmitted to the secondary circuit at the resonance frequency. In the circuit of Fig. 3.2, the impedance of primary circuit is the sum of the impedance of primary circuit itself and the impedance coupled to the primary circuit by the existence of secondary circuit, hence $Z_p + \frac{(\omega M)^2}{Z_s}$. If primary voltage is e_1, then primary current i_p is

$$i_p = \frac{e_1}{Z_p + \frac{(\omega M)^2}{Z_s}} \quad \cdots\cdots\cdots\cdots\cdots\cdots\cdots \quad (3.3)$$

Secondary induced voltage e_2 is

$$e_2 = -j\omega M i_p = \frac{\omega M e_1}{Z_p + \frac{(\omega M)^2}{Z_s}} = \frac{e_1}{\frac{Z_p}{\omega M} + \frac{\omega M}{Z_s}} \quad \cdots\cdots \quad (3.4)$$

where

$\frac{(\omega M)^2}{Z_s}$ = Coupled impedance affected to primary side by means of primary and secondary coupling

M = Mutual inductance

Z_p = $R_p + j \times P$ = Primary series impedance without secondary circuit coupled

Z_s = $R_s + j \times P$ = Secondary series impedance without primary circuit coupled

e_1 = Applied voltage

Among critical coupling coefficient K_c, coil Q and primary and secondary inductance L, there is the following relation.

If

$$K_c = \frac{M}{\sqrt{L_p \cdot L_s}} \quad \frac{L}{\sqrt{Q_p \cdot Q_s}}$$

$$L_p = L_s = L \qquad R_p = R_s = R \qquad Q_p = Q_s = Q$$

then

$$K_c = \frac{M}{L} = \frac{1}{Q}$$

$$M = K_c L = \frac{L}{Q} = \frac{L}{\frac{\omega L}{R}} = \frac{LR}{\omega L}$$

If Q_p and Q_s are 100 respectively, the critical coupling coefficient $K_c = 0.01$. Substituting M for formula (3.4), to obtain secondary induced voltage e_2,

$$e_2 = \frac{e_1}{\frac{Z_p}{\omega M} + \frac{\omega M}{Z_s}} = \frac{e_1}{\frac{Z_p}{\omega \cdot \frac{LR}{\omega L}} + \frac{\frac{LR}{\omega L} \cdot \omega}{Z_s}} = \frac{e_1}{\frac{Z_p}{R} + \frac{R}{Z_s}}$$

Since $Z_p = Z_s = R$ at the time of resonance, formula (3.5) is

$$e_2 = \frac{e_1}{\frac{Z_p}{R} + \frac{R}{Z_s}} = \frac{e_1}{1 + 1} = \frac{e_1}{2} \quad \ldots\ldots\ldots\ldots \quad (3.6)$$

That is, if the conditions of the primary and secondary circuits are the same, the output voltage e_2 becomes half of e_1. The load impedance at resonance, too, will become half of the impedance of single-tuned circuit.

3.3 Single Tuned Amplifier

A tuned amplifier is selective to frequencies, and amplifies only the frequencies in the vicinity of resonance. The simplest circuit and an equivalent circuit are shown in Fig. 3.4. The degree of amplification at resonance is as shown below, according to the equivalent circuit.

$$e_0 = I_p Z_L$$
$$e_g = -\frac{I_p}{gm} \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (3.7)$$

Therefore,

$$\frac{e_0}{e_g} = -\frac{I_p}{I_p} \cdot \frac{Z_L}{/\ gm} = -gm \cdot Z_L = A \quad \cdots\cdots\cdots \quad (3.8)$$

where A is amplification degree; gm is mutual conductance; and Z_L is the overall composite impedance of tuned circuit. Since formula (3.8) contains Z_L, the amplification degree varies according to the frequency. The variation is like a resonant curve of a tuning circuit, as shown in Fig. 3.5.

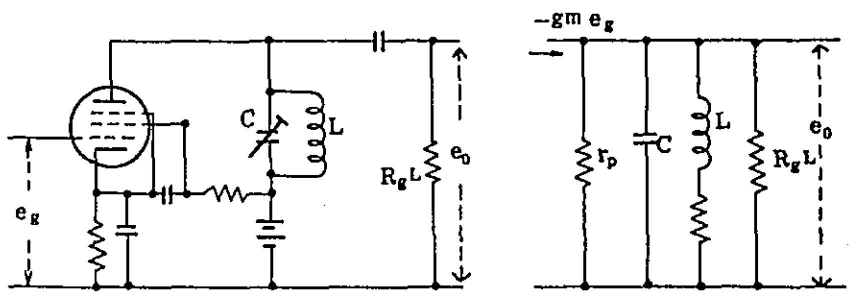
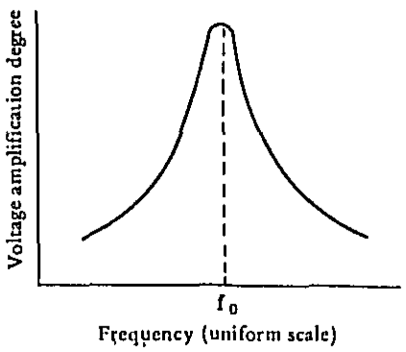


Fig. 3.4 Single-tuned circuit and equivalent circuit



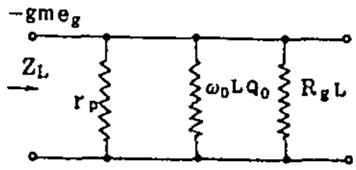


Fig. 3.5 A Variation of Amplification degree with frequency of single-tuned circuit

Fig. 3.6 Parallel impedance Z_L

As obvious from the figure, the maximum amplification degree occurs at the resonance frequency at Z_L. The resonance impedance of tuned circuit is parallel impedance consisting of $\omega_0 L\ Q_0$, resistances $R_g L$ and r_p. Therefore, this parallel impedance Z_L is

$$Z_L + \frac{1}{\frac{1}{r_p} + \frac{1}{\omega_0 L Q_0} + \frac{1}{R_{gL}}} = \frac{\omega_0 L Q_0}{1 + \frac{\omega_0 L Q_0}{r_p} + \frac{\omega_0 L Q_0}{R_{gL}}} \quad \ldots (3.9)$$

In an ordinary circuit, as the grid-leak resistance R_{gL} and plate resistance r_p is sufficiently large, compared to the parallel impedance of the resonance circuit, the amplification degree A_0 at resonance can be expressed by

$$A_0 = gm\ \omega_0 L Q_0 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3.10)$$

The bandwidth of tuned circuit refers to the width at the point the amplification degree becomes -3dB, and in the case of one stage single tuning, the bandwidth Δ is single tuning, the bandwidth Δ is

$$\Delta = \frac{f_0}{Q_{eff}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (3.11)$$

where f_0 is resonance frequency (kHz), and Q_{eff} is effective Q of amplifier circuit. The selectivity of single-tuned circuit is expressed by

$$S(dB) = 10 \log_{10} [1 + 4(\frac{\Delta f}{f})^2 \ Q^2] \quad \ldots\ldots \quad (3.12)$$

where f is resonance frequency; Δ f is detuned frequency; and Q is the effective Q of the circuit.

3.4 Double Tuned Amplifier

An amplifier with tuning circuits connected to the primary and secondary side as shown in Fig. 3.7 is called double-tuned amplifier. The tuning circuits are tuned to the same frequency of the primary and secondary, and are adjusted to have band-pass characteristics. Fig. 3.8 shows curves to indicate how the ratio K/K_c of critical coefficient K_c to actual coupling coefficient K affects the amplification degree. Calculation was made with the same Q for both primary and secondary.

In the double-tuned circuit, all the necessary side-bands contained in signal waves are amplified by a constant degree, and for other frequencies, sudden attenuation is caused. Therefore, it is very convenient for amplification of modulated waves.

In this amplifier, reactances become zero at resonance frequency for both primary and secondary circuits, and resistance component only remains. The coupled impedance from secondary circuit to primary circuit is $\frac{(\omega M)^2}{R_s}$, and it lowers the effective Q of primary circuit.

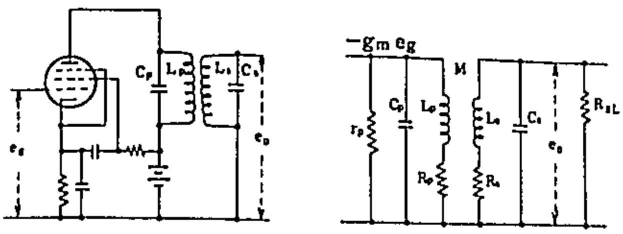


Fig. 3.7 Double tuned amplifier and equivalent circuit

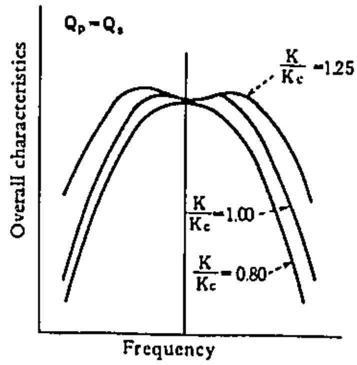


Fig. 3.8 Change of amplification degree with the coupling changed in double tuned circuit

The amplification degree is $K/K_c = 1$, viz. becomes maximum at critical coupling $K_c = \frac{1}{\sqrt{Q_p Q_s}}$. The maximum possible amplification degree A in this state is

$$A = gm \frac{\omega_0 \sqrt{L_p L_s} \sqrt{Q_p Q_s}}{2} \quad \dots\dots\dots\dots \quad (3.13)$$

where Q_p is $\omega L_p / R_p$ with plate resistance r_p taken into consideration in the primary circuit, and Q_s is $\omega L_s / R_s$ with grid resistance R_{gL} taken into consideration in the secondary circuit. To compare formula (3.13) with formula (3.10) with formula (3.10), the amplification degree of double-tuned circuit drops to $1/2$ in voltage ratio, compared to that of primary circuit type single-tuned circuit.

The selectivity of critical coupling state in double-tuned circuit is given by the following formula.

$$S(dB) = 10 \log_{10} [1+4 (\frac{\Delta f}{f})^4 Q^4] \quad \dots\dots\dots \quad (3.14)$$

where f is resonance frequency; f is detuned frequency; and Q is averaged Q of respective circuits, being $Q = \sqrt{Q_p Q_s}$.

3.5 Phase and Delay Time of Tuned Amplifier

When current progresses in a tuning circuit, the phase-angle depends upon the frequency, and if it is not in a linear proportional relation with frequency, distortion will occur. The distortion increases with the number of stages. If modulation frequency is low, phase distortion is generally small, being trivial, but if frequency range is wide like television signals, sufficient care should be taken.

An important characteristic in the case of a tuned amplifier is the phase-shift received by the side-bands of modulated waves compared to the phase-shift received by carrier-frequency. This relative phase-shift remains even after demodulation, and appears as time-delay distortion in output waves.

In the case of amplitude-modulated waves, if the upper and lower side-bands receive equal phase-shift of 45 degrees in the direction reverse to the carrier, it causes similarly phase diviation of 45 degrees in the envelope. The phase shift thus occurring in a tuned-amplifier will be phase-deviation in radio frequency, and it provides a 45-degree phase-shift also in the modulated-wave output after detection. The envelope delay in amplitude-modulated waves is shown in Fig. 3.9.

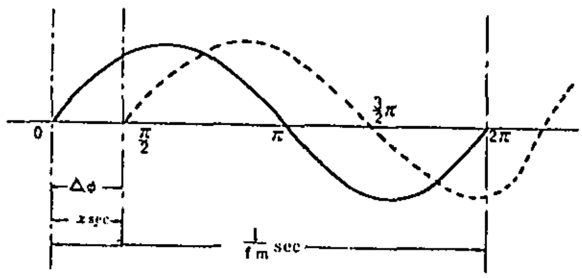


Fig. 3.9 Envelope delay

If $m = 2\pi f_m$ and delay is x seconds, then

$$x \ : \ \frac{1}{f_m} = \Delta\phi \ : \ 2\pi$$

Hence

$$x = \frac{\Delta\phi \cdot \frac{1}{f_m}}{2\pi} = \frac{\Delta\phi}{2\pi f_m} = \frac{\Delta\phi}{\omega m} \ \ldots\ldots\ldots\ldots \qquad (3.15)$$

where $\Delta\phi$ = Phase of side-bands compared to phase of carrier (radian), and $\omega m/2\pi$ = modulation frequency.

In the above formula, it is assumed that the upper and lower side-bands receive a phase-shift $\Delta\phi$ same in magnitude but reverse in sign, compared to that of the carrier, and that $\Delta\phi$ is positive when the side-bands of high-frequency delay in phase, and when side-bands of low-frequency advance in phase.

3.6 Self Oscillation of Multistage Tuned Amplifier

In a high-frequency amplifier, the electrostatic capacity between plate and grid may be the causes of positive feedback, and oscillation will occur. The capacities C_{pg} of tubes are

Large triodes	5 to 100 PF
Large tetrodes	0.05 to 1 PF

If the frequency is high, the reactance of C_{pg} becomes small, and will be the cause of oscillation.

In addition to C_{pg}, common impedance employed in the power-supply circuit for respective amplifiers may be the cause. Even if the amount of feedback is extremely small, the energy of output side is very large in comparison with the energy of 1st stage, and therefore slight coupling between input and output generates large energy. If amount of feedback becomes large, the amplifier oscillates and cannot be used. Even if the feedback is not so large as to cause oscillation, it generally changes the amplification degree, changes the frequency characteristic, greatly. Feedback must be reduced as far as possible.

The band-pass characteristic obtained by critical coupling of double-tuned circuit is also destroyed by slight feedback.

The elimination of regenerative action of tuned-amplifiers is particularly difficult when frequency used is very high. The reason is that capacitive reactance decreases with frequency, and that even small stray capacity causes considerable inverse current to flow. The inductive reactance commonly put in the respective sections of amplifier becomes large in proportion to frequency, being an important item at high-frequency.

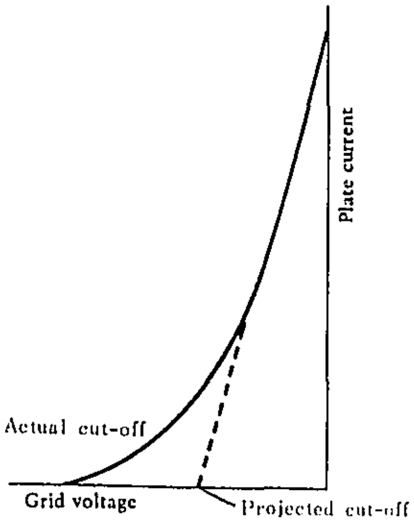
3.7 Linear Amplifier

A linear amplifier is Class B or Class C amplifier adjusted to have output power proportional to exciting power. To provide this characteristic, bias is selected at the point corresponding to the projected cutoff as shown in Fig. 3.10. This is Class B amplifier provided with tuning circuit, but since the tuning circuit, but since the tuning circuit removes higher harmonics, a push-pull amplifier is not required. The linear amplifier allows power amplification without distorting modulation.

The relation between exciting power and output with load changed in a typical linear amplifier is shown in Fig. 3.11. The linear characteristic is held almost upto the critical output power, and beyond it, saturation is caused.

Saturation is caused when the maximum value of AC voltage generated in the load impedance of plate circuit becomes close to the plate supply voltage, and however large the excitation voltage may be, the plate AC voltage does not exceed the supply voltage.

If the excitation voltage is small, the grid of linear amplifier is always kept at negative potential, not allowing the grid current to flow, and exciting power can be neglected. On the



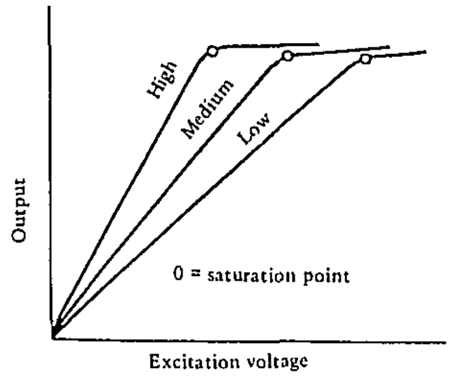


Fig. 3.10 Projected cut-off curve

Fig. 3.11 Input/output characteristics with load impedance changed in linear amplifier

contrary, at the peak of modulation, the grid is driven to positive potential, and the exciting power becomes considerably large.

With regard to the plate efficiency of an ideal linear amplifier, if the tube characteristic is linear, and the plate current flows accurately 180 degrees in each cycle, then the efficiency η is

$$\eta = \frac{(E_b - E_{min}) I_{max} / 2}{E_b I_{dc}} \times 100\%$$

If the plate current is a half-wave of sine-wave, the characteristic is linear and the mean DC plate current is

$$I_{dc} = \frac{2}{\pi} I_{max}$$

Substituting it for the above formula,

$$\eta = \frac{(E_b - E_{min})\frac{I_{max}}{2}}{E_b \frac{2}{\pi} I_{max}} \times 100\% = \left(1 - \frac{E_{min}}{E_b}\right)\frac{\frac{1}{2}}{\frac{2}{\pi}} \times 100\%$$

$$= \frac{\pi}{4}\left(1 - \frac{E_{min}}{E_b}\right) \times 100\% \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3.16)$$

where E_b is plate supply voltage, and E_{min} is the minimum instantaneous value of plate voltage during the cycle. The plate efficiency of maximum output in actual state is about 50 to 60%, as in the case of untuned load.

If the signals to be amplified are 100% modulated carrier waves, the amplitude of the carrier waves is half of the peak, and with no modulation efficiency is half of the maximum efficiency.

(1) Grounded grid amplifier:

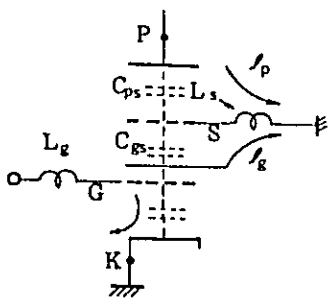
The circuits of grounded-cathode amplifier and grounded-grid amplifier are shown in Fig. 3.12. For amplification of modulated waves, the grounded grid circuit is widely used.

(a) Grounded cathode circuit (b) Grounded grid circuit

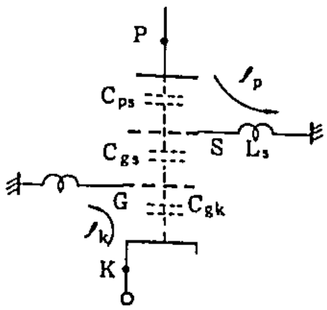
Fig. 3.12 Circuits of amplifiers

Grounded-grid circuit can be used for triodes and tetrodes. The grounded-grid amplifier of tetrodes is the most popular apparatus in VHF band transmitters.

Fig. 3.13 shows the flow of current in tank circuits of grounded-cathode amplifier and grounded-grid amplifier of tetrodes. In the grounded-cathode amplifier of (a), plate current $\mathscr{I}_p$ and control grid current $\mathscr{I}_g$ flow through L_s of screen grid. If the value of L_s is not zero, potential appears at the screen grid. This voltage couples the plate with the grid, to cause oscillation. The occurrence of oscillation is also concerned with C_{pg}. If the value of L_s is made proper, C_{pg} can be cancelled, to neutralize the coupling. An actual grounded-cathode circuit is provided with this neutralizing circuit, but the adjustment is troublesome.



(a) Grounded cathode amplifier



(b) Grounded grid amplifier

Fig. 3.13 Tank currents of amplifiers

The tank current of grounded grid amplifier is as shown in (b), and since currents ℓ_k and ℓ_p flow individually, the coupling is separated. As for the inter-electrode capacity, C_{pk} is coupled instead of C_{pg}, but the value is smaller by one digit, separating the input circuit from the output circuit perfectly, with no fear of oscillation.

A feature of the grounded grid amplifier is that excitation power appears partially in the output, and this is called through-power. Another features is that the input impedance is low. Input impedance Z_{in} is

$$Z_{in} = \frac{\beta_{rp} + R_{\ell}}{\mu(1 + \frac{1}{\mu_s} + \frac{1}{\mu})} \cdots\cdots\cdots\cdots \qquad (3.17)$$

In Class B amplifier, it is constant with $\beta = 2$, but in Class C amplifier, the input impedance depends upon bias value, since β depends upon the operation angle.

(2) Final stage linear amplifier of television transmitter equipment:

Here will be described the circuit with tetrode 7F71R used as final stage linear amplifier for amplification of modulated waves in 1 kW television transmitters. The connection diagram of power amplifier circuit and the conceptual view of the structure are shown in Figs. 3.14 and 3.15.

This amplifier allows to obtain 1.1 kW peak output free from distortion in Class AB operation, since television signals are amplified linearly. Since the input impedance of the circuit is about 26 Ω, an input of 20 to 50 W and 50 Ω impedance is converted, to be applied to the middle-point of filament.

The load of plate circuit is made to resonate with 20 PF output capacity, to obtain about 1,000 Ω output impedance. For conversion to 50 Ω output impedance, M coupling circuit is used with no contact portions to avoid troubles due to variable adjusting portions.

Examples of typical operations are shown in Table 3.1.

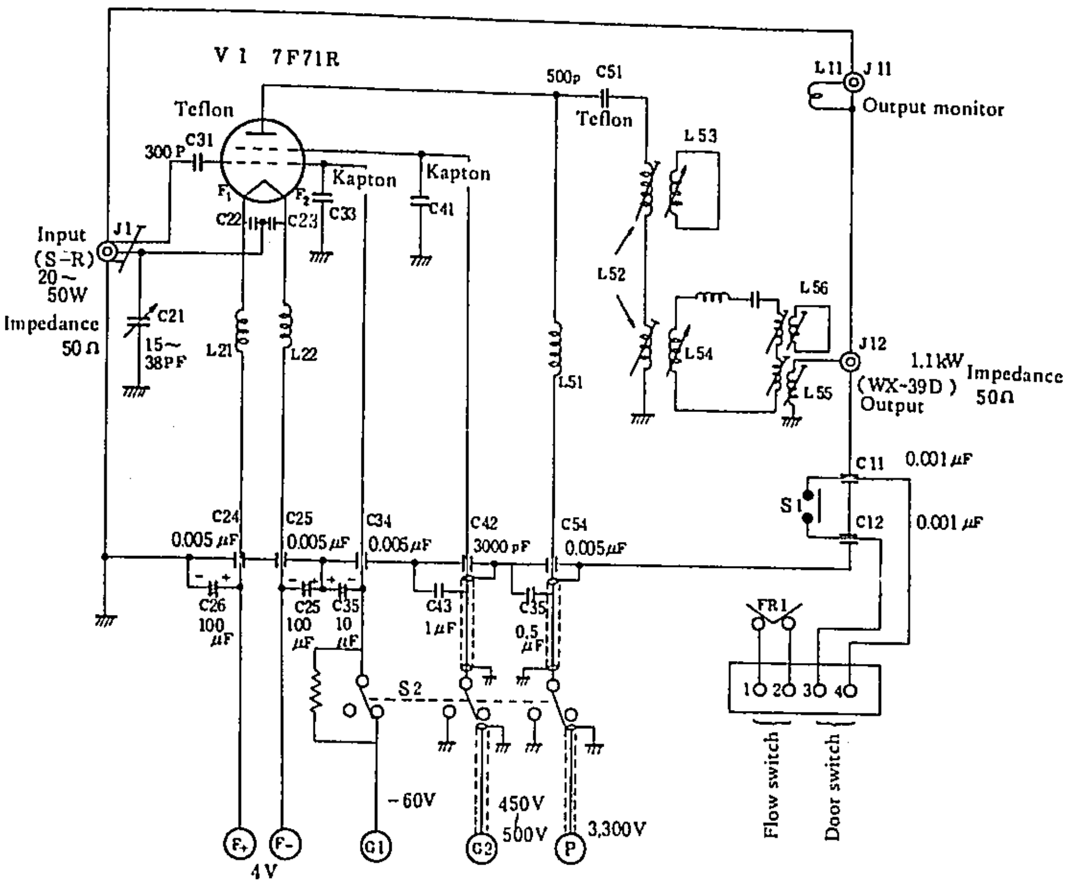


Fig. 3.14 Connection diagram of 7F71R cavity resonator (HCH)

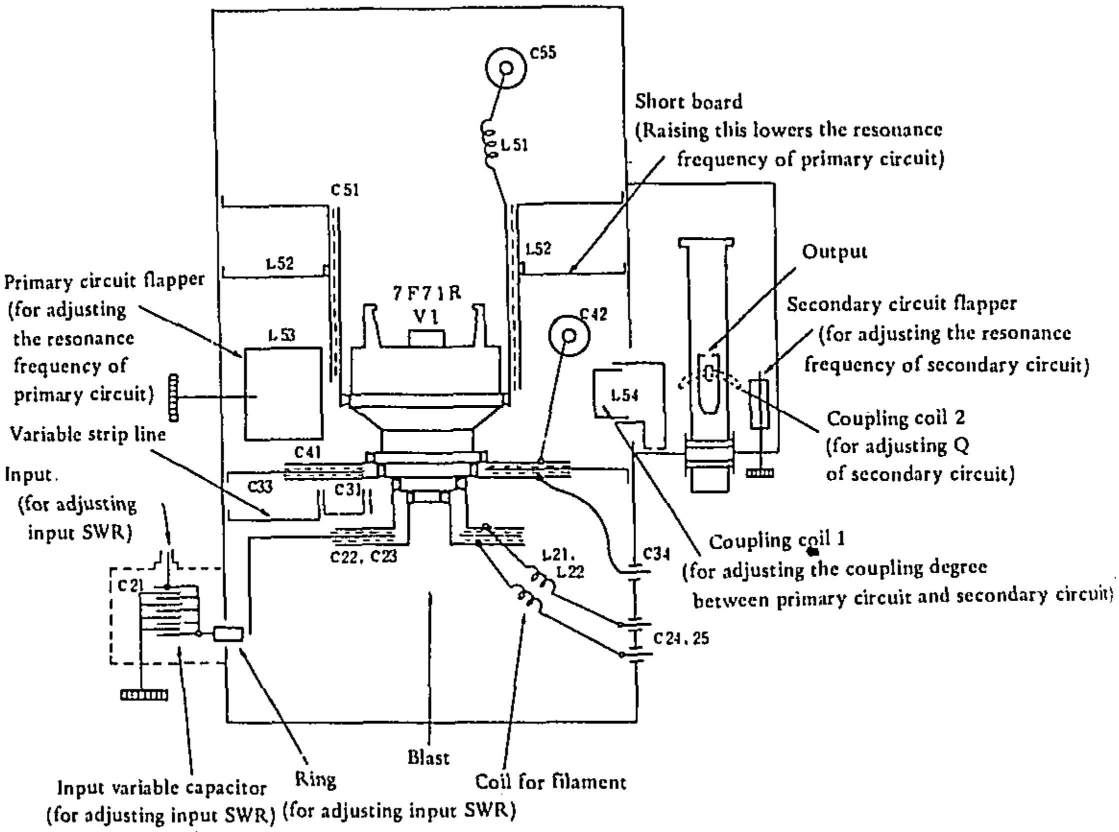


Fig. 3.15 Conceptual view of 7F71R cavity resonator structure (HCH)

Table 3.1 Examples of 7F71R linear amplifier operations

			CW 500W	CW 1,000W	CW 1,500W
Plate voltage	DC	E_b (V)	3,300	3,300	3,300
	Fundamental wave	E_p (V)	1,000	1,410	1,730
Plate current	Peak	i_{pm} (A)	2.0	2.82	2.46
	DC	I_b (A)	0.85	1.17	1.42
	Fundamental wave	I_p (A)	1.0	1.41	1.73
Operating angle of plate current flow		θ_p (deg)	180	160	150
Output		P_0 (W)	500	1,000	1,500
Plate input		P_i (W)	2,805	3,861	4,684
Plate efficiency		η_p (%)	17.8	25.9	32.0
Grid voltage	DC	E_c (V)	-60	-60	-60
	Fundamental wave	E_g (V)	26	38	45
Grid excitation power		P_d (W)	13	27	39
Passing power		P_{th} (W)	13	27	39
Load resistance		RL (Ω)	1,000	1,000	1,000
Excitation input impedance		Z_{in} (Ω)	26	27	26

$E_{SG} = 500$ V

CHAPTER 4. MODULATION METHODS

4.1 High Level Modulation and Low Level Modulation

The modulator of transmitters can roughly be classified into high-level (power) modulation and low-level (power) modulation, and in terms of modulation methods, it includes plate modulation, grid modulation, screen grid modulation and diode modulation, etc. Usually, modulation is provided at one place, but there is a multi-stage modulation system applying modulation at more than one place.

In low level modulation, carrier waves are modulated at a low-level power, and power-amplified by a linear amplifier, to be supplied to the antenna. Since the power is low at the modulation point, high quality modulation is secured. This is very popularly employed in latest television transmitters. Since it is uneasy to amplify modulated waves distortionlessly in radio transmitters, the characteristics in light of the transmitters as a whole are not quite favorable.

In this regard, high-level modulation does not require the amplification of modulated waves, and is used from old times in radio transmitter for its easy adjustment and maintenance. The high level modulation method is still the most popular method now. At the time of starting TV broadcasting, high-level modulation was the most popular method, but with the development of solid state technology, it is being substituted by low-level modulation apparatus which cause few trouble and allow easy maintenance.

4.2 Plate Modulation

If proper grid bias and grid excitation voltage are applied to Class C high frequency amplifiers, the output current of closed circuit is proportional to the plate DC voltage. Fig. 4.1 shows the relation between plate DC voltage and output current, with grid excitation voltage E_g changed. If grid excitation voltage E_g is low, the voltage is proportional to the current at the low voltage range, but proportional relation cannot be obtained in the high plate voltage range. If the grid excitation voltage is high, the proportional relation is maintained, upto further higher plate voltage.

If audio voltage is superimposed on the plate DC voltage, with proper grid excitation applied in radio transmitter, plate modulation can be made.

Plate voltage requires double plate voltage amplitude at the time of 100% modulation peak, compared to that at the time of carrier. This voltage amplitude is supplied from the modulator. When sine-wave modulation is provided by Class C plate modulator as shown in Fig. 4.2, the average high frequency output current is

$$\text{Carrier effective value} \times \sqrt{1 + m^2/2} \text{ times.}$$

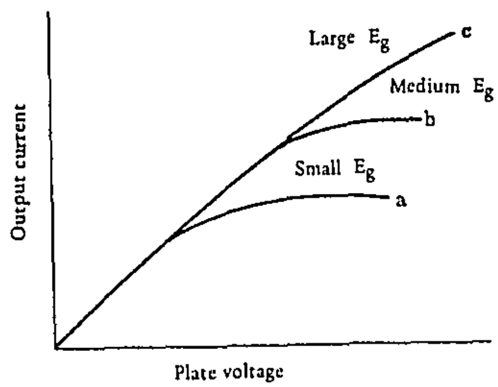


Fig. 4.1 Plate voltage vs. output current with E_g changed

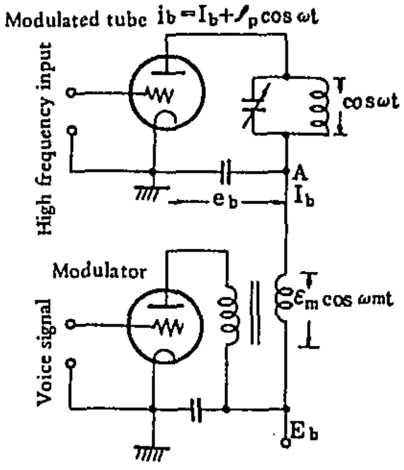


Fig. 4.2 Plate modulator circuit

If modulation degree $m=1$, viz. 100%, the number of times is $\sqrt{1+\frac{1}{2}} = 1.225$. Interms of power, $(1.225)^2 = 1.5$ times. For example, if the output at the time of carrier is 1.1 kW, the output after 100% modulation of sine waves is $1.1 \times 1.5 = 1.65$ kW. The difference 0.55 kW from 1.1 kW is covered by the modulator, and if the efficiency of the modulated tube is 0.78,

$$0.55 / 0.78 = 0.705 \text{ kW}$$

That is, 0.705 kW must be supplied from the modulator to the modulated tube. In general, the output of modulator is designed to be 0.75 of carrier output.

Table 4.1 shows the operation of modulated tube 5T31 used in 1 kW radio transmitter. The tube 5T31 is a triode, and since the output of one tube is 584 W, two of these tubes are used in parallel for a 1 kW transmitter.

Table 4.1 Operation of 5T31 plate modulation

			At the time of carrier	100% modulation positive peak
Plate voltage	DC	E_b (V)	3,000	6,000
	Fundamental wave	E_p (V)	2,790	5,580
Plate current	Peak	i_{pm} (A)	0.9	1.6
	DC	I_b (A)	0.237	0.49
	Fundamental wave	I_p (A)	0.418	0.83
Operating angle of plate current flow		θ_p (deg.)	65	80
Output		P_0 (W)	584	2,320
Input		P_i (W)	711	2,950
Plate efficiency		η_p (%)	82	78
Grid voltage	DC	E_c (V)	-270	-270
	Fundamental wave	E_g (V)	480	555
Grid current	DC	I_c (A)	0.06	0.09
	Fundamental wave	I_g (A)	0.12	0.17
Exciting power		P_d (W)	28	48
High frequency load resistance		R_L (Ω)	6,680 Ω (3,340 Ω 2 tubes)	
Modulation transformer load resistance		R_m (Ω)	12,600 Ω (6,300 Ω 2 tubes)	

The table shows that the voltages and currents of respective electrodes are greatly changed by modulation. Among them, the fundamental wave component of grid voltage must be increased from 480 V to 555 V for modulation. This is why prior stage modulation is required. At the time of 100% modulation, the degree of prior stage modulation is

$$\frac{555 - 480}{480} = 15.6\%$$

Applying the above modulation to the prior stage allows to improve modulation distortion. An actual plate modulator applies modulation of about 10 to 30%.

4.3 Grid Modulation

In radio transmitters, plate modulation is mainly used in light of favorable characteristics and favorable stability. But in television transmitter, because of the difficulty to obtain high voltage video output and impossibility to realize modulation transformers, plate modulation is not used, but grid modulation has exclusively been used. This grid modulation is, however, being gradually substituted by diode modulation.

Grid modulation refers to the method of obtaining modulated waves by applying high frequency voltage to the grid and changing the bias voltage by the video signals, as shown in Fig. 4.3.

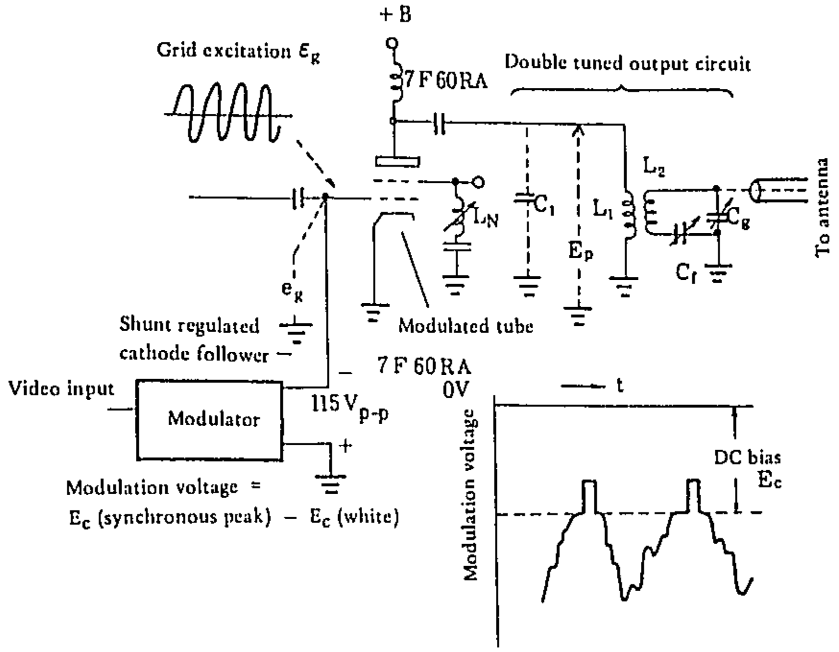


Fig. 4.3 Grid modulator circuit

An example of calculation of operation for delivering a synchronous peak value output of 1.2 kW by using 7F60RA is shown in Table 4.2.

Table 4.2 Operation of 7F60RA grid modulation

			Synchronous peak value (100%)	Pedestal level (75%)	White level (125%)
Plate voltage	DC	E_b (V)	2,700	2,700	2,700
	Fundamental wave	E_p (V)	1,500	1,125	187
Plate current	Peak	i_{pm} (A)	3.2		
	DC	I_b (A)	1.02	0.73	0.1
	Fundamental wave	I_p (A)	1.6	1.2	0.2
Operating angle of plate current flow		θ_p (deg.)	90	81	39
Instantaneous power (average power)		P_0 (W)	1,200	675 (714)	18.7 (164)
Input		P_i (W)	2,754	1,971	270
Plate efficiency		η_p (%)	44.5	34.0	7.0
Grid voltage	Bias	E_c (V)	-80	-110	-195
	Max. voltage	e_g (V)	105	75	-10
	Excitation voltage	ε_g (V)	185	185	185
Load resistance		R_L (Ω)	940	940	940

For grid modulation, a constant high frequency voltage must be applied to the grid of modulated amplifier. However, because the bias changes according to the modulation signal, more grid current flows by synchronous signal and little current flows by white signal. Therefore, since the load of grid terminal is variable viewed from excitation stage, the high frequency voltage tends to decrease by synchronous signal, and tends to increase by white signal. In general, the grid of modulated amplifier receives modulation reverse to that of plate, and this is called inverse modulation. The inverse modulation not only requires more modulation voltage but also is liable to cause phase modulation which causes buzz and aggravation of differential phase.

To decrease inverse modulation, a swamping resistance is put in parallel to the grid, for reducing the influence of change of I_g.

A means to decrease inverse modulation is taken also in circuit configuration. That is, the excitation stage and modulated stage are connected by λ/4 circuit, as shown in Fig. 4.4. Passing through λ/4 circuit allows to reduce the influence of change of input impedance at modulated stage.

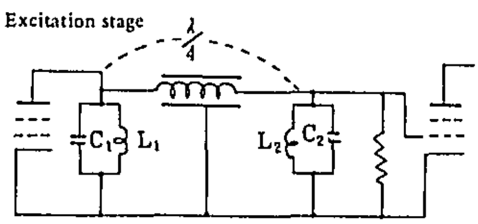
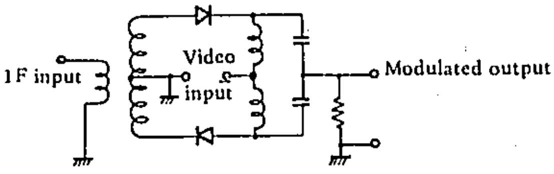


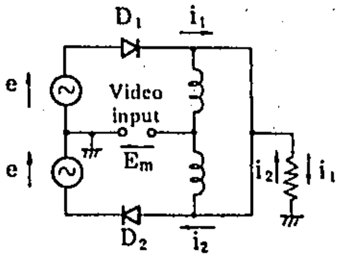
Fig. 4.4 Connection of excitation stage and modulated stage

4.4 Balanced Modulators for Television Low Level Modulation

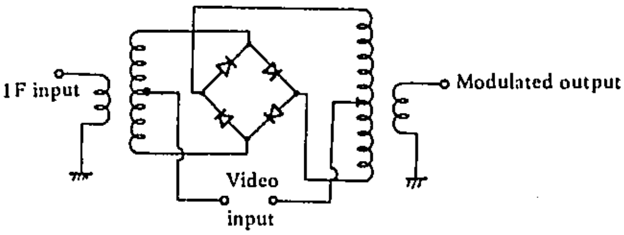
For low level modulation in television transmitter, a balanced modulator with diodes applied is used. In the balanced modulator circuit, 1F carrier input and video input are applied to the modulator, to take out modulated output, as shown in Fig. 4.5.



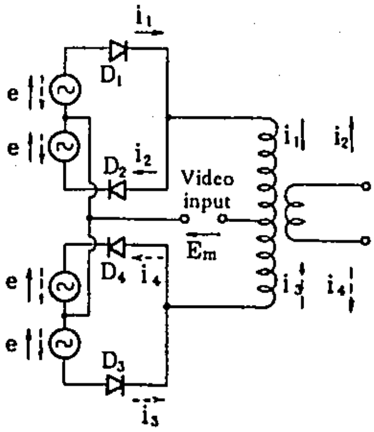
(a) Single balanced circuit



Equivalent circuit of (a)



(b) Double balanced circuit



Equivalent circuit of (b)

Fig. 4.5 Balanced modulator circuit

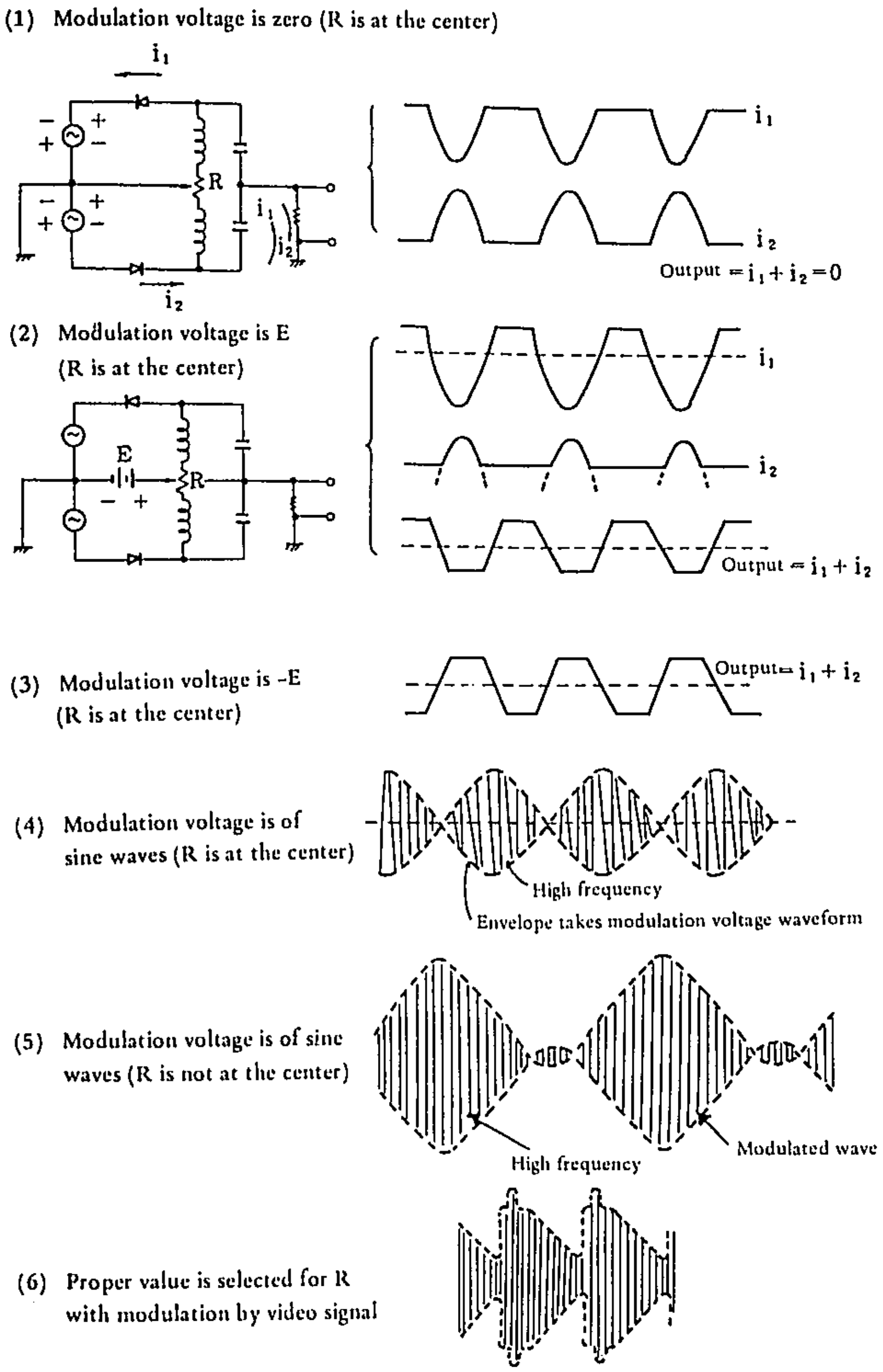


Fig. 4.6 Operation of balanced modulator

The operation of balanced modulation will now be considered, taking the single balanced circuit of (a) for example. Fig. 4.6 shows how the currents of diodes changes according to the presence of modulation voltage and the polarity of modulation voltage.

The polarity of modulation voltage affects the currents flowing in diodes. If the modulation voltage is of sine waves, the carrier suppression double-sideband waveform as shown in the figure appears. In this case, by throwing diode loads out of balance, the carrier component can be left. By selecting a proper value for load resistance R and modulating by video signal, the double-sideband modulated waves like (6) can be obtained. To obtain vestigial sideband waves, a vestigial side-band filter is inserted at the output of balanced modulator, for obtaining standard type signals.

This modulation output is converted into transmission frequency, and is amplified to 30 - 50 W by a solid-state amplifier. The output of the solid-state amplifier is applied to the linear amplifier described in Chapter 3, for power amplification, thereby television video transmitter by low level modulation being formed.

4.5 Frequency Modulation by Reactance Tubes

(1) Theory of reactance tube operation

In Fig. 4.7, if the voltage applied to the plate of tube is ε_p, the voltage is split by C and R to be applied to the grid. If the angular frequency is ω, the voltage applied to the grid ε_g is

$$\varepsilon_g = \frac{R\varepsilon_p}{\frac{1}{j\omega c} + R} \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (4.1)$$

If R and C are selected to have $\frac{1}{\omega c} \gg R$, formula (4.1) is

$$\varepsilon_g = j\omega c\, R\varepsilon_p \quad \cdots\cdots\cdots\cdots\cdots\cdots \quad (4.2)$$

If the mutual conductance of tube is gm, plate current $\mathcal{J}_p$ is

$$\mathcal{J}_p = gm\,\varepsilon_g = j\omega c\ R\ gm\ \varepsilon_p \quad \ldots\ldots\ldots\ldots \quad (4.3)$$

Therefore, the impedance viewed from plate side to tube is

$$\frac{\varepsilon_p}{\mathcal{J}_p} = \frac{1}{j\omega c\ R\ gm} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (4.4)$$

Thus, it is capacitive, and its value changes according to gm. Therefore, if this reactance tube is connected in parallel to the tank circuit of oscillator, and voice signal is super-imposed on the grid voltage of reactance tube, to change gm, the oscillation frequency is changed according to the voice signal, allowing frequency modulation.

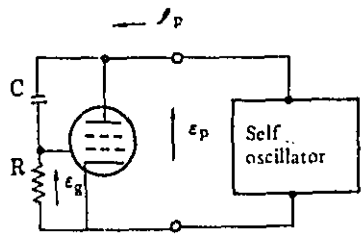


Fig. 4.7 Theory of reactance tube

The reactance tube is desirable to be a tube, the mutual conductance of which changes linearly to bias. As obvious from the above theory, if feedback with the phase difference of 90 degrees from the plate voltage is given to the grid and the grid voltage is changed by the audio signal, a reactance tube is provided.

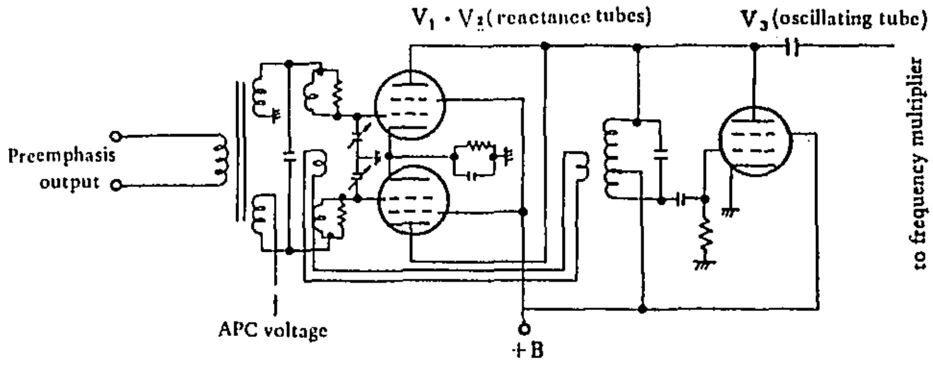


Fig. 4.8 Reactance modulator

(2) Actual reactance frequency modulator

An actual reactance frequency modulator is as shown in Fig. 4.8. The pick-up output from an oscillating circuit is applied to the grids of reactance tubes V_1 and V_2 by coils respectively with phase difference of ±90 degrees. Since the two reactance tubes operate like a push-pull circuit, the frequency deviation becomes wide, and linearity becomes better, allowing to obtain favorable characteristics.

Since a frequency modulator by reactance tubes uses a self oscillator, the stability of the modulator itself is not good. Therefore, a frequency stabilizing circuit, for example APC (automatic phase control) circuit must be provided, to keep transmission frequency stable.

4.6 Frequency Modulation by Variable Capacitance Diodes

For frequency modulation, the frequency of self oscillator must be changed in proportion to the magnitude of modulation input. Reactance tubes were used for this purpose, but recently, variable capacitance diodes are used. How the inter-terminal capacitance changes when bias voltage is changed by variable capacitance diodes is shown in Fig. 4.9, as an example.

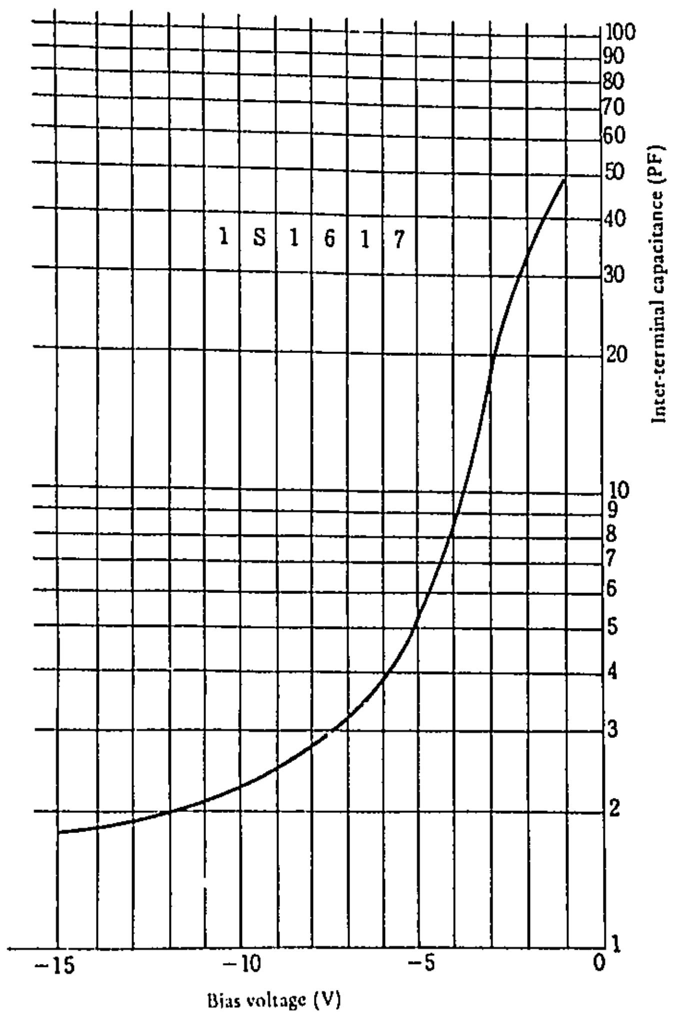


Fig. 4.9 Characteristic of variable capacitance diodes

An example of frequency modulator circuit is shown in Fig. 4.10. Three variable capacitance diodes are inserted in parallel to the oscillating circuit of self oscillator with a transistor applied, for frequency modulation. In this case, the self oscillation frequency is about 15 MHz, and the frequency deviation is about 12.5 kHz. A frequency multiplier must be provided to obtain necessary carrier frequency and frequency deviation.

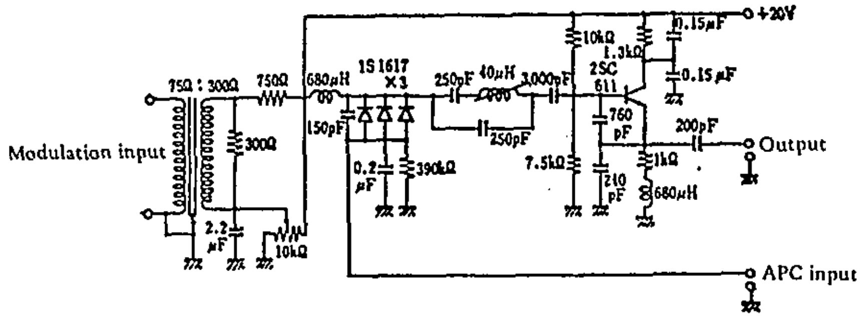


Fig. 4.10 Frequency modulation by variable capacitance diodes

Since this circuit is a self oscillator, the stability of frequency is not good. Therefore, a APC circuit is added to detect the discrepancy of frequency, and to feedback positive or negative voltage corresponding to the discrepancy to the variable capacitance diodes, for automatic frequency correction.

The modulation output is applied to a frequency multiplier and a solid-state amplifier, and then to a power amplifier for power amplification, thereby frequency modulation transmitter being formed.

4.7 Serrasoid Modulation

Serrasoid modulation was popularly used for television aural transmitters, before the frequency modulation by variable capacitance diodes came into practical use. This method is still employed in some stations. Fig. 4.11 is an example of serrasoid modulator circuit. The output of crystal oscillator is fed through oscillator, Class C amplifier, differentiation circuit, saw-tooth generator and linearity corrector between V_{1a} and V_{3b}, to be correct saw-tooth waves, being coupled with the grid of V_{4a}.

The cathode voltage of V_{4a} is adjusted to allow the plate current to flow when the amplitude of saw-tooth wave becomes 1/2, and the plate voltage of V_{4a} is set to a low value. By this arrangement, the grid current flows in a very short time after the plate current starts flowing, to stop the charge of saw-tooth generator, and the latter half of a saw-tooth wave is clipped. These relations are shown in Fig. 4.12. In Fig. 4.12 (a), since V_{4a} is cut-off between a and b, the plate voltage rises, but since current starts flowing at b, the plate current saturates in a short time, and the plate voltage drops suddenly like B-C. Between c and d when the plate current flows, the plate voltage drops at C-D, but at d, V_{4a} is cut-off suddenly, and the plate voltage rises by the time constant of plate load 30 kΩ of V_{4a} and output capacity, like D-E. At e, V_{4a} conducts again, and the plate voltage drops.

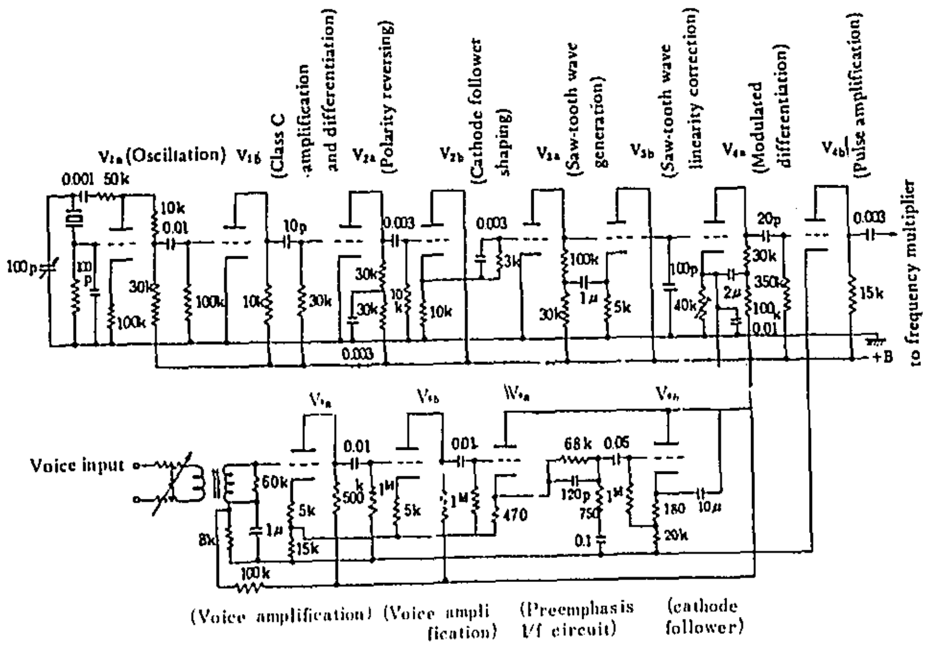
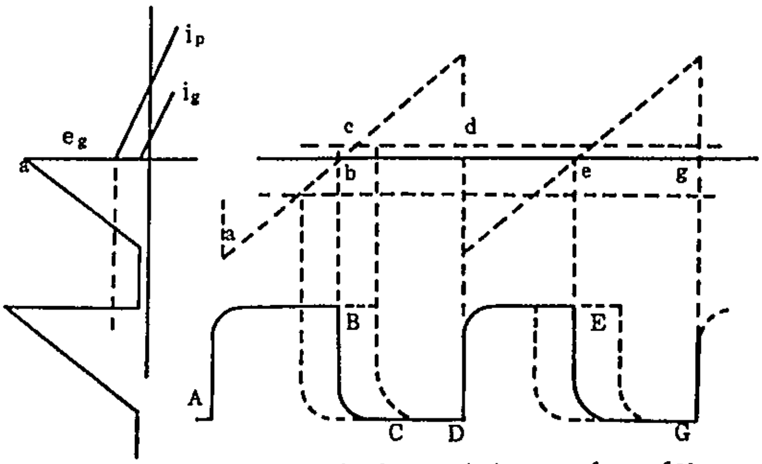


Fig. 4.11 Circuit of serrasoid modulator



(a) Grid voltage and plate waveform of V_{4a}

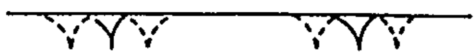

(b) Grid voltage waveform of V_{4b}

Fig. 4.12 Waveforms at the respective portions of V_4

If voice signal is applied to the cathode of V_{4a}, the point b corresponding to the cut-off voltage changes along the gradient of saw-tooth wave, and therefore the point for the current to start flowing, viz. B-C changes. And the plate voltage changes at the after-edge of square wave according to voice voltage.

The plate waveform of V_{4a} is differentiated and is applied to the grid of V_{4b}. The resistance 350 kΩ of differentiation circuit is connected to +B circuit. These are shown in Fig. 4.12 (b). The negative pulses changed in phase by voice voltage are amplified by V_{4b}, and are applied to the next frequency multiplier, to be the required FM waves. Since these waves are phase modulated, 1/f circuit must be provided for the modulation input circuit.

Symbols V_5 and V_6 indicate voice amplifiers. Between the cathode of V_{6a} and the grid of V_{6b}, there are inserted a 1/f circuit and a preemphasis circuit.

In serrasoid modulation, since the transmission frequency is determined by a crystal oscillator, the frequency is stable. However, because wide-band modulation cannot be made and the circuit is complicated, it is not used in new equipment.

4.8 Radio High Efficiency Modulation

In the plate modulation of radio, the larger the voltage utilization factor ε_v and the current utilization factor ε_i, the higher the efficiency. Therefore, if the waveform of plate output voltage and current is closer to square, the output can be obtained more efficiently. As a method to make the plate waveform square, an even-harmonics or odd-harmonics order tank circuit is provided to the plate circuit, to synthesize fundamental wave and higher harmonics.

The theoretical view of this circuit is shown in Fig. 4.13. To obtain ideal square waves, higher harmonics of higher order are required, but to simplify the circuit configuration, 3rd harmonic is actually used, for obtaining output waveform close to square. In synthesizing fundamental wave with 3rd harmonic, the waveform becomes the closest to square, when the amplitude of 3rd harmonic becomes about 1/6 of that of fundamental wave. For this purpose, a circuit to obtain 1/6 amplitude is provided by $3f_0$ tuning circuit on the grid and plate sides.

At the time of carrier, the plate efficiency is im roved to high 85 - 90%, being higher than general plate modulation by 5 to 10%. In the case of large output radio transmitting station, the plate efficiency greatly affects the size of power supply equipment. Circuits of higher efficiency will be considered also in future.

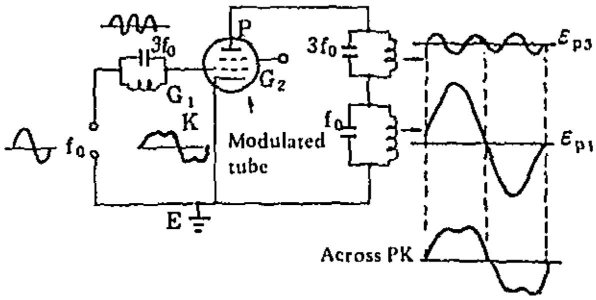


Fig. 4.13 Circuit for high efficiency modulation

CHAPTER 5. MICROWAVE TUBES

Tubes operating in different theory from that of electrode tubes, in ultra-high-frequency, such as klystrons and travelling wave-tubes, are called microwave tubes.

In some cases, the electrode tubes are used as amplification tubes for UHF television translators. But microwave tubes which meet the requirements of solid-state circuits, facilitate the maintenance and adjustment are used for transmitters of high-power stations. The features of microwave tubes are

- High gain with large output power.
- Less discrepancy in adjustment and maintenance.
- Wide-band amplification facilitates the manufacture of transmitters and alternation of frequency.
- Low efficiency makes power supply unit large in size, and power consumption is large.
- Because of the above, cooling system becomes large in size.
- Requires long pre-heating time and therefore, interruption of service becomes long.
- Large structure, heavy in weight makes replacement work troublesome.

5.1 Klystrons

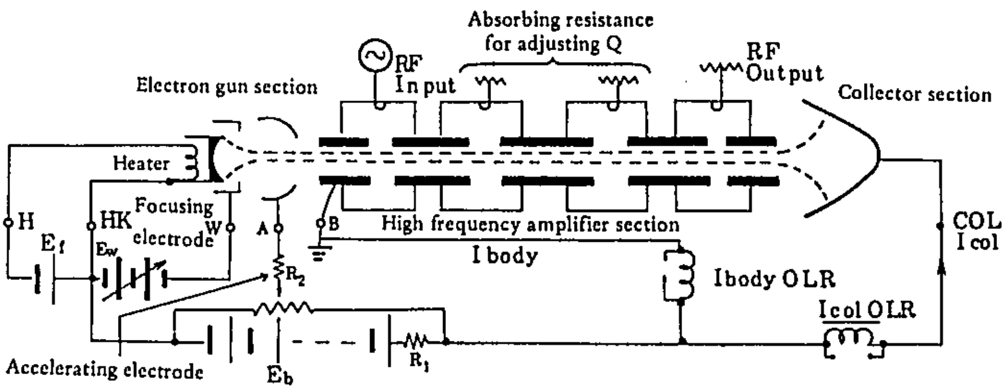
Klystrons are also called velocity modulation tubes. They are now widely used for final stages of UHF television transmitter with output of 3 kW or more. Mainly used Klystrons of 4 cavity resonators are mostly used.

Power gain of a klystron is about 40 dB, being very high, and 1 W excitation power is sufficient to obtain 10 kW output. Klystrons are very effective to simplify the structure of equipment by employing solid-state circuits completely up-to the excitation stage.

(1) Structure and operation theory of klystrons

Fig. 5.1 shows the basic connection diagram of a klystron amplifier. As obvious from the diagram, it comprises three major sections, the electron gun, high frequency amplifier and collector. High frequency is not applied at all to the electron gun section and the collector section. This is a great difference between electrode tubes and klystrons. Even if the frequency becomes high, there is a degree of freedom in design to lower the current density of cathode and the heat-loss density of collector, being the main reason to allow production of highly reliable tubes.

In a klystron, DC electron beams flow at a constant velocity in the high frequency amplifier section. When the DC electron beams pass through the gap of input cavity resonator, some electrons are accelerated and the others are decelerated, according to the phase of high frequency input voltage at that time. This is called velocity modulation.



Ef : Heater power supply
Ew : Wehnelt power supply
Eh : Beam power supply
R_1 : Short protection resistance
R_2 : Flash protection

Fig. 5.1 Basic connection of klystron amplifier

While velocity-modulated beams run through the sapce free from electric field (drift space), bunching action causes density modulation, to generate high frequency output in the output cavity resonator. Fig. 5.2 shows this state, indicating the relation between time and position of individual electrons. Therefore, gradients of straight lines express the velocities, and the density of lines expresses current density.

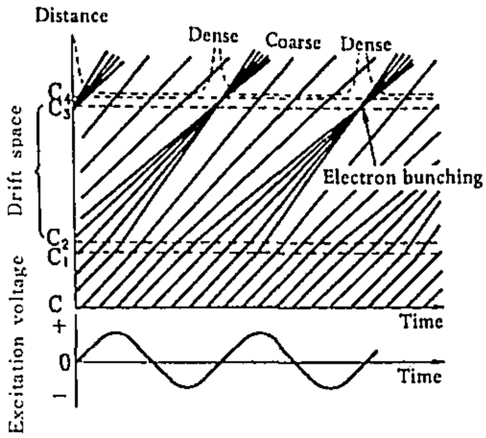


Fig. 5.2 Electron motion of klystron

When density-modulated electron current passes through the cavity gap of high frequency output, induced current flows in the cavity resonator, allowing to take out amplified UHF output power.

(2) Band characteristics

The respective cavity resonators of multi-cavity klystron are not coupled at all in terms of circuit. The respective cavity resonators are coupled by electron beams only. If the coupling of cavity resonator is disregarded, the frequency characteristic of the entire klystron will be expressed by the product of 4 cavity resonator characteristics. That is, the situation is as if there are 4 parallel resonance circuits. Therefore, the band can be windened in the same way as stagger adjustment made for wide band amplification by the intermediate frequency amplifier of television receiver.

For example, by adjusting the 1st cavity resonator to f_0 + 2.5 MHz, 2nd cavity resonator to f_0 + 1 MHz, 3rd cavity resonator to f_0 + 6 MHz and 4th cavity resonator to f_0 + 1 MHz, the overall band characteristic is arranged to have a band of 8 MHz with -1 dB attenuation. Fig. 5.3 shows these relations.

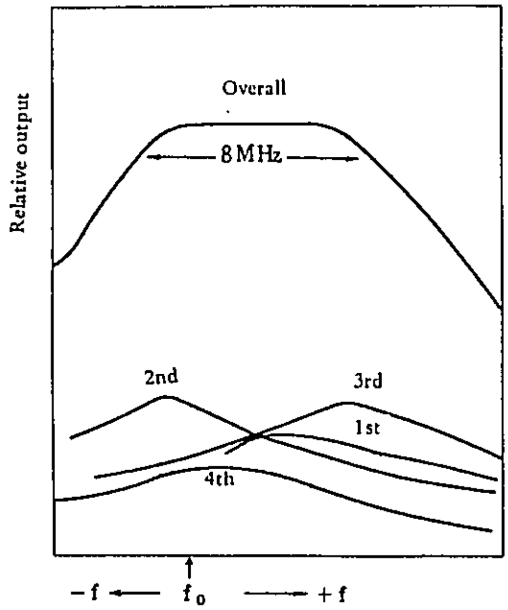


Fig. 5.3 Band characteristics of klystron

(3) Operation characteristics

The output characteristics of 1AV56 chiefly used for 5 to 10 kW UHF television transmitters are shown in Fig. 5.4.

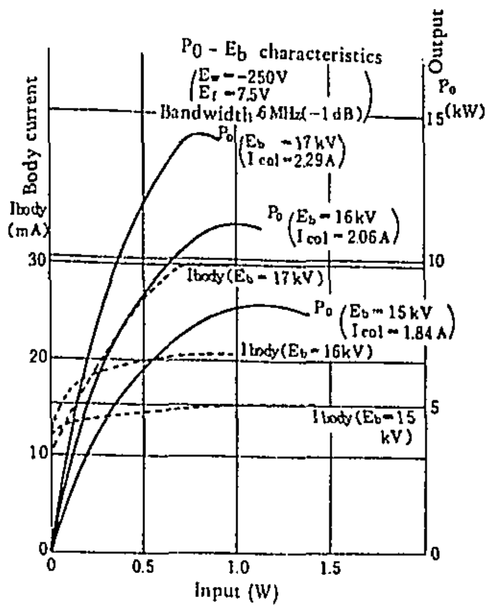


Fig. 5.4 Output characteristics of 1AV56

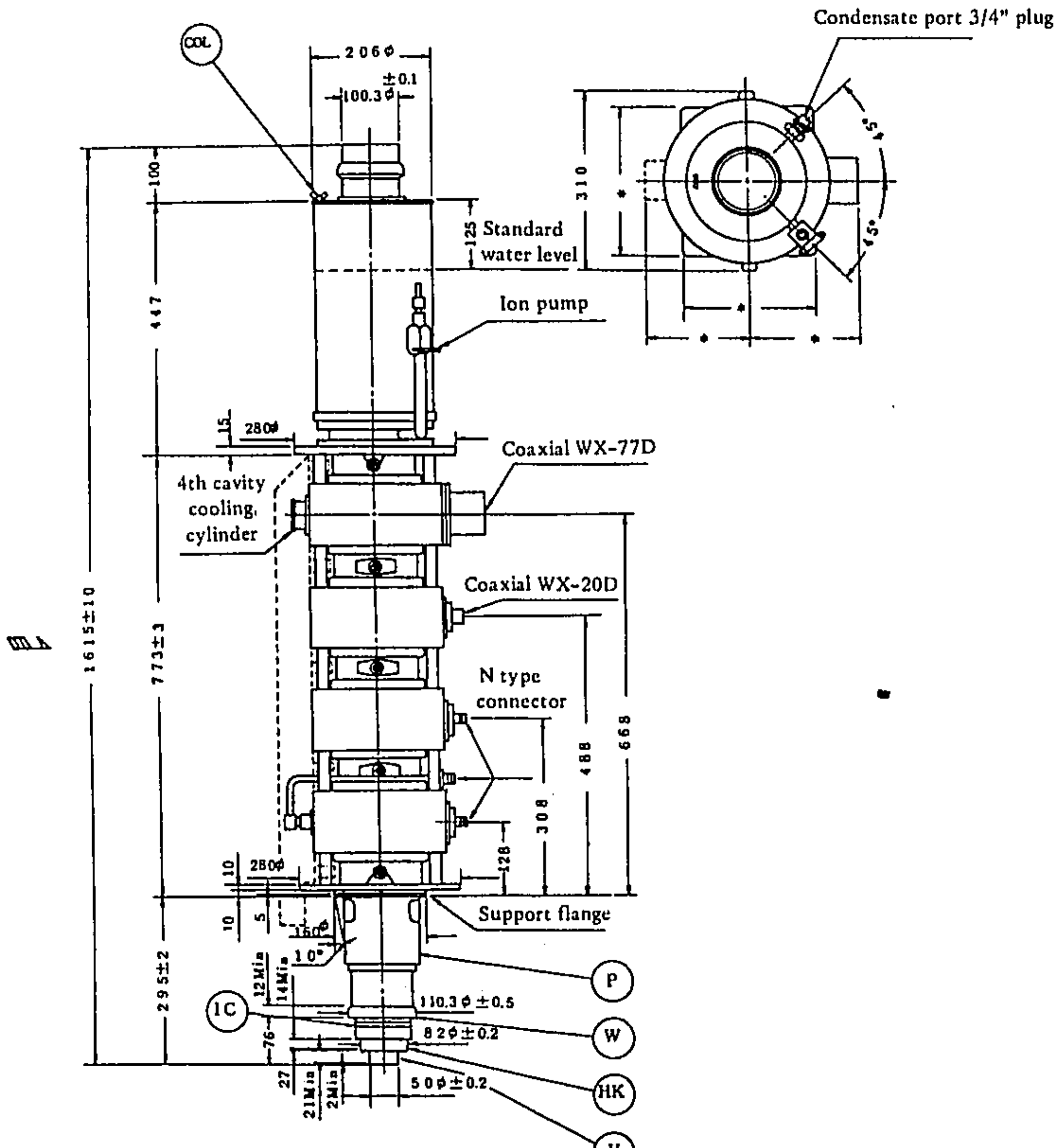
If Eb voltage is changed to 15 kV, 16 kV and 17 kV, the maximum output becomes larger. Actual apparantus are used at 13 to 15 kV for 5 kW output, and at about 17 kV for 10 kW output. In any case, with 1 W input power, sufficiently saturated output can be obtained.

The ratio of output to collector input power, viz. efficiency is about 25 to 30%, when used for video transmitters. The rest becomes heat, requiring a large-scale cooler.

(4) Specifications of klystrons

Table 5.1 shows the specifications of typical klystron 1AV56, and Fig. 5.5 shows its external view.

In mm



Indicates front side.

* External form depends upon application to some extent

--- Indicates the state with cooling air cylinder fitted

Fig. 5.5 External view of 1AV56

Table 5.1 Specifications of 1AV56

		1AV56
Cathode		Matrix type
Heater voltage		7.5 V
Heater current		33 A
Cavity type		Package 4 cavities
Focusing		Electromagnet
Overall length		1,579 mm
Diameter of largest section		385 mm
Weight		Approx. 100 kg
Cooling	Collector	Evaporation cooling
	Body	Forced air cooling
	Output cavity	Forced air cooling
	Electron gun section	Forced air cooling
Max. collector loss		60 kW
Operation example	Frequency	Specified CH
	Band (-1 dB)	6 MHz
	Output	11 - 11.5 kW
	E_b	17 kV
	Cathode current	2.65 A
	Power gain	37 dB

5.2 Traveling Wave Tubes

Traveling wave tubes are used for output stage of 1 kW class UHF television transmitters. A traveling wave tube is a distributed constant amplifier to create high frequency field traveling almost at the same velocity as electron current, for continuous interaction of electron beams.

Many traveling wave tube are also used for 10 W to 100 W class television translators, but together with the development of solid-state circuits, they are gradually being substituted by solid-state circuits in light of saving of power consumption and improvement of reliability, except for high power transmitter.

(1) Structure and operation theory of traveling wave tubes

Fig. 5.6 shows the theoretical structural diagram of a traveling wave tube. As shown in the diagram, it comprises electron gun section to emit electron beams, helix section for high frequency amplification and collector section.

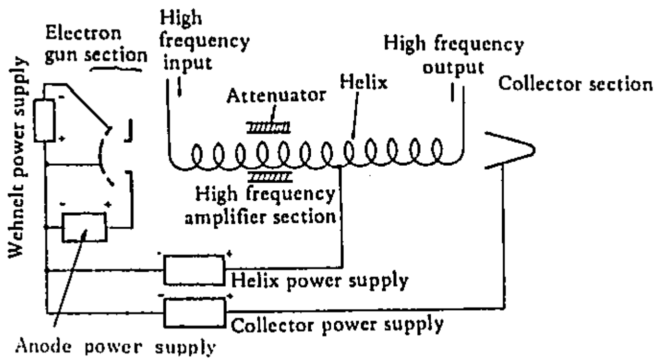


Fig. 5.6 Theoretical structural diagram of traveling wave tube

The helix is the most important portion for amplification, and is respectively connected to input and output, being well matched to the frequency in use. Intermediately, an attenuator for prevention of oscillation is provided, and separated in terms of high frequency. The pitch of the helix is arranged to make the velocity of waves on the helix be almost equal to the velocity of electrons.

If high frequency input is applied to the helix, axial electric field is generated inside along the spiral, and the field trav travels according to the progress of high frequency. The propagation velocity of UHF waves in the axial direction of spiral becomes slow, to be almost as fast as the velocity of electron beams passing through it, as shown in Fig. 5.7. The field of UHF waves forms accelerating fields and decelerating fields for electron beams, traveling together with electron beams toward the output. Therefore, the electrons in the accelerating fields are always accelerated to be higher in velocity, and the electrons in the decelerating fields are always decelerated, to be lower in velocity. That is, electron beams are velocity-modulated in the helix section.

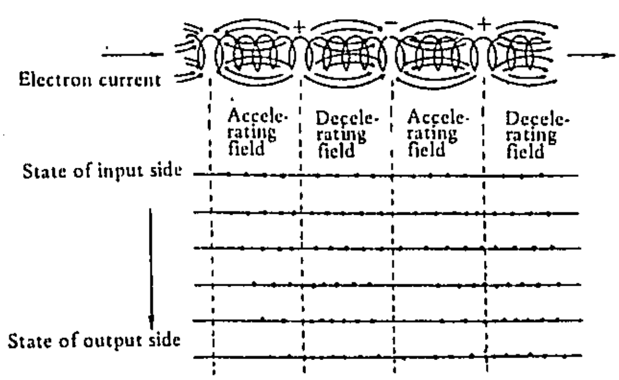


Fig. 5.7 Bunching of electrons in the decelerating fields of output side

The velocity-modulated electron current forms further intensely coarse and dense portion, to be density-modulated, as it travels to the output side.

If the axial propagation velocity of UHF wave field is made a little higher than the velocity of electron beams, the electrons accelerated in the accelerating fields enter into the forward decelerating fields. Therefore, the number of electrons in the decelerating fields increases gradually. As soon as the electrons in the decelerating fields are decelerated, the motion energy is converted into UHF energy. Since the number of electrons in the decelerating fields increases according to approach to the output side, the energy of microwaves is amplified.

Since a traveling wave tube is very large in amplification degree, mismatched impedance at the input and output ends generates reflected waves, causing oscillation by positive feedback. Therefore, an attenuator is put almost at the center between input and output, to attenuate the microwaves velocity-modulated at the input side, and to attenuate the reflected waves generated at the output end.

(3) Operation capabilities

Fig. 5.8 shows the input/output characteristics of 1W80 used for UHF 1 kW telecasting equipment. By 1 W high frequency input, about 1.8 kW output can be obtained. The gain is 32.5 dB. Collector power consumption is 7.2 kW, and the efficiency is about 25%.

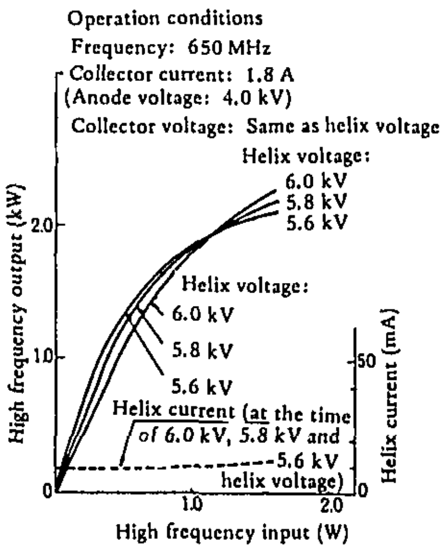


Fig. 5.8 Input/output characteristics of 1W80

Since a traveling wave tube as a whole forms wide-band, it has no portion to be adjusted with power source removed. Therefore, maintenance is very easy.

(4) Specifications of traveling wave tube

Table 5.2 shows the specifications of typical traveling wave tube 1W80, and Fig. 5.9 shows the external view.

Table 5.2 Specifications of 1W80

		1W80
Cathode		Heater-type impregnated cathode
Heater voltage		6.0 VDC
Heater current		12.5 A
Overall length		1,300 mm
Diameter of largest section		300 mm
Weight of main body		Approx. 30 kg
Weight of electromagnet		Approx. 70 kg
Cooling		Forced air cooling
Max. collector loss		14 kW
Operation example	Frequency	650 MHz
	Collector voltage	5.6 kV
	Helix voltage	5.6 kV
	Collector current	2.0 A
	Helix current	15 mA
	Excitation power	0.63 W
	Output	1.26 kW
	Power gain	33 dB

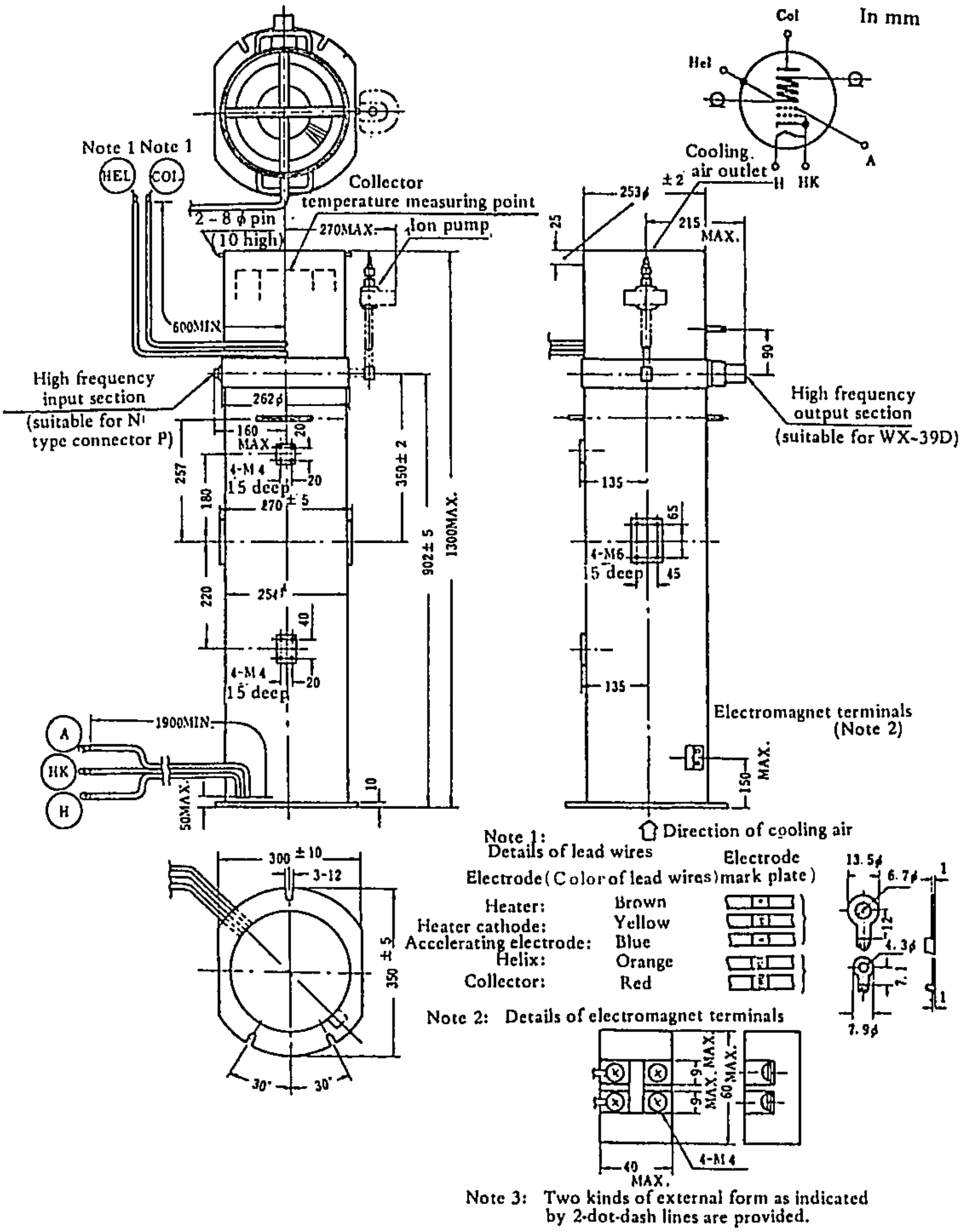


Fig. 5.9 External view of 1W80