MEDIUM WAVE BROADCAST TRANSMITTER

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Medium Wave Broadcasting Transmitter

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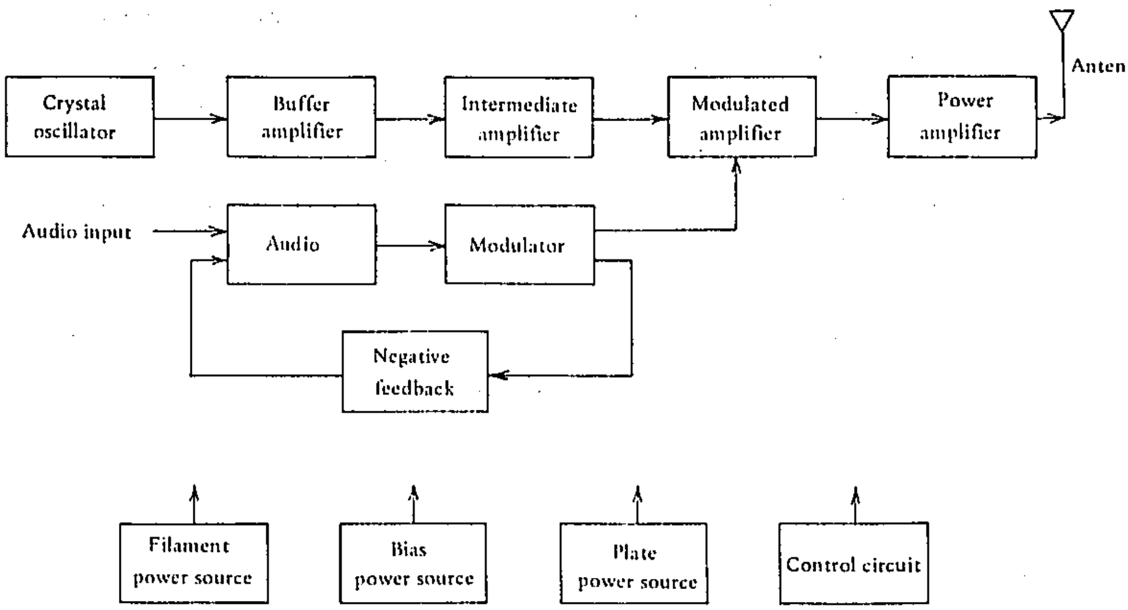


Fig. 1 Configuration of medium wave broadcast transmitter

Fig. 1 shows an configuration of medium wave broadcast transmitters. With regard to the high-frequency part, the high-frequency current generated in the crystal oscillator is amplified by the buffer amplifier and the intermediate amplifier, and then applied to the modulated amplifier for modulation. It is then fed through the power amplifier to the antenna. A buffer amplifier is a high-frequency amplifier which prevents the influence of modulation to vary the oscillation frequency of the crystal oscillator. The power amplifiers is in general, not used except for particular cases.

On the other hand, the audio input to the transmitter is amplified by one or more stages of audio amplifiers, and then applied to an audio power amplifier called modulator, to be modulated by its power. In the modulation system, negative feedback is usually applied and can be classified into feedback of modulation output and feedback of rectified output of modulated wave.

In modulating broadcast transmitter, modulation applied to last stage of high-frequency amplifier and fed directly to antenna is called high-power modulation. On the other hand, modulation applied at low-power stage of high-frequency and then amplified by power amplifier, before entering antenna is called low-power modulation.

As the modulation power required for low-power system is low, the design and construction of modulator is easy, but it is difficult to amplify the modulated high-frequency without distortion. As the high-power modulation system requires a large amount of modulation power, the design and construction of modulator is difficult but adjustment and maintenance is easy.

Recently, broadcasting transmitters of high power-efficiency are required, and in this regard, the high-power modulation system is advantageous. This is because in case of low-power modulation system, the power-efficiency of linear amplification of modulated wave is substantially low. To fulfill this shortcoming, a high efficiency amplification system, such as Doherty Amplification is deviced, but it is not used so much because of difficulty in maintenance.

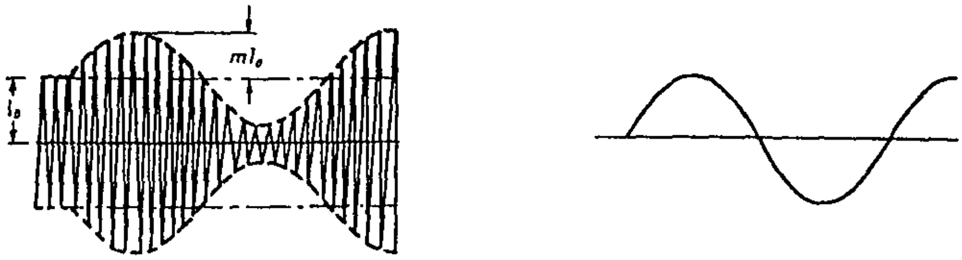
In respect to the modulated tube, modulation is classified into grid-modulation and plate-modulation, according to whether the audio voltage is applied to the grid or the plate.

As the modulation power of grid-modulation is low, the modulator can be small in size, but the modulation characteristic is not so good and the plate-efficiency of modulated amplifier is low. However, the modulation characteristic can be improved by applying negative feedback.

The modulation characteristic of the plate modulation system is excellent and the plate-efficiency of the modulated amplifier is also high. However, the modulator itself requiring a large amount of power, necessitates an excellent power-efficiency and characteristic. In this respect, at present, the adoption of class B pushpull circuit with high-efficiency modulator completely satisfies this purpose.

2. Modulation

The medium-wave broadcasting employs amplitude modulation (AM). Namely, the amplitude of the high-frequency current is varied in accordance with the waveform of audio current as shown in Fig. 2.



(b) Modulated wave

(a) Modulation wave

Fig. 2 Amplitude modulation

Presuming a sinusoidal audio waveform of frequency f_m, and an amplitude of high-frequency current before modulation I_0, the amplitude after modulation is expressed by the following formula.

$$I = I_0 + mI_0 \cos \omega_m t \qquad (1)$$

where m indicates the degree of amplitude change, i.e. modulation degree, $\omega_m = 2\pi f_m$

Therefore, the instantaneous value of high-frequency current is expressed by the following.

$$i = I_0(1 + m \cos \omega_m t)\sin \omega_0 t \qquad (2)$$

where $\omega_0 = 2\pi f_0$ and f_0 is the frequency of high-frequency current.

By expanding the f. (2), the following is obtained.

$$i = I_0 \sin \omega_0 t + mI_0 \cos \omega_m t \sin \omega_0 t$$

The second term of the above equation includes a product of sine and cosine which can be expanded by the trigonometric function equation to a function of added angles and subtracted angles.

Therefore, (2) can be expressed as follows.

$$i = I_o \sin \omega_o t + \frac{mI_o}{2} \sin(\omega_o - \omega_m)t + \frac{mI_o}{2} \sin(\omega_o + \omega_m)t \quad (3)$$

From the above expression, it is understood that the modulated wave consists of the summation of the following three terms. The first term expressed by $I_o \sin \omega_o t$ is called carrier-wave, the amplitude of which is constant regardless of the modulation and is equal to the mean-value of the amplitude of the modulated wave. The second and third terms have identical amplitudes. However, one of the frequency is lower than the carrier frequency by the amount of the modulation frequency and the other is higher by the same amount. These two components are called sidebands. Their lower one is called lower sideband and the higher one is called upper sideband. The sidebands of modulated waves play a role of transmitting information with modulated waves, and the amplitude is determined by the modulation degree.

Fig. 3 (b) is a vectorial presentation of the modulated wave expressed by (3). If we presume a referential vector of the angular-velocity of carrier wave ω_o, since it is constant, then the upper sideband U is expressed by the counter clockwise vector because it has an angular velocity of $(\omega_o + \omega_m)$ and the lower sideband L is expressed by the clockwise vector because of an angular velocity of $(\omega_o - \omega_m)$. The synthesis of the upper and the lower sidebands are carried out always in the same or reverse phase relationship and these are added or cancelled in correspondence with the phase of the modulated wave, resulting in sole variation of amplitude.

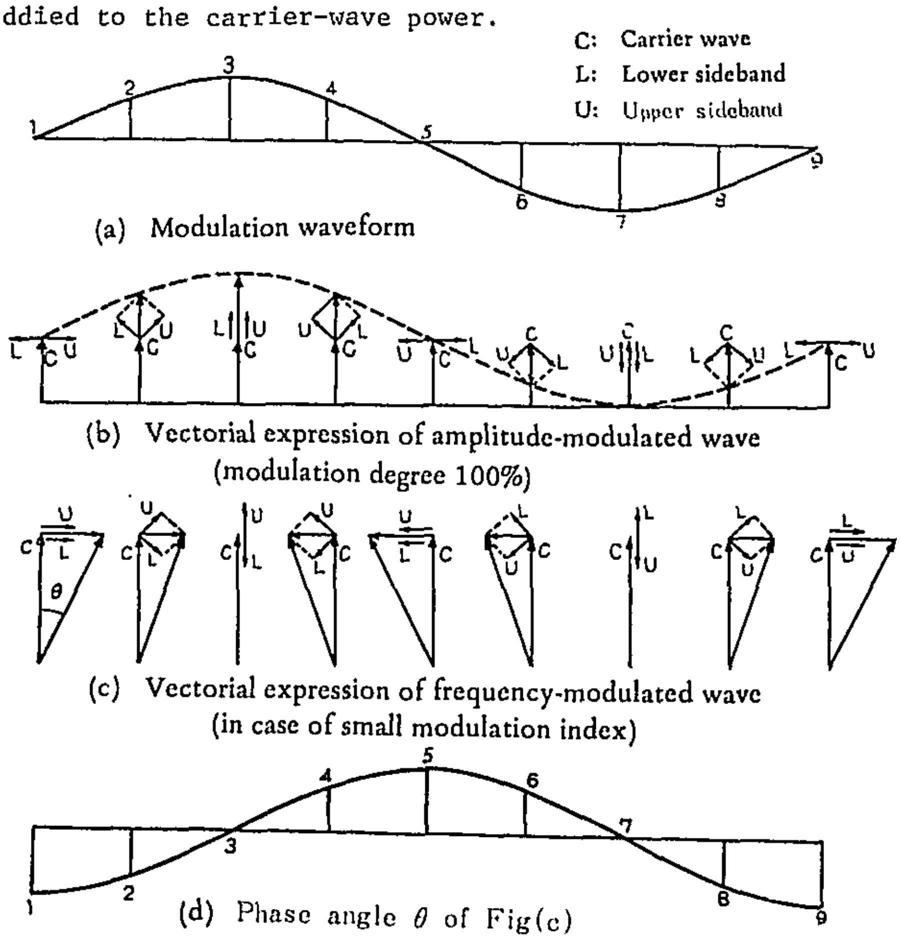
The effective value of the modulated high-frequency varies in accordance with the modulation. Although the effective value of each high-frequency cycles is $1/\sqrt{2}$ of the amplitude as already known, the modulated wave is calculated by the root mean square of the amplitude, since the amplitude itself changes in accordance with the equation (1).

$$\text{Effective value of the modulated wave} = \frac{I_o}{\sqrt{2}} \sqrt{\frac{1}{T}\int_0^T (1 + m \sin \omega_m t)^2 \, dt}$$

$$= \frac{I_o}{\sqrt{2}} \sqrt{1 + \frac{m^2}{2}}$$

where $T = \frac{1}{f_m}$

Thus the effective value of the modulated value becomes $\sqrt{1 + m^2/2}$ times that of the carrier-wave. In case of $m = 1$, i.e. 100% modulation, it becomes $\sqrt{1 + 1/2} = 1.225$ times. The reason why the effective value of the modulated wave increases is because a sideband power is of an amount of (carrier power x $m^2/2$) is addied to the carrier-wave power.

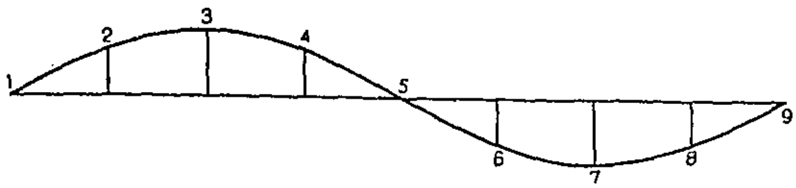


(a) Modulation waveform

(b) Vectorial expression of amplitude-modulated wave (modulation degree 100%)

(c) Vectorial expression of frequency-modulated wave (in case of small modulation index)

(d) Phase angle θ of Fig(c)



(c) Angular velocity $= \frac{d\theta}{dt}$

Fig. 3 Vectorial expression of the modulated wave

The above description relates to the case of modulation provided by a single sinusoidal wave. However, in audio current of actual programs, various components are contained. Therefore, when a carrier wave is modulated by this kind of signal, the frequency spectrum will be a distribution as shown in Fig. 4. If we presume the maximum frequency of the audio current as f_m, then the width of side-bands will become 2 f_m. If the tuning of the tuning circuit of the transmitter (including antenna system) is excessively sharp, the ends of both side-bands, i.e. higher component of audio frequency will be eliminated.

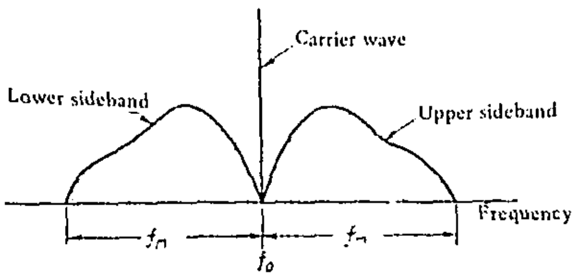


Fig. 4 An example of frequency spectrum of modulated wave

When a modulated wave is applied to a single-tuning circuit, the amount of elimination of high-frequency components can be calculated by the following,

$$\frac{\text{Current of detuning}}{\text{Current of tuning}} = \frac{1}{\sqrt{(1+\delta)^2 + [Q\delta(\frac{2+\delta}{1+\delta})]^2}} \quad (4)$$

where

$$\delta = \frac{\text{Frequency deviation from tuning frequency}}{\text{Tuning frequency}}$$

$Q =$ Effective Q of tuning circuit

Reference of the f. (4)

Terman: Electronic and Radio Engineering, 4th Edition

Presume the modulation frequency as 10 kHz, then the calculated effective value of a tuning circuit of which the side-band will descease in terms of 1 dB, 0.5 dB and 0.2 dB are shown in Table 1.

Table 1 Relation between output decreased and Q at modulation frequency of 10 kHz

		Carrier frequency		
		540 kHz	1000 kHz	1500 kHz
		$\delta = 0.0185$	0.01	0.0067
Output decrease	1 dB	Q = 13	25	37
	0.5 dB	Q = 8	16	25
	0.2 dB	Q = 3	8	14

The characteristics of the broadcasting transmitter by BTS standard is specified at (+)1 (-)1.5 dB at 10,000Hz. Also, since the frequency allocation of the medium wave is made with intervals of 10 kc, sidebands seems to interfere virtually the adjacent channel at high modulation frequencies, but actually it does not occur. The reason is that the occupied frequency band-width is limited to 15 kHz by the Radio Regulations. The occupied frequency band-width means a frequency band-width where mean radiated power in that frequency band occupies 99% of the mean-power of total radiated waves. In other words the mean-power of each side band caused by the modulation frequency exceeding 7,500 Hz is required to be 0.5% or less than the total mean-power. However, this condition is almost always satisfied in actual broadcasting programs.

Here, some explanation will be given with regard to the frequency modulation (FM) below.

Presume high-frequency current is expressed by the following.

$$i = I_o \sin \phi(t) \qquad (5)$$

where $\phi(t)$ is an angular displacement at time t.

Instantaneous angular velocity ω_i and instantaneous frequency f_i are given by the following

$$\omega_i = 2\pi f_i = \frac{d\phi(t)}{dt} \qquad (6)$$

Then presume that the frequency of high-frequency current is deviated to maximum frequency-deviation Δf by modulation frequency f_m. Then the instantaneous angular velocity will be given by the following expression.

$$\omega_i = 2\pi (f_o + \Delta f \cos 2\pi f_m t) \qquad (7)$$

where f_o is a frequency without modulation

By combining (6) and (7)

$$\begin{aligned} \phi(t) &= \int 2\pi(f_o + \Delta f \cos 2\pi f_m t)dt \\ &= 2\pi f_o t + \frac{\Delta f}{f_m} \sin 2\pi f_m t + \theta \\ &= \omega_o t + m_f \sin \omega_m t + \theta \qquad (8) \end{aligned}$$

where $\omega_o = 2\pi f_o$, $\omega_m = 2\pi f_m$, $m_f = \frac{\Delta f}{f_m}$ = modulation index,

θ = integral constant (angle desplacement at $t = 0$)

By neglecting θ in (8) for simplification, and put into (5)

$$i = I_o \sin [\omega_o t + m_f \sin \omega_m t] \qquad (9)$$

Then (9) can be expanded with Bessel functions.

$$\begin{aligned} i = I_o[&J_o(m_f)\sin \omega_o t \\ &+ J_1(m_f)(\sin(\omega_o + \omega_m)t - \sin(\omega_o - \omega_m)t) \\ &+ J_2(m_f)(\sin(\omega_o + 2\omega_m)t + \sin(\omega_o - 2\omega_m)t) \\ &+ J_3(m_f)(\sin(\omega_o + 3\omega_m)t - \sin(\omega_o - 3\omega_m)t) \\ &+ \ldots\ldots\ldots\ldots] \end{aligned} \qquad (10)$$

where $J_n(m_f)$ is the n-th order Bessel function of first kind with argument m_f.

The values are graphically indicated in Fig. 5.

As understood from (10), the carrier wave component of frequency modulation is not constant, but changes in accordance with the modulation index, and in some cases, it happens to disappear. With regard to side-bands, one pair appears above and below the main carrier, same as in the case of AM. But even when modulated with a single frequency, an infinite number of side-bands will occur in accordance with the intervals of modulation frequency. Thus, the frequency spectrum theoretically has an infinite spreading, but when the modulation index is under 0.5, the side-bands beyond the second order are small and can be neglected as shown in Fig. 5, so the frequency band-width can be deemed almost identical to that of AM.

When the frequency deviation is constant, the modulation index is inversely proportional to the modulation frequency as seen in (8).

When the modulation frequency is high, the modulation index will not be so large, so the primary side-bands will be located considerably apart from the center frequency, as described above, but the higher order side-bands can be neglected. On the contrary, when the modulation frequency is low, the modulation index will be large and side-bands up to higher orders must be taken into account, but as the frequency intervals are small, even if the higher order side-bands are taken into account, the necessary frequency speading will not be so large. In practical, the main side-bands of frequency modulation to be considered are those of "frequency deviation + modulation frequency" apart from the center frequency.

In case of phase modulation, (5) can be converted to the following form.

$$i = I_o \sin(\omega_o t + m_p \sin \omega_m t) \qquad (11)$$

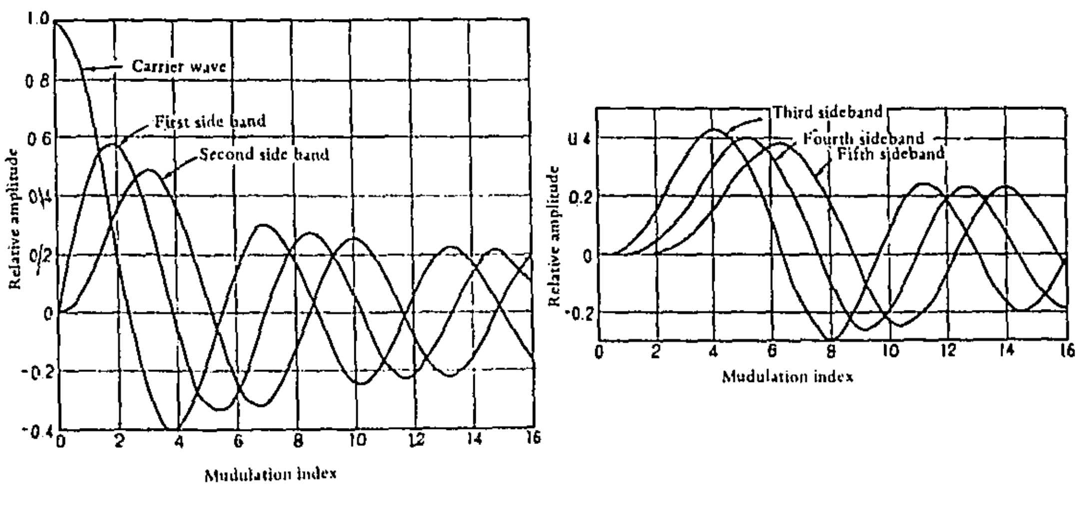


Fig. 5 Relative amplitude of each frequency component in frequency modulation (amplitude of carrier wave without modulation is taken as 1)

where m_p = Modulation index of phase modulation
= Maximum displacement of phase angle

By comparing the above expression with (9), the conversion is performed only by the replacement of the modulation index m_f with m_p, so that the condition of sidebands generating is completely identical to the case of frequency modulation.

By using formula (6), the instantaneous angular velocity of the phase modulated wave is,

$$\omega_i = \frac{d\phi(t)}{dt}$$

$$= \frac{d}{dt}(\omega_o t + m_p \sin \omega_m t)$$

$$= \omega_o + \omega_m m_p \cos \omega_m t$$

By comparing this with (7),

$$2\pi\Delta f = \omega_m m_p$$

or

$$\Delta f = f_m m_p \qquad (12)$$

is obtained.

The meaning of this is that phase-modulation is equivalent to frequency modulation and the frequency deviation is proportional to the modulation frequency. Therefore, it phase modulation is applied to an audio circuit by adding a network of frequency responses inversely proportional to the frequency, then a correct frequency modulation with a constant frequency deviation will be obtained.

Fig. 3 (c) indicates the vectorial relationship between carrier and first sideband of formula (10), representing frequency modulated waves. This vector is similar to the vector (b) of amplitude modulation. However, in this case as the sign of the lower sideband L is opposite from that of the upper sideband U, the composition of upper and lower sidebands becomes a vector with varying amplitude in a direction rectangular to the carrier wave. The composition of this rectangular direction vector and the carrier wave C is phase-deviated from the carrier wave of non modulation by the amount of θ.

Since the phase angle θ is changed by the modulation frequency, it conclusively indicats that frequency modulation is equivalent to phase modulation. It would be also understood that the amplitude of modulated wave is constant irrespective of the modulation, if the angle θ is small.

3. Operation of vacuum tube

3.1 Fundamental items

In Fig. 6, E_c is the grid bias voltage, E_b is the DC plate voltage. Z_L is the load impedance and in case of audio amplifiers, it is a non-tunned circuit and in case of high-frequency amplifiers, it is regarded as pure resistance.

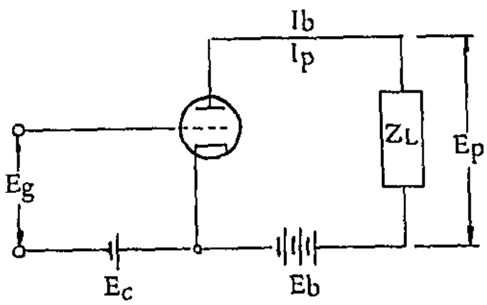


Fig. 6 Fundamental vacuum tube amplifier circuit

When a sinusoidal wave of amplitude E_g is provided to the grid of the amplifier, assume that the amplitude of the sinusoidal wave voltage appearing at both terminal of the load is E_p. Assume that the DC component of plate current is I_b, and the amplitude of the sinusoidal wave of the fundamental wave is I_p.

In considering the circuit operation of this circuit above, the voltage utilization-factor and current utilization factor can be presumed as follows.

$$\text{Voltage utilization factor } \xi_\gamma = \frac{E_p}{E_b} \tag{13}$$

$$\text{Current utilization factor } \xi = \frac{I_p}{I_b} \tag{14}$$

Presuming that the load impedance is pure resistance, and as E_p and I_p are in phase the output P_o applied to the load is given by,

$$P_o = \frac{E_p}{\sqrt{2}} \times \frac{I_p}{\sqrt{2}} = \frac{E_p\, I_p}{2} \qquad (15)$$

On the other hand, the input P_i applied from the plate source to the plate circuit is given by

$$P_i = E_b\, I_b$$

Then, the ratio of the amount of power converted to the output from the power supplied to the plate source, is called plate-efficiency and can be simply expressed by the following formula.

$$\text{Plate efficiency} \quad \eta_p = \frac{P_o}{P_i} = \frac{1}{2}\left(\frac{E_p}{E_b} \times \frac{I_p}{I_b}\right) \qquad (16)$$

$$= \frac{1}{2}\,(\text{Voltage utilization factor} \times \text{Current utilization factor})$$

The residual after subtraction of the output from the plate input, i.e.

$$P_p = P_i - P_o$$

is the plate loss, which makes the plate temperature rise. If no AC voltage exists at the grid, the plate input will totally turn out as plate loss. However, it can be considered that by applying AC voltage, a part of the plate input will be converted to the output.

When a sinusoidal wave is applied to the grid of a circuit as shown in Fig. 6, the plate current conducts throughout one cycle of the input sinusoidal wave or otherwise, only one part of the cycle.

In this case, the range of current flow is normally expressed by an electrical angle as shown in Fig. 7, where the period of plate current conduction in one cycle is expressed by an electrical angle and is called the conduction-angle, and one half of this is called the operational-angle.

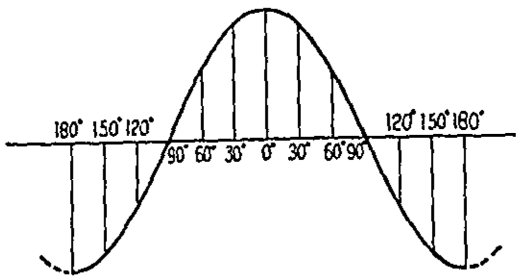


Fig. 7 Electrical angle of voltage and current in vacuum tube

In general, plate current I_b of a triode is given by the following formula when the grid voltage is negative.

$$I_b = K_1 \left(E_c + \frac{E_b}{\mu}\right)^{3/2} \tag{17}$$

Where K_1 is a constant determined by the construction and dimension of the vacuum tube, and μ indicates the amplification factor. When $(E_c + E_b/\mu)$ is negative, the plate current becomes zero. Therefore, the value given by the following formula is the cut-off bias, since $(E_c + E_b/\mu) = 0$ is the boundary where the plate current becomes just zero.

Cut-off bias = E_b/μ

When the gird voltage is positive, grid current I_c flows, so that (17) must be corrected as follows.

$$\text{Cathode current} = I_b + I_c = K_1(E_c + E_b/\mu)^{3/2} \tag{18}$$

The ratio of this cathode current distribution to the grid and the plate is determined according to the construction of the vacuum tube and the voltage of each electrode. To say qualitatively and simply, the lower the plate voltage is, the higher the grid current component becomes.

In case of tetrode, pentode or beam tube, (18) is corrected as follows.

$$\text{Cathode current} = I_b + I_{c2} + I_{c1}$$
$$= K_2\ (E_{c1} + E_{c2}/\mu_2)^{3/2} \qquad (19)$$

Where K_2 is a constant, I_{c1} and E_{c1} are voltage and current of the control grid, I_{c2} and E_{c2} are voltage and current of the screen grid respectively and μ_2 is an amplification factor of the screen grid. In this case, the amplification factor of the control grid μ_1 is very large, so that the influence of the plate voltage to the plate current is practically negligible.

Therefore, the cut-off bias is given by the following formula,

$$\text{Cut-off bias} = -\ E_{c2}/\mu_2$$

3.2 Classification of operation of vacuum tube

The operation of vacuum tubes are divided into three categories according to the operation angle of plate current such as class A, class B and class C.

Class A relates to an operational angle of 180° where the plate current conducts throughout the whole cycle of the input sinusoidal wave. Class B relates to an operation angle of 90°, where the plate current conductes only during half period of the input sinusoidal wave. Class C relates to an operation angle less than 90° where the plate current conducts only during a period less than half cycle of the input sinusoidal wave. Intermediate operation between class A and B is sometimes called class AB where the tube is considered to operate as class A with regard to small input, and at large input, it operates closely to class B.

In some cases, suffix 1 or 2 is added to the operational symbol of the vacuum tubes, for instance, class A_1, class B_1, class A_2, class B_2, etc. In these cases, suffix 1 indicates a system where no grid current conducts throughout a whole cycle of the input sinusoidal eave and suffix 2 indicates a system where the grid current conducts. In case of class C, the grid current is usually conducting, so there is no particular discrimination.

3.3 Class A operation

As shown in Fig. 8, the plate current conducts throughout a whole cycle of the input signal. If the vacuum tube characteristic is linear, the waveform of AC component in the plate current is identical with that of the input signal and, in particular, this holds even if the input signal waveform is not sinusoidal. Therefore, class A operation is most suitable to amplify a complicated waveform, e.g. audio current, etc. without distortion.

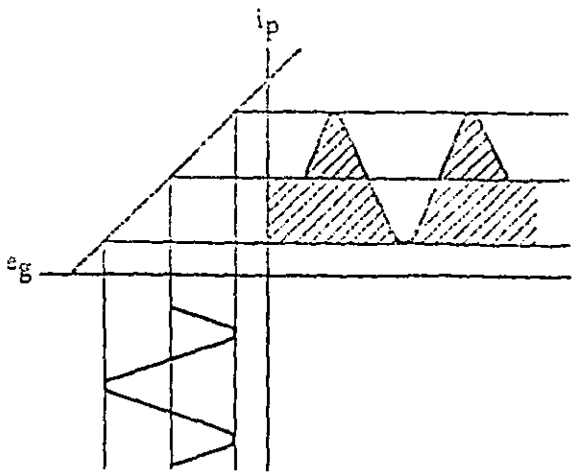


Fig. 8 Class A operation

The current utilization factor of class A operation differs in accordance with the amplitude of the input signal, but will not exceed 1 at the maximum. Therefore, the plate efficiency of class A operation is low, and it is not applicable to high-power amplification but applicable to low power amplification.

Class A operation can be divided theoretically into class A_1 and A_2, but as the power to deal with is low as mentioned above, and the necessity of reducing distortion, class A is generally referred as class A_1.

3.4 Class B operation

In class B operation, the plate current conducts only during a half cycle of the input signal as shown in Fig. 9.

Therefore, the AC component of the plate current includes considerable amount of even order harmonics in addition to the fundamental component of the input signal frequency.

When two vacuum tubes are connected in pushpull and operated at class B, each tube will work at positive and negative half cycles, thus resulting in a equivalent working condition of one tube operated at class A, and the amplification output becomes less distorted. In this case, even order harmonics generated in each tube are cancelled when conducting through

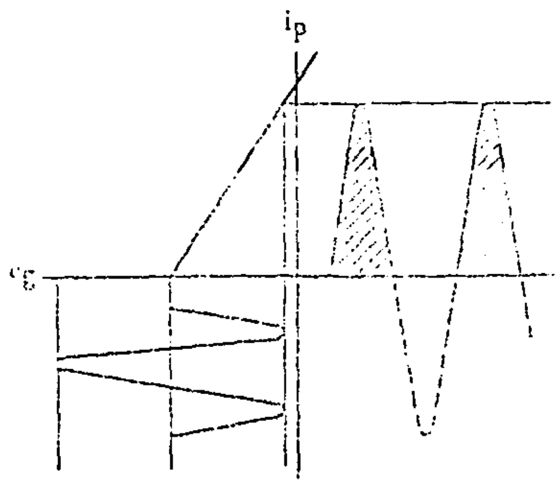


Fig. 9 Class B operation

the primary winding of the pushpull output transformer, so that no other component appears on the secondary side, except components of the same frequency as the input signal waveform. Therefore, class B pushpull circuit can be used for the amplification of audio current.

Class B operation is also applicable to high frequency amplification. In particular, class B operation is necessary for amplification of 100% modulated radio frequency wave. In this case, plate circuit only takes out fundamental wave through a tuning circuit and harmonics are bypassed. Therefore, even distortion exists in the plate current waveform, it does not harm the performance. In addition, the output amplitude of the fundamental wave is proportional to that of the input signal, therefore the envelope of the modulated wave is linearly amplified without distortion.

Current utilization factor at class B operation is equal to $\pi/2$ when calculated with the presumption that the plate current is sinusoidal half wave. Although this is inferior to the class C operation described in the following it is far superior to class A operation. Therefore, the plate efficiency of the power amplifier of this system is considerably excellent, so that it is suitable for a high power audio amplification and modulated radio frequency wave mplification.

Class B operation is divided into class B_1 and B_2, the former is applicable to an audio amplifier having relatively small output. Class B_2 amplifier requires special precaution in the design of input circuit, however, since it can handle the largest output of the given vacuum tubes. Therefore an audio amplifier of class B_2 pushpull is widely applicable to the modulator of the broadcasting transmitters. Almost all class B linear amplifiers for the modulated radio frequency wave employs class B_2.

3.5 Class C operation

In class C amplification, the operational angle of plate current is less than 90° and the plate

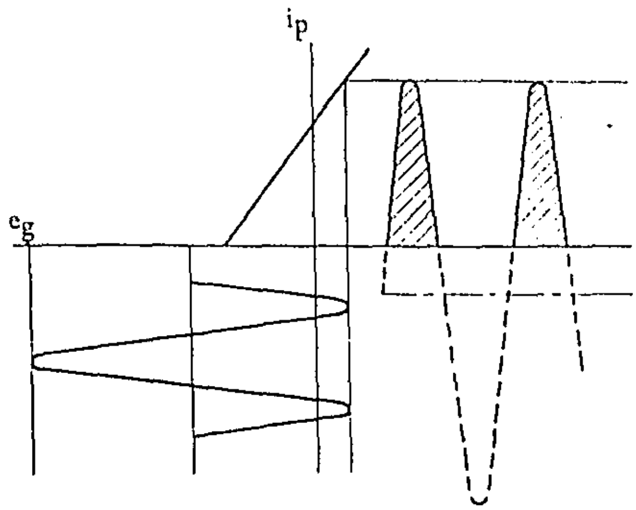


Fig. 10 Class C operation

current conducts in a narrower range than a half-cycle sinusoidal wave as shown in Fig. 10. In this case, in addition to the fundamental wave of frequencies equivalent to the input wave frequency, a large amount of even and odd order harmonics are included in the plate current wave form.

When pushpull connection is incorporated with class C operation, even order harmonics will cancel each other. However, the odd order harmonics will be doubled and forced out at the output. Therefore, this system is not applicable to audio amplification.

Class C operation is exclusively applicable to power amplification of high frequency wave. In this case, the fundamental components in the plate current are only taken out by a tuning circuit and the harmonic components are bypassed and removed.

Class C high frequency amplification is applicable to the modulated radio frequency amplifiers, where the most practical methods are the plate modulation system and the grid modulation system. In case of multielectrode tubes, there are various systems, e.g. screen grid modulation, screen grid and plate simultaneous modulation, suppressor grid modulation, etc., which are applicable only to, low power or auxiliary purposes.

Class C operation is rarely applied to power amplification of modulated radio frequency wave, where rectified negative feedback must be incorporated to suppress occurence of distortion due to non-linearity of amplification characteristics.

Distinctive features of class C operation are high plate current-utilization factor with high power-efficiency. Presuming that the plate current wave form is a part of a sinusoidal wave, the calculated current utilization factor will be within 1.57 to 2, and in practical, it is considered to be about 1.8.

4. Audio amplification circuit

4.1 Outline

In this article, an audio amplification circuit means the overall modulation system containing an amplifier and modulator, in a construction of a broadcasting transmitter as shown in Fig. 11.

The audio amplifier cation circuit of a braodcasting transmitter is classfied into the voltage amplification circuit and the power amplification circuit in correspondence with its operational functions. In a normal braodcasting transmitter, audio input is applied to the power amplifier circuit called a modulator, after voltage amplification one stage or more. In case of the modulator of class B_2 pushpull, a power amplifier circuit called sub-modulator is norma-ly provided before it.

Voltage amplifier circuit is mainly intended to obtain a large voltage amplification degree and it employs a high load impedance in the plate circuit. On the contrary, power amplification circuit is intended to obtain a large power output, so that the load impedance is usually not so large. Of course, voltage amplification can supply the amplified power to the load and power amplification often accompanies voltage amplification. The above is classified according to the main purpose of the amplifier.

4.2 Voltage amplification circuit

The following features are necessary to voltage amplification circuits.

(1) High amplification degree

(2) Large output voltage

(3) Less distortion

(4) Good frequency response with excellent phase characteristics

(5) Small input capacity with high input resistance

(6) Simple circuit and low cost of parts

In a broadcasting transmitter the condition of (4) and (5) are important. (Because negative feedback is almost always incorporated in the broadcasting equipment, so that the frequency response is required as flat as possible and a good phase characteristics is necessary for stable negative feedback.) Further, the frequency response and the phase characteristics have a close relationship, where the better the frequency response is, the better the phase characteristics becomes. In addition, the input capacity has an influence upon the high frequency response, and it relates to the grid to plate (GP) capacity and the amplification degree of the tube. In a triode, the GP capacity is large, so that, if the amplification degree is put large, the input capacity becomes extremely large. But in case of multi-electrode tubes, this capacity is very small.

The voltage amplifier circuit can generally be classified into the transformer coupled circuit and the resistance coupled circuit as shown in Fig. 11. There is a choke coupled circuit as a modification of the transformer coupled circuit.

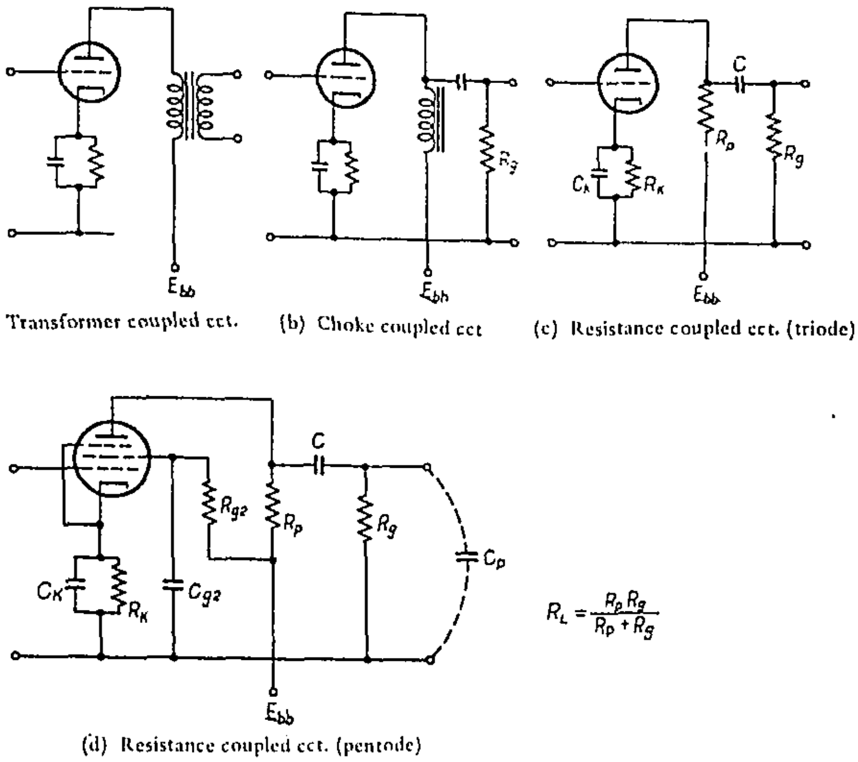


Fig. 11 Types of voltage amplifier circuit

In case of transformer coupled circuits, the maximum output voltage becomes large. However, the amplification degree is not so large, and in addition, it has shortcoming of poor frequency response and phase characteristics. Also, the cost of transformers of good performance is high. On the contrary, resistance coupled circuit has extremely excellent frequency response and phase characteristics and the cost of parts is low, so that if designed properly, it affords very high amplification degree as well as considerably high output voltage. Therefore, for voltage amplification, a resistance coupled circuit is exclusively used.

For the resistance coupled circuit, there are two methods, such as using triodes and pentodes. Table 2 indicates the comparison of maximum output voltage and amplification degree, based on RCA Tube Manual. According to the table, pentodes are excellent with regard to amplification degree. Moreover, as input capacity is very small, pentodes are suitable for resistance coupled circuits.

Table 2 Comparison of vacuum tubes regarding resistance coupled system

Kinds	Medium-μ triode	High-μ triode	Pentode
Type	12AU7	12AT7	6AU6
Amplification factor	12	60	230
Maximum output voltage	82	78	108
Circuit	Fig. 11 (c)		Fig. 11 (d)
Circuit condition	E_{bb}=300V, R_p=0.22MΩ, R_g=0.47MΩ		

As well know, the amplification degree of medium frequency range of a resistance coupled circuit can be calculated by the following formula.

$$\text{Amplification degree } A = \frac{\mu R_L}{\gamma_p + R_L} \qquad (20)$$

where μ = Amplification factor

γ_p = Internal resistance of vacuum tube

$$R_L = \text{Load resistance} = \frac{1}{\frac{1}{R_p} + \frac{1}{R_g}}$$

Since $\gamma_p \gg R_L$ prooves in a multielectrode tube, neglect R_L in comparison with γ_p in (20), then the following expression will be obtained.

$$\frac{\text{Amplification degree of}}{\text{multielectrode tube A}} = \frac{\mu}{\gamma_p} R_L = g_m R_L \qquad (21)$$

where g_m = trans-conductance

If only the characteristic constants of the vacuum tube are known, the amplification degree can be calculated easily as shown above. As a matter of practice, however, plate voltage and screen grid voltage are considerably low because of the voltage drop due to series resistance, so that the calculation can not be carried out unless the characteristic curve is known precisely in that range. For designing without using characteristic curves, it is convenient to refer to the design values of resistance coupled amplifier in the RCA Tube Manual.

However, be careful to the following items when using values published in the Tube Manual. First, maximum output voltage relates to the value of distortion factor of 5%, so that, in case of a broadcasting transmitter, some reduction must be incorporated. Second, values of coupling condenser and each bypass condenser are selected such that the low frequency response decreases by 3 dB at 100 Hertz. This is not sufficient for the broadcasting purposes. It is preferable to select the bypass condenser by 10 times of the specified value and the coupling condenser of more than by 10 times as far as possible.

General precautions with regard to the design of the resistance coupled amplifier are described below.

(1) It is more preferable to supply grid bias voltage and screen grid voltage through series resistances R_k and R_{g2} of the cathode circuit and the screen circuit respectively, than to supply those from fixed voltage source, e.g. breeder, etc. This is because the fluctuation of the plate current and the screen current caused by the variation of plate supply voltage and vacuum tube replacement is automatically suppressed by the

voltage drop across the resistance.

(2) The value of load resistance R_L is determined in accordance with high frequency response. In Fig. 11 (d), if the capacity C_p indicated by the broken line is presumed to be the stray capacity of the plate circuit (including input capacities of following stages), the amplification degree will decreases by 3 dB at the frequency f_c indicated in the following formula.

$$R_L = \frac{1}{2\pi f_c C_p}$$

Therefore,

$$f_c = \frac{1}{2\pi R_L C_p}$$

Thus, the high frequency response is determined by the product of $R_L C_p$. For instance, if we presume R_L = 0.2 MΩ, C_p = 25 pF, 3 dB drop occurs at f_c = 32 kc. In case of broadcasting equipments, maximum practical limit of R_p is approx. 100 kΩ in consideration of the stability of the negative feedback.

(3) It is preferable to apply a large grid leak R_g as possible to obtain a large amplification degree. Practically, the applicable value is about 5 times R_p. If the grid leak is excessively large, and when the plate current of the vacuum tube is large, positive ions separated from tube residual gas will adhere to the grid of the tube, thus resulting in loss of grid control function. Therefore, there is a limitation in this regard.

4.3 Class A power amplifier

In general, the power amplifiers must conform to the following conditions.

(1) Low plate source voltage with large output power
(2) Low excitation voltage with large output power
(3) High power efficiency
(4) Less distortion
(5) Good frequency characteristics and phase characteristics
(6) Small input capacity and high input resistance

Since the broadcasting transmitter consumes large power, power efficiency is an important problem with regard to the operation cost. In particular, this requirement is critical in case of the high power transmitters. It is preferable however to employ low plate supply voltage in consideration of maintenance, so that in large power transmitters, it is important to cut down the high plate voltage as far as possible.

In audio power amplification circuits, both class A and sclass B are employed, the former will be described first.

Class A power amplification has an advantage of simplicity of circuit configuration and less distortion. However, since the plate efficiency is low, it can not be applicable to a large output. Class A power amplification is exclusively applicable to a sub-modulator or i.e. a driver, in the preceding stage of class B modulators.

If comparison is made between a triode and a multi-electrode tube, with regard to the applicability to class A power amplifiers, the latter can afford larger output, better plate efficiency with less excitation voltage, under the same plate supply voltage. The multi-electrode tube has larger distortion than the triode and this is particularly prominent in case that the load changes according to the signal level, e.g. the case of a sub-modulator tube. But this distortion can easily be suppressed by applying negative feedback. As for drivers of class B_2 modulators, a cathode follower of beam tetrodes in pushpull connection are generally used.

It is convenient for the analysis of class A power amplifier operation to utilize plate voltage/plate current characteristics, i.e. E_b-I_p characteristic curve, as shown in Figs. 12 and 13. In these figures, the point Q can be determined if the plate voltage E_b (DC voltage between plate and cathode obtained by subtracting the voltage drop due to output transformer, cathode bias, etc. from the plate supply voltage E_{bb}) and the grid bias E_{co} are given. Point Q indicates the plate voltage E_b and plate current I_b in case of no input signal, and this is the basic point of operation. If the plate current increases beyond this point, the operating point moves to the upper left side, due to the voltage drop in the load. If the plate current decreases, the plate voltage raises due to the decrease in voltage drop, so that the operating point moves to the lower right. In case of a audio amplifier, the load can be normally deemed as pure resistance, therefore, the locus of these operating points form a straight line passing the base point Q. This line is called a working line or a load line. The slope of this line is given by

$$\frac{\text{Increase of plate voltage}}{\text{Decrease of plate current}} = \text{Load resistance } R_L$$

When a sinusoidal excitation voltage of amplitude E_g is applied to the grid, the grid voltage varies between "$E_{co} \pm E_g$", so that, in accordance with this maximum and minimum value, the following values can be obtained on the operating line, i.e. plate current maximum I_{max}

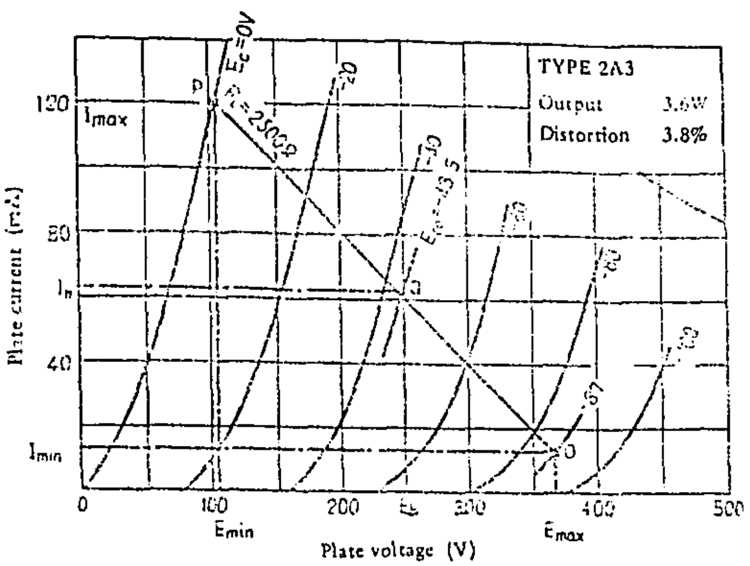


Fig. 12 Example of class A operation of triode

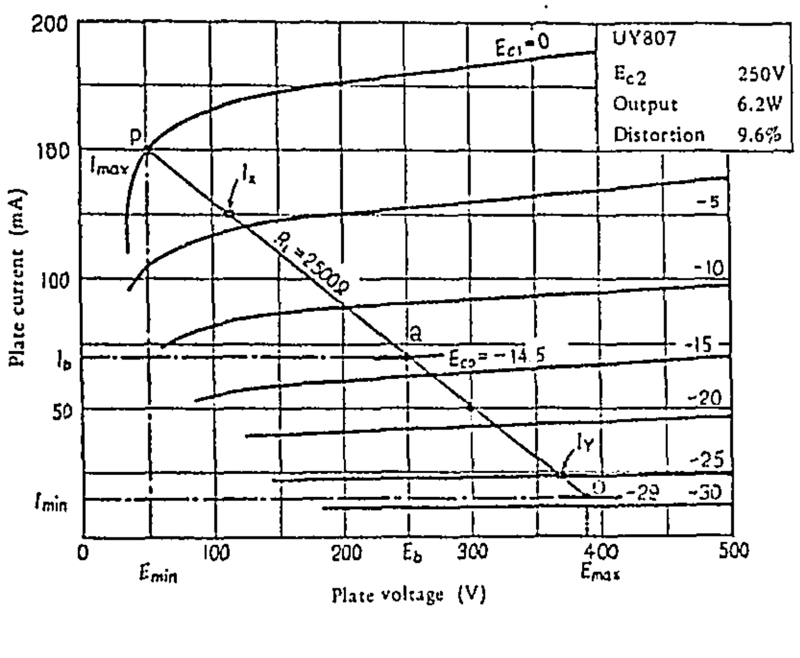


Fig. 13 Example of class A operation of beam tetrode

and minimum I_{min}, and plate voltage minimum E_{min} and maximum E_{max}. Correspondingly the operating line is indicated by PO.

In a class A amplifier of triode, the output and distortion factor can be obtained by using these values, the in the following formula.

$$\text{Output } P_o = \frac{(I_{max} - I_{min}) \times (E_{max} - E_{min})}{8} \tag{22}$$

$$\text{Distortion factor} = \frac{\dfrac{I_{max} + I_{min}}{2} - I_b}{I_{max} - I_{min}} \times 100\% \tag{23}$$

In order to reduce the distortion factor, grid voltage must not exceed zero to avoid the effect of grid current flow. Also, if I_{min} is drived to zero, the distortion increases because of the bend of the characteristic curve in the neighborhood. Therefore, load resistance value is varied within the range of these two conditions, and the output and distortion are calculated to determine the load resistance R_L which makes the output maximum, in correspondence with the distortion factor.

Description will be made with regard to the meaning of the maximum output. When we take out power from a circuit having a constant electromotive force and a constant internal resistance, we know that the maximum power is obtainable on condition that the load resistance is equal to the internal resistance, namely on the internal resistance, namely on the condition of impedance matching. However, the maximum output of a power amplifier has a different meaning. In other words, it expresses the ability of power obtainable of a given vacuum tube, without distortion. The grid excitation voltage becomes larger than the case of impedance matching, although, the extention of output power causes better economy of facility cost and consumption of power.

Les us calculate the condition of maximum output on the presumption of an ideal triode, where e_p - i_p characteristic is linear and γ_p is constant, with a given plate supply voltage E_b.

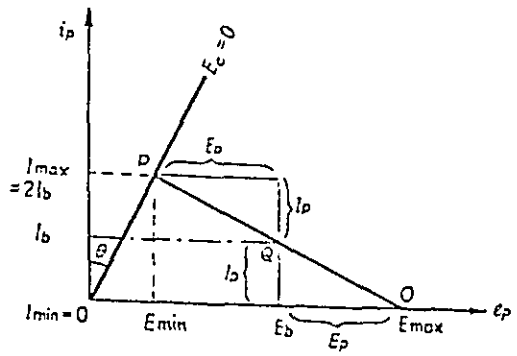


Fig. 14 Operational characteristics of idealized triode

Referring to Fig. 14, determine a base point of operation Q on the vertical line passing E_b in accordance with a suitable I_b. Then determine an operational line OP, such that both ends agree with a line of E_c = Q and a line of I_{min} = 0 respectively, and the point Q positioned at the center. As is obvious from the drawing,

$$I_p = I_b$$

$$E_p = E_b - E_{min}$$

$$= E_b - 2I_b\gamma_p \quad \text{[Because } \tan\theta = \frac{de_p}{di_p} = \text{Internal resistance } \gamma_p\text{]}$$

$$\text{Output } P_o = \frac{E_pI_p}{2}$$

$$= I_b(E_b - 2I_b\gamma_p) \times \frac{1}{2} = (I_bE_b - 2I_b^2\gamma_p) \times \frac{1}{2}$$

In order to obtain the condition of maximum output, differenciate P_o with respect to I_b and put it equal to zero.

$$\frac{dP_o}{dI_b} = (E_b - 4I_b\gamma_p) \times \frac{1}{2} = 0$$

From this,

$$I_b = \frac{E_b}{4\gamma_p}$$

And

$$R_L = \frac{E_p}{I_p} = \frac{E_b - 2I_b\gamma_p}{I_b} = \frac{4I_b\gamma_p - 2I_b\gamma_p}{I_b}$$

As described above, first determine I_b, then take the load resistance two times the internal resistance. However, I_b can not exceed the permissible plate loss. The method described above is related to the ideal case. Practically the characteristic curve must be used for the calculation. However, the result of this ideal case will play a role of reference.

In case of multielectrode tubes, graphical analysis is applicable with $e_p - i_p$ characteristics similarly to the case of triodes. As obvious from Fig. 13, upper end point P of the load line must be selected at the shoulder of the zero bias line. In multielectrode tubes, the line intervals are not uniform with regard to the same amount of change of bias voltage, and there will be much distortion. Therefore, in order to calculate the output and the distortion, use the current I_X, I_Y of intermediate points on the operating line (refer to Fig. 13) in addition to I_{max}, I_{min}, I_b, E_{max} and E_{min}, and apply it to the following equation.

$$\text{Output } P_o = \frac{[I_{max} - I_{min} + 1.41(I_X - I_Y)]^2 R_L}{32} \qquad (24)$$

$$\text{Distortion factor of second harmonic} = \frac{I_{max} + I_{min} - 2I_b}{I_{max} - I_{min} + 1.41\ (I_X - I_Y)} \times 100\%$$

$$\text{Distortion factor of third harmonic} = \frac{I_{max} - I_{min} - 1.41\ (I_X - I_Y)}{I_{max} - I_{min} + 1.41\ (I_X - I_Y)} \times 100\%$$

$$\text{Total distortion factor} = \sqrt{\left(\begin{array}{l}\text{Distortion}\\ \text{factor of}\\ \text{second harmonic}\end{array}\right)^2 + \left(\begin{array}{l}\text{Distortion}\\ \text{factor of}\\ \text{third harmonic}\end{array}\right)^2} \qquad (25)$$

Where I_X and I_Y will be the plate current which correspond to points apart from the base point Q, by the amount of grid voltage of 0.707 times the maximum exciting voltage (this is equal to the base point bias voltage) in positive and negative directions on the operating line. In other words, those are the plate currents corresponding to the bias voltage of 0.293 E_{co} and 1.707 E_{co}. In case of multielectrode tubes, optimum load resistance does not have correspondency with the internal resistance, contrary to the case of triodes.

Next, the large amplitude operation of the amplifier will be examined in case of a load including the reactance. The load impedance of an audio. Amplifier is considered as pure reactance in the medium frequency range. At the low frequency range, the reactance caused by the inductance of the output transformer is inserted parallel to the load resistance. At the high frequency range, the reactance caused by the stray capacitance of the plate circuit is inserted parallel to the load. As a result, amplification degree decreases together with the decrease of the load impedance, as well as with the shift of the output voltage phase. More explicitly, referring to Fig. 15, if the load is pure resistance of 10 kΩ and the output voltage is 200V, AC component of plate current is 20 mA and the operating line becomes the line ab. When the load is pure reactance of 10 kΩ, both output voltage and AC component of plate current are identical to the foregoing, i.e. 200V and 20 mA, respectively. However, since the voltage has a leading phase angle of 90° with respect to the current, so that the operating line becomes an ellipse expressed by cdefc. (Although a circle is drawn since the same scale was employed with respect to 20 mA and 200V, an ellipse is generally formed, having vertical or horizontal long axis.)

Next, in case that a resistance of 10 kΩ and a reactance of 10 kΩ are connected parallel as a load and the output voltage is 200V, plate current consists of vectorial summation of the resistance current and the reactance current. Therefore, referring to Fig. 15, the ellipse agcbhea becomes the operating line, which is composed by the line ab and the ellipse cdefc, then the plate current becomes 28 mA. This is easily understood because the load impedance

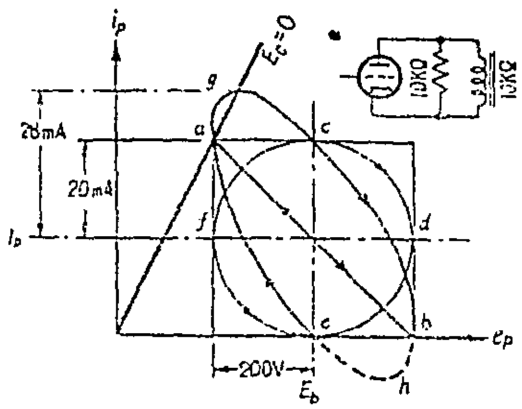


Fig. 15 Example of amplifier operation on reactance load

decreases to 1/ 2 so that, to keep the output voltage at the same value as before, current must be increased to 2 times. In this case, however, operating line agcbhea enters the range of positive grid voltage. Also, it enters the range of negative plate current. Since the plate current can not flow inversely, however, lower side is chopped and generates the output waveform distortion. If to be operated in a distortionless range, output decrease must be inevitable. The same results are obtained whether the reactance is inductive or capacitive, except for the reversed direction of the operating line sevolution.

4.4 Class B power amplifier

The feature of class B power amplifier is that the plate efficiency is higher than that of class A. It can be connected to a B class pushpull circuit as an audio amplifier, to obtain high efficiency amplification with less distortion.

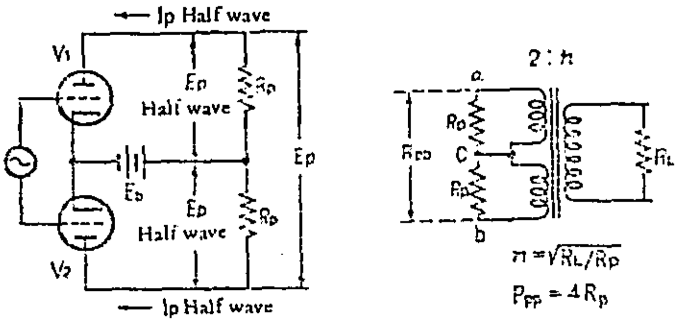
First, let us examine the plate-load distribution in class B pushpull amplification.

Referring to Fig. 16 (a), the load connected between both plates becomes $2R_p$ on the presumption that the load of one vacuum tube is R_p. In consideration of the output transformer, however, one half of the primary winding, i.e. ac and bc must match R_p respectively, so that the impedance between a and b of the primary winding is proportional to the square of the number of turns, i.e. $4R_p$. In other words, when V_1 is operating at a positive half cycle of the input signal, V_2 has no plate current, so the winding bc becomes open and the winding ac only works as impedance R_p. Then at a negative half cycle, the winding bc only works as impeda impedance R_p to the contrary and the winding ac becomes open. Namely, ac and bc work as impedance R_p alternately.

In class B amplification, circuit configuration consists of pushpull but the operation of the vacuum tube is only of push of each other without the pull operation.

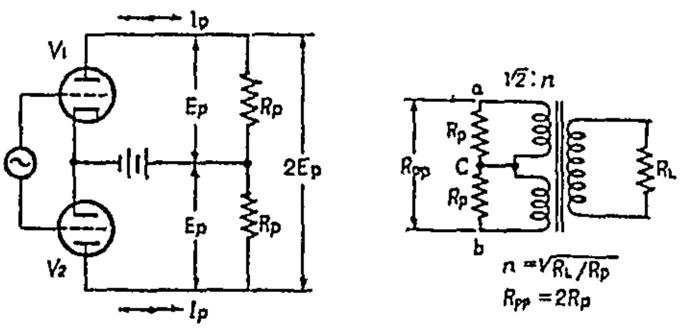
Then we shall compare this with class A pushpull circuit as shown in Fig. 17.

For class A, V_1 and V_2 are completely in pushpull operation.



(a) Class B pushpull circuit (b) Output transformer

Fig. 16 Load distribution in class B pushpull circuit



(a) Class A pushpull circuit (b) Output transformer

Fig. 17 Load distribution in class A pushpull circuit

and it is obvious that the impedance between terminals ab of the primary side of the output transformer must match the impedance $2R_p$.

Referring to vacuum tube catalogs, it is a normal practice to indicate the plate/plate load impedance of pushpull circuit operation as R_{pp}. Since this means the impedance between both plate terminals of the output transformer, load resistance of one vacuum tube becomes $R_{pp}/4$ for class B and $R_{pp}/2$ for class A.

In class B pushpull amplification, when the grid is exited by a sinusoidal wave, the plate current is sinusoidal half-wave, so the DC plate current flow is the mean value thereof, i.e. I_{max}/π per one vacuum tube and 2 I_{max}/π per two vacuum tubes. This current always varies according to the amplitude of the input signal, so that the plate power supply must have excellent voltage regulation property.

In class B_1 pushpull circuit, grid current does not conduct, therefore a normal voltage amplifier circuit can be employed in the preceding stage, so resulting in a simple configuration of circuit. When the triode and the pentode are compared as for applicable vacuum tube, the latter is superior with respect to both plate efficiency and output. For instance the plate efficiency of the pentode can be raised up close to 60% at maximum. Any way, in class B_1 amplifiers the circuit configuration of the preceding stage is simple, so it is suitable for modulator of low power broadcasting transmitters.

Fig. 18 indicates an example of class B_1 operation of 4B85 beam tetrode tube. Since its plate to plate load impedance R_{pp} is 2,200 Ω, operating line OP can be drawn on the condition of 550Ω load resistance per vacuum tube. In this case, point Q determined by the plate voltage and the grid bias has not any relation with the operating line if pushpull is involved, but it only affords plate current during no signal condition. When the signal is small, however, it approaches class A pushpull, so that the operating line passes the point Q, thus being corrected as shown by the broken line. Presuming that the maximum plate current is I_{max} and the minimum plate voltage is E_{min}.

Amplitude of plate current AC component $I_p = I_{max}$

Amplitude of plate voltage AC component $E_p = E_b - E_{min}$

Then the output of two vacuum tubes is given by the following formula,

$$P_o = \frac{E_p I_p}{2} = \frac{(E_b - E_{min}) I_{max}}{2}$$

If the tube is operated on the condition that the plate current of no-signal state equals to zero, a distortion may be caused at the junction of positive and negative waveforms, because of the bend of the plate current characteristics. Therefore, the circuit is used on the condition that some amount of current is drawn, and this current value is determined at approx. 5% of the plate current at the maximum signal condition.

Next, in case of class B2 pushpull grid current conducts at the maximum signal input. Therefore, the input impedance of this amplifier varies to a great extent depending on the amplitude of the signal, so that, if a normal voltage amplifier is applied to the preceding stage, prominent distortion of waveform may be introduced. It is effective for the prevention of this phenomenon to make the output impedance of the preceding state far lower than the fluctuation of the input impedance of the preceding stage far lower than the fluctuation of the input impedance. In actual broadcasing transmitters, it is an usual practice to apply a class A pushpull cathode follower of maultielectrode tubes as to the sub-modulator.

Triodes and the pentodes are applied to the class B_2 pushpull amplification. However, since it is difficult and expensive to employ an adequate output multielectrode tube for high power plate modulation, the triode is exclusively used. Furthermore, it is a practice to use the same type of tube as the modulated tube.

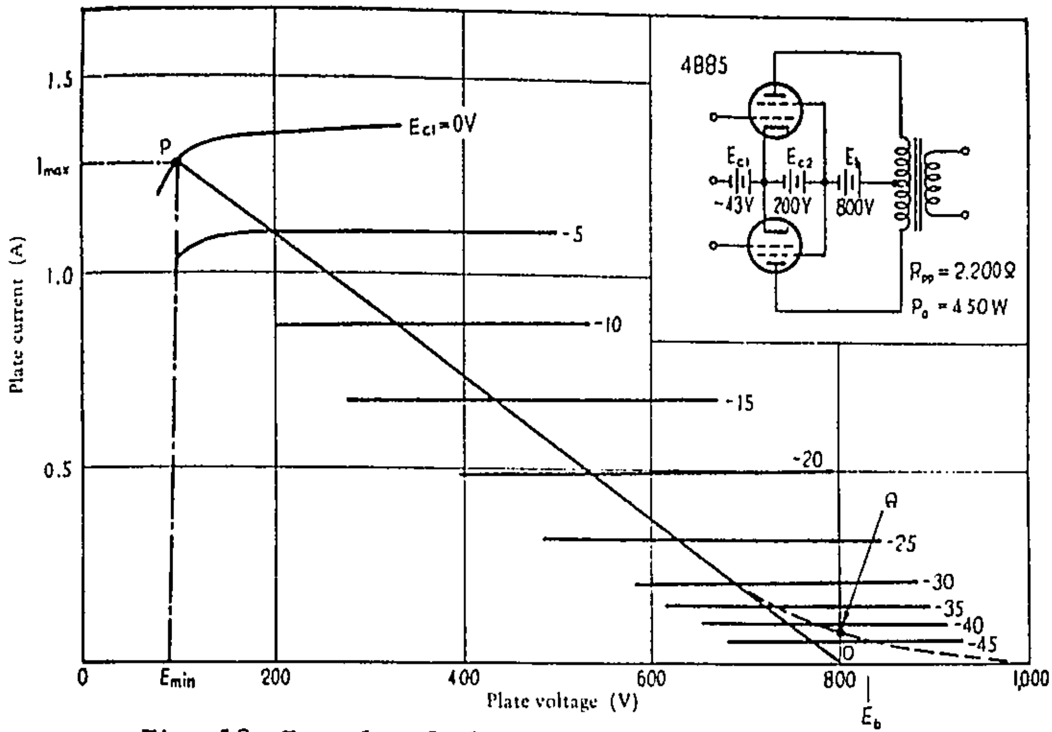


Fig. 18 Example of class B_1 pushpull operation of 4B85

In case of class B_2 amplifier, the plate efficiency can be raised up to approx. 60 % at maximum even if a triode is employed. The plate efficiency, when a multielectrode tube is used at class B_2 pushpull, is almost similar to that when the same tube is used at class B_1, i.e. approx. 60 %, however, the class B_2 output is larger of course.

An example of class B_2 operation of 5T31 triode is shown in Fig. 19. It is more convenient for the analysis of class B_2 operation on the characteristic curve to use the e_p - e_g characteristics, i.e. constant current curves than the e_p - e_i characteristics. The reason is that, when the grid voltage is maximum, plate current and grid current become maximum and plate voltage becomes minimum. In addition, the relation between plate current and grid current is determined by the minimum plate voltage and the maximum grid voltage, so that the constant current characteristic curve is more preferable to indicate this relationship more obviously on a sheet of paper. Description will be made later with regard to the constant current characteristics.

4.5 Frequency response and phase characteristics

As to audio amplifiers, frequency response and phase characteristics are very important. In particular, the phase characteristics are important for negative feedback amplifiers.

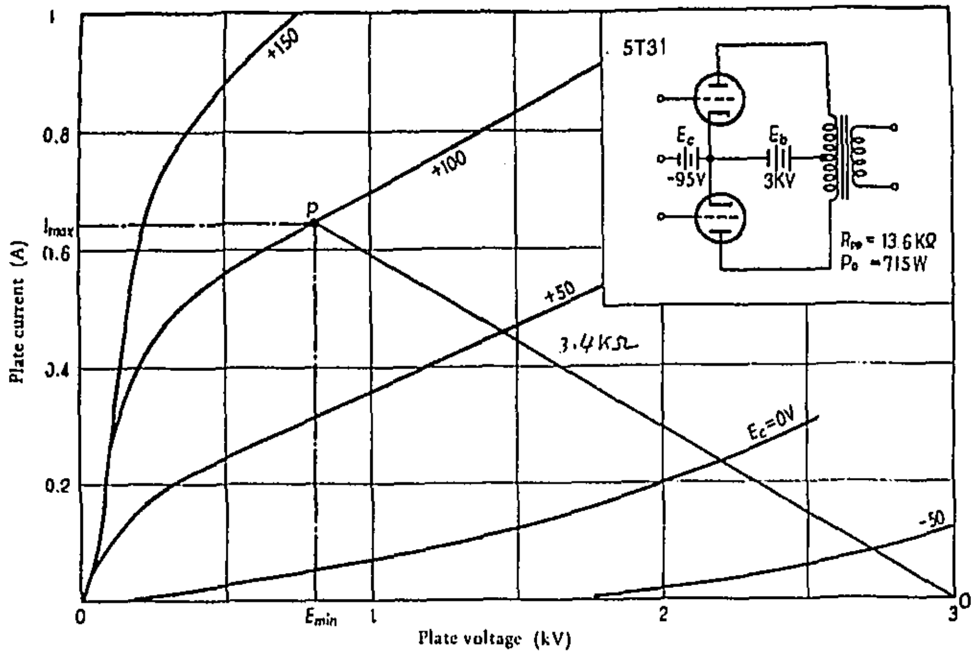


Fig. 19 Example of B_2 class pushpull operation of triode 5T31

Except for particular exceptions, in general, certain relationship holds between frequency response and phase characteristics of circuits. If a frequency response is given over the whole frequency band, from zero to infinitive frequency the corresponding phase characteristics can be theoretically obtained by calculation. The reverse is also possible. The calculation is somewhat mathematically complicated, however, the outline is as follows. If we plot the frequency characteristic curve with the amplitude in dB on the ordinate, and the frequency in log-scale on the abscissa,

(1) When the slope of frequency response is positive, the phase is leading, and when negative, it is lagging. Therefore, when the frequency response is flat, the phase angle is zero.
(2) When the frequency response is a constant slope over the

whole frequency range, the phase angle is simply expressed by the following equation.

(Phase angle) = (Slope of frequency response) x 15

The unit of phase angle is in degree, and that of the frequency response slope is in dB/octave. The octave means the relationship of frequencies where the ratio is two-fold or one-half. For instance, when a constant current is applied to an inductor, the output voltage increases proportionally to the frequency and, each time the frequency becomes twice, it increases by 6 dB, so that the phase angle is leading by 90°.

(3) When the frequency response is curved, the phase angle at an arbitrary frequency is influenced by the slopes of the response of other frequencies over the whole frequency range as well as affe-ted by the slope at that frequency. However, the degree of influence becomes less in correspondence with the increase of the distance to the frequency. Therefore, if the slope is linear through several octaves, the phase angle can be calculated by the slope of that point without significant error.

(Reference) Literature on the relationship between response and phase angle

Bode: "Network Analysis and Feedback Amplifier Design", 1949.

Terman: "Radio Engineer's Handbook", 1943

According to the above literature,

$$B_e = \frac{\pi}{12}\left(\frac{dA}{du}\right)_c + \frac{1}{6\pi}\int_{-\infty}^{+\infty}\left[\left(\frac{dA}{du}\right) - \left(\frac{dA}{du}\right)_c\right] \log_e \coth \frac{|u|}{2}\, du$$

where

B_c: Phase angle (radian) at the frequency f_c

A: Response (dB)

$\frac{dA}{du}$: Slope of response (dB/oct) u: $\log_e(f/f_c)$, f is the frequency, f_c is the frequency, where B_c is to be obtained.

coth x: Hyperbolic cotangent

$$\log_e \coth \left|\frac{u}{2}\right| = \text{Real part of } \log_e \coth \frac{u}{2}$$

$$= \log_e \left|\frac{f + f_c}{f - f_c}\right|$$

This term is the weighting factor which indicates the degree of the influence of the response slope at the frequency f to B_c. The suffix c indicates the value when $f = f_c$.

The frequency response and the phase characteristics of major amplification circuits are shown in Figs. 20 and 21. In Fig. 20(a), it drops a slope of 6 dB/oct at the low frequency range because of the coupling condenser and, in Fig. 20 (b), the same drop occurs at the high frequency range because of the output stray capacity, so that the limit of the phase angle is ($\pm$) 90°. In case of the transformer coupled amplifier of Fig. 20 (c), it drops in a slope of 6 dB/oct at the low frequency range because of the influence of the inductance, so that the phase angle is leading 90°. At high frequency range, it drops a slope of 12 dB/oct because of the resonance between leakage inductance and stray capacity, so that the phase angle is lagging 180°.

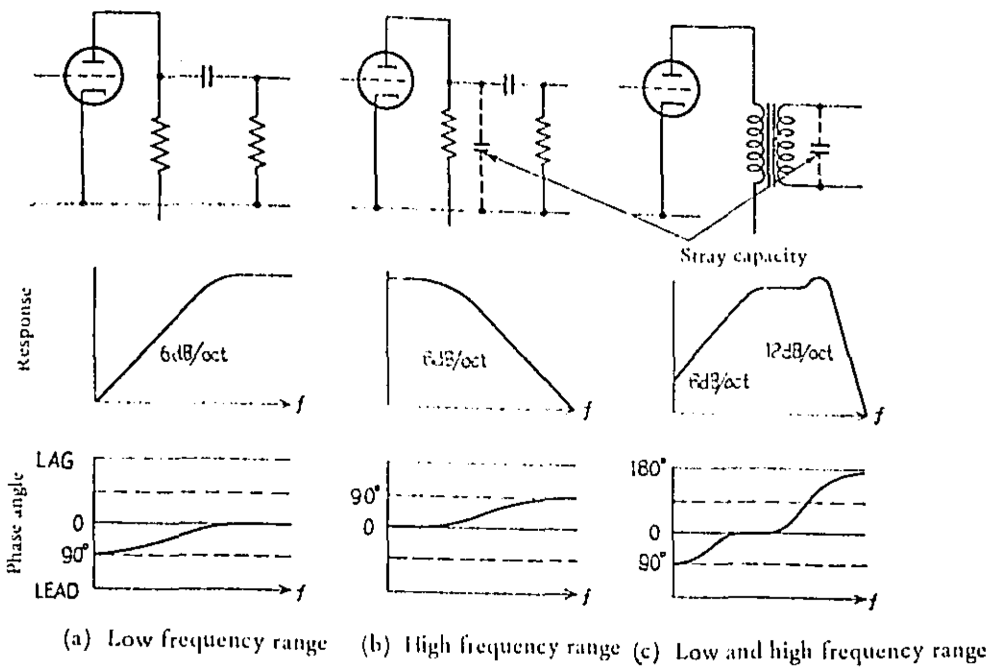
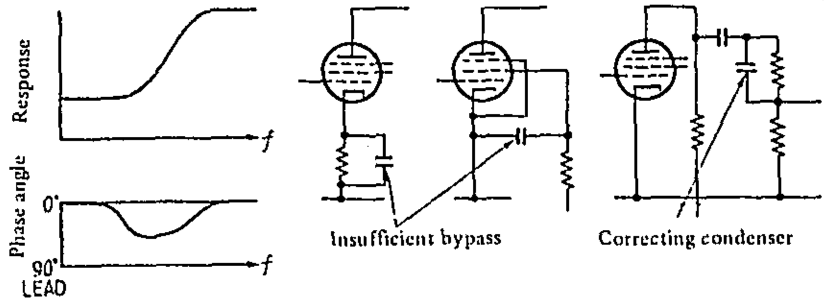


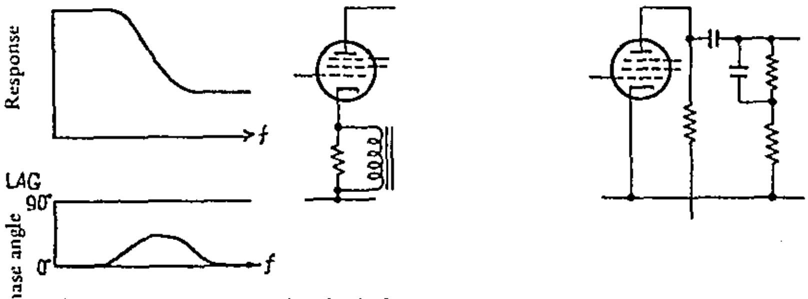
Fig. 20 The case that the response drops in a constant slope at extremities of low and high frequencies

In various circuits shown in Fig. 21, the response drops at low and high frequencies, however there are two cases, i.e. one is where a constant response value is attained in the extreme frequency and the other is where a peak appears at a certain frequency. In these circuits, the phase angle deviation is generated only at the portion where the response changes and, at other portions, it is zero. These are utilized as a phase compensation circuits in for negative feedback which will be stated later.

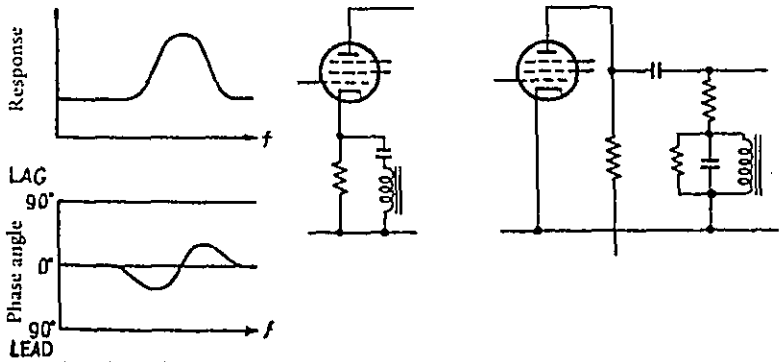
Frequency response and phase characteristics of the amplifier have common properties in correspondence with the various combinations of the reactance and the resistance, so that charts can be generated with regard to some representative combinations by means of previous calculations. A group of such charts is called universal curves.



(a) A step is generated at low frequency range



(b) A step is generated at high frequency range



(c) A peak is generated at certain frequency

Fig. 21 The case that the response drop is finite at low and high frequencies

(Reference literature)

Tsuruo Shimayama: Design and adjustment of audio amplifier

4.6 Fundamentals of negative feedback circuit

Negative feedback is widely applied since its application to the audio amplifier causes drastic improvement of the properties. The a portion shown in Fig. 22 is an amplification circuit. If the amplification degree is presumed as a, it holds

$a = e_2/e_1$

Further, the portion indicated by β is a voltage divider circuit which divides a part of the output voltage e_2,

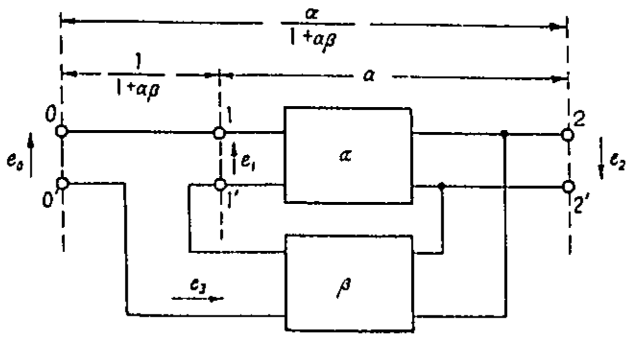


Fig. 22 Negative feedback circuit

and takes out the voltage e_3 having the following relationship.

$\beta = e_3/e_2$

Then this voltage is applied to the terminals 1 - 1' with reverse phase to the input signal e_0.

If presuming that the net input voltage e_1 appears at the terminals 1 - 1' when input signal e_0 is applied to the input terminals 0 - 0', the following relation holds between these voltages.

$e_0 = e_1 + e_3$

However, the following relation is valid

$e_3 = \beta e_2$

$e_2 = ae_1$

Therefore, the following formula is established.

$e_0 = e_1 + a\beta e_1$

$$\frac{e_1}{e_0} = \frac{1}{1 + a\beta}$$

This is the ratio between voltage at the terminals 0 - 0' and the terminals 1 - 1'. Then the ratio of voltage at the terminals 0 - 0' and the terminals 2 - 2', i.e. the overall amplification degree A is given by the following equation.

$$A = \frac{1}{1 + a\beta} \times a \tag{26}$$

The first term of the right hand side of the formula is less than 1, indicating attenuation and the second term is larger than 1, indicating gain.

Thus, the negative feedback circuit is considered the combination of the attenuation indicated by $1/(1 + a\beta)$ and the gain indicated by a. In addition, the most distinctive feature is the attenuation has a relationship with the gain.

In case that $a\beta$ is sufficiently large in (26), i.e. $a\beta$ 1, the following equation holds.

$$A = \frac{1}{a\beta} \times a$$

$$= \frac{1}{\beta}$$

In other words, it becomes independent of a.

It virtually appears disadvantageous to apply negative feedback to amplifiers, since the amplification degree drops. However, the following advantages are ensured beyond the necessary costs to raise the gain of the preceding stage to compensate for the drop.

(1) Stability of amplification degree

As described above, the amplification degree is determined only by β when negative feedback is sufficiently applied, so that it will not be influenced by the supply voltage fluctuation, change in performance of vacuum tubes, etc. However, in case that the vacuum tube emission drastically decreases, the negative feed back does not compensate it since a becomes small. Also it is necessary that β will be firmly constant without being affected by frequency, temperature or voltage.

(2) Reduction of non-linear distortion

With regard to the internal distortion of the amplifier, the existance of an equivalent electromotive force is presumed in the amplifier. Then, by feeding the distortion voltage appearing at the output in reverse phase, it will work to cancel the internal electromotive force, so that the distortion will be lowered with regard to certain range of output voltage.

(3) Reduction of noise

This is also improved by the same reason as the non linear distortion. If the noise frequency band is wide, the negative feedback must be applied to cover the whoes band-width. Also, the β circuit itself should not generate noise nor pickup noise. The degree of noise improvement by the negative feedback is dependent upon whether the noise source is close to the input stage or to the output stage of the amplifier. For instance, if noise source exists in the grid side of the first stage amplifier, noise voltage is equally amplified regardless of the existence of the negative feedback, so that SN ratio stays unchanged. On the contrary, if noise source exists in the last stage vacuum tube, a part of the output noise voltage will enter the amplifier, with a reverse phase through the β circuit and be sufficiently amplified on the way to the noise source location, so that the original noise voltage will be effectively cancelled and the SN ratio will be improved.

(4) Improvement of frequency response

As obvious from section (1), if the characteristic of β is excellent, the amplification degree will be determined by β, and in case of resistor divided circuits, the frequency response becomes flat independant of the frequency.

(5) Reduction of internal resistance

The change of load impedance usually causes change of amplification degree α. However, that the overall amplification degree A is kept almost constant by application of negative feedback, it virtually means a drastical decrease in internal resistance. Actually the internal resistance also decreases to $1/(1 + a\beta)$. This means it is ineffectiveness of load impedance fluctuation, which is particulary advantageous with respect to the multielectrode tube of high internal resistance. As a result of internal resistance reduction, due to negative feedback, a question may arise whether the impedance matching shall be applied to this decreased resistance, but this is not necessary, because the load resistance generating the maximum output without distortion is given by the characteristic curve of that vacuum tube and is independent of the negative feedback. Also load matching means abandoment of the feature of internal resistance being low in comparison with the load resistance. However, when matching is required for prevention if reflection in case of connection to long lines, the bridge type feedback which internal resistance does not vary, by means of feedback, should be applied.

As described above, the negative feedback is an useful means of improving the characteristics of the audio amplifier circuit and its main advantage is the reduction of nonlinear distortion and noise. As a result, high efficiency, high power class B_2 audio amplifier is realized and the drastic reduction of cost is enabled with regard to AC heating of the vacuum tubes, simplification of smoothing circuit in plate power supply, etc.

On the other hand, the necessary costs are related to the slight improvement of the preceding stage circuit, which is almost negligible.

However, the negative feedback is not almighty. For instance, it cannot recover the distortion of the transmitter due to over modulation or the distortion generated by plate current cutoff, caused by over excitation of output tubes. Because, even by negative feedback the plate current can not conducted to reverse direction.

4.7 Method of negative feedback

In a negative feedback circuit, if the feedback voltage is in reverse phase to the input signal, $(1 + a\beta)$ is larger than 1 as described above, so that $1/(1 + a\beta)$ is less than 1, resulting a reduction of distortion and noise. If the phase angle of the output voltage is different by 180° from that of the input voltage, the negative feedback will be in phase and the feedback will become positive. The value $a\beta$ is negative in this case, so that, if $1 > a\beta$ holds, $1/(1- a\beta)$ becomes larger than 1, therefore both amplification degree and distortion increases correspondingly. In an extreme case, oscillation will take place if $a\beta = 1$.

In general, $a\beta$ is a complex number varying dependent of the frequency and the locus of the $a\beta$ on a complex plane makes a closed curve along the frequency sweep from zero to infinity as show in Fig. 23. In this case, Nyquist has proved that the point (1, 0) shall not be included inside of the curve for stable operation without oscillation occuring at any frequency.

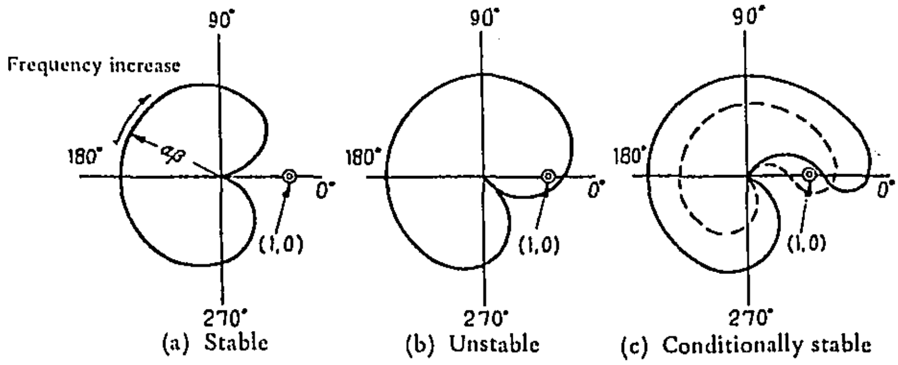


Fig. 23 Nyquist diagram

In Fig. 23 (a) the point (1, 0) is outside of the aβ curve, so that it is stable, and in (b), the point is inside the curve, so that it is unstable. In (c), the condition is stable for the moment but, if the amplification degree a drops such as to result in aβ of the broken line, it becomes unstable. Therefore, this type of form is not preferable.

It is usual practice to apply a resistance divider circuit as a β circuit, therefore the phase angle of aβ may be deemed to be the phase angle of the amplification degree a. Thus, it shall hold for the stability of the negative feedback circuit that the phase angle of the amplification degree is less than 180° when $|a\beta|$ $|a\beta| = 1$ and that, in such frequency as the phase angle is (±) 180°, $|a\beta| < 1$.

In a resistance coupled single stage amplifier, there is no possibility of oscillation, since the phase angle is 90° at high and low frequency range as shown in Fig. 20. In case of 2 stages amplifier, the phase angle becomes 180°. However, aβ in this case becomes 0 as shown in Fig. 23 (a), therefore there is no possibility of oscillation. In case of 3 strages or more, the possibility of oscillation arises since the phase angle becomes 270° or larger. In case of transformer coupled circuit, no oscillation is caused in case of 1 stage. However, if a resistance coupled stage is added to it, there would be a risk of oscillation.

As a conclusion, it is a vital condition for establishing stable negative feedback to keep the phase angle less than 180° in a frequency range of $a\beta > 1$, i.e. to keep the slope of the response less than 12 dB/oct. For instance, suppose the curve (a) of Fig. 24 is a response of a certain amplifier. To apply 20 dB negative feedback thereto.

$$1 + a\beta = 20 \text{ dB} \qquad \therefore a\beta = 19.08 \text{ dB}$$

And, if we take marginal 6 dB for safety, it is required to work stably up to the following value.

$$1 + a\beta = 26 \text{ dB} \qquad \therefore a\beta = 25.7 \text{ dB}$$

In fig. 24 the line of (-) 25.7 dB indicates $a\beta = 1$. The slope of the response close to the point where the above line crosses with the curve "a" is approximately 12 dB/oct (or 40 dB/decade), so that the phase angle is close to 180° as shown by curve "b", resulting in possible oscillation. To prevent this, it is necessary to suppress the phase angle near that point down to about 150° at maximum. For this, it is effective to take steps like curve "c" on the response. Fig. 21 indicates various kinds of compensating circuits to be applied therein.

In the actual design, at first calculate the response and the phase characteristics without negative feedback using the universal curve and check whether the phase angle is 150° or less at the frequencies between f_1 and f_2 where $a\beta = 1$. If this condition is not satisfied, use the additional compensation circuit, and compensation should be made by the cut-and-try method.

4.8 Cathode follower

In class B_2 audio amplifiers, grid current conducts in part of the positive side of the grid voltage waveform, so that voltage drop is caused by the internal resistance of the preceding stage (called

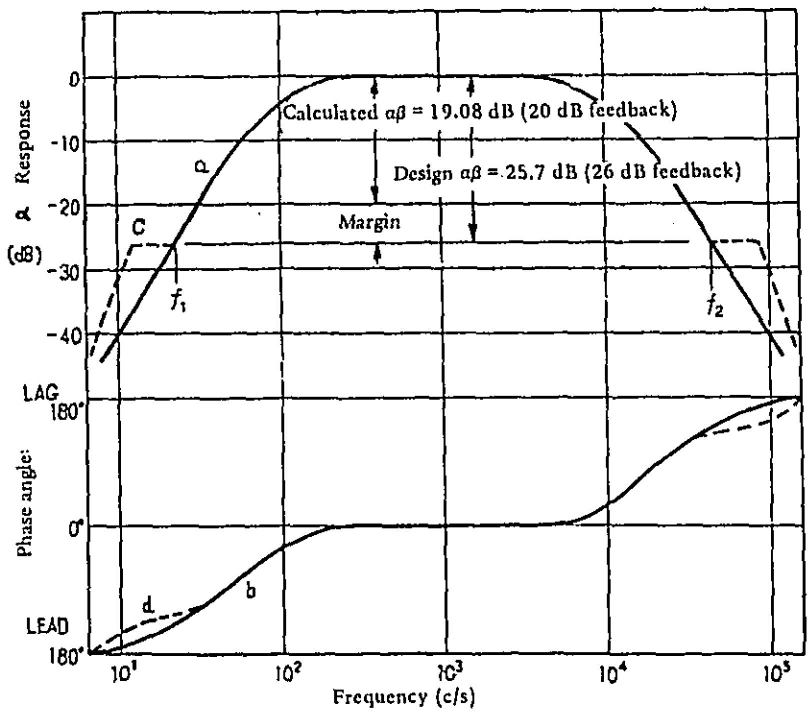


Fig. 24 Example of negative feedback

driver) resulting in the distortion of waveform. For the reduction of this, the internal resistance of the driver should be sufficiently small. For this purpose, a cathode follower is usually applied.

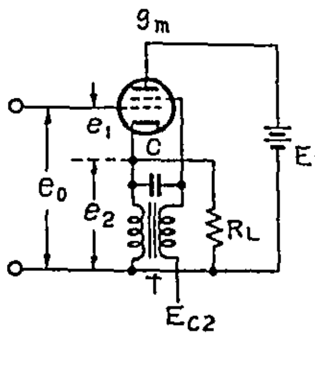


Fig. 25 Cathode follower

A cathode follower is a circuit where a load is connected to the cathode. Presuming that the transconductance of the vacuum tube is gm, then

$e_2 = e_1$ gm, $R_L = ae_1$

where a is an internal amplification degree and

$e_0 = e_1 + e_2$

Therefore,

$$\text{Amplification degree } A = \frac{e_1}{e_0} \quad \frac{g_m R_L}{1 + g_m R_L}$$

$$= \frac{a}{1 + a}$$

The above equation indicates that the amplification degree is always less than 1 and, when a is large, it is close to 1.

By reducing the above equation, the following equation is obtained.

$$A = \frac{R_L}{\frac{1}{gm} + R_L}$$

This equation indicates the amplification degree of an amplifier using a vacuum tube of amplification factor $\mu = 1$ and the internal resistance = 1/gm. Therefore, the equivalent internal resistance is equal to 1/gm. For instance the internal resistance of a beam tetrode tube 807 is several tens kΩ but, if gm = 4 m℧ in case of cathode follower, then

$$\text{Equivalent internal resitance} = \frac{1{,}000}{4} = 250\,\Omega$$

Thus it becomes extremely small.

Next, we shall consider the phase relationship. Referring to Fig. 26, the input voltage e_0 is a vectorial summation of the grid voltage e_1 and the output voltage e_2. In a frequency range where

e_2 is extremely larger than e_1, the phase difference θ between input and output does not become large irrespective of the phase between e_1 and e_2. For instance if a pure reactance load is involved, the phase angle between e_1 and e_2 is 90° and the ratio of e_1 and e_2, i.e. the internal amplification degree a = 50,

$$\theta = \tan^{-1}\frac{e_1}{e_2} = 1^\circ \quad g'$$

so it does not become significant. This is extremely advantageous in case of negative feedback.

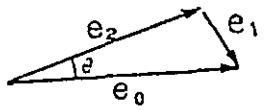


Fig. 26 Phase relationship

Then let us consider the input impedance.

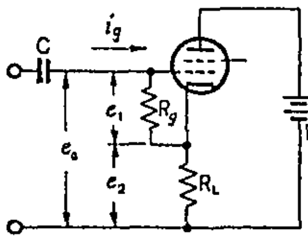


Fig. 27 Input impedance relationship

If we presume the input impedance of the vacuum tube as R_g, the following equation is obtained.

$$R_g = \frac{e_1}{i_g}$$

However, when viewed from the input terminal of the amplifier, current is i_g and the voltage is e_0. Therefore, the following relationship holds.

$$\text{Input impedance} = \frac{e_0}{i_g}$$

$$= \frac{e_1\ (1 + a)}{i_g}$$

$$= R_g(1 + a)$$

Thus, the apparent input impedance becomes very high, so that the necessary capacity of the input coupling condenser becomes extremely small. This is effective for the prevention of the blocking oscillation.

The cathode follower is a kind of negative feedback amplifier and provided with general features of negative feedback in addition to the above. Therefore, it is useful for the driver of the class B_2 modulator. Referring to Fig. 25, if a negative voltage corresponding to the grid bias of the modulator, is provided to the cathode as part of plate supply of the vacuum tube, direct coupling to the modulator grid is enabled without any coupling condenser. The transformer T is used to keep the screen grid at the same AC potential as that of the cathode and its ratio of transformation is 1 : 1. Further, the condenser C is a bypass to prevent the shift of the screen grid potential caused by the leakage inductance of T at higher frequency range.

4.9 Application of negative feedback to broadcasting transmitters

Recently there is no broadcasting transmitters which do not incorporate negative feedback. In the final stage plate modulation system, negative feedback is applied to overall audio amplification system including the modulator.

In the grid modulation system, distortion is generated by the modulation itself, therefore, rectified feedback system is applied, where the modulated wave is rectified and the audio frequency signal is extracted and feed back to the modulation system.

In case of plate modulation, rectified feedback is not required because the distortion by modulation is small. If the rectified feedback is to be applied, it would be difficult to provide a large amount of feedback, because of the large amount of phase shift caused by multi-stage audio amplifiers, especially of the modulated transformer included in the feedback loop, and furthermore, if the amount of rectified feedback is increased, phase shift of high frequency circuit will be added. On the contrary, in grid modulation, the number of amplification stages of the modulator is less, and phase shift due to modulator transformer will not be included. The adoption of rectified feedback will be easy, even if the phase shift of high frequency circuit is to be added.

(Example of negative feedback application)

Explanation is given to a practical 1 kW broadcasting transmitter shown in annex Fig. 2. Negative feedback is provided to the first audio amplifier of the modulator portion, but the feedback is limited only to that stage. The main feedback is applied to the grid of the second audio amplification tube from the plate of the modulator tube. The voltage amplification degree from the grid of the second audio amplifier tube to the plate of the modulator tube is given by,

$$a = 5950$$

Therefore, if we presume the negative feedback quantity at 18 dB, then

$$1 + a\beta = 7.93$$

$$\beta = 0.00116$$

must hold.

As for β circuit, a ladder voltage divider circuit is applied as shown in Fig. 28. VR is a potentiometer to balance the pushpull circuit. As is shown in Fig. 28 (b), if with a voltage divider circuit consisting of parallel resistance and condenser.

$$C_0 \quad R_0 = C_1 \quad R_1$$

then the following equation is established.

$$\beta = \frac{\dot{E}_2}{\dot{E}_1} = \frac{\dot{Z}_2}{\dot{Z}_1} = \frac{R_1}{R_\theta + R_1}$$

Therefore, β becomes independent of frequency, and no phase shift is generated. The selection of $C_0 R_0 = 20 \times 10^{-6}$ means that the reactance of C_0 becomes equal to R_0 at 8 kc.

R_1 = A parallel resistance of (1.5 kΩ + one half of VR) and 30kΩ

= 2.3 kΩ

$$\beta = \frac{2.3}{200 \times 10 + 2.3} = 0.00115$$

which meets with the requirement.

$$C_1 = \frac{20 \times 10^{-6}}{2.3 \times 10^3} = 8{,}700 \text{ pF}$$

Where C_1 includes the capacity of wiring from the voltage divider circuit to the grid. Therefore, if we presume this as 200 pF, a capacitor of 8,500 pF should be applied to C_1 of the voltage divider circuit.

Some results of response and phase angle at lower frequency range are calculated in Fig. 29 and 30.

Similarly in Fig. 31 and 32, the calculated results at high frequency range are shown. If we take a margin of 6 dB with regard to the negative feedback quantity of 18 dB, the phase angle at a gain drop of 24 dB must be 180° or less, which is understood to be sartisfied by the calculated results. The phase angle of the modulation transformer will not be involved in this calculation, because it is outside of the negative loop.

Fig. 33 indicates the measured result of the modulation frequency response with regard to the above example. As shown therein, maximum feedback is applied at about 200 - 800Hz. where the energy distribution of the audio current is high and the power-supply hum is included, therefore it is considered reasonable from the view point of improving distortion and reducing noise.

(Reference literatures of negative feedback)

Tsuruo Shimayama:	Disign and adjustment of audio amplifier
Terman:	Electronic and Radio Engineering
Terman:	Radio Engineers Handbook

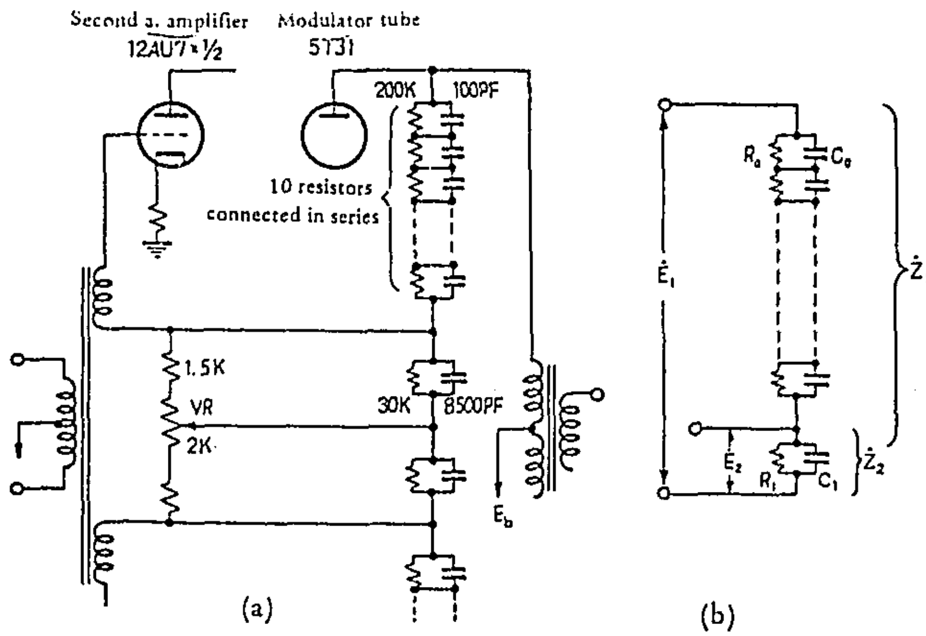


Fig. 28 Example of negative feedback circuit

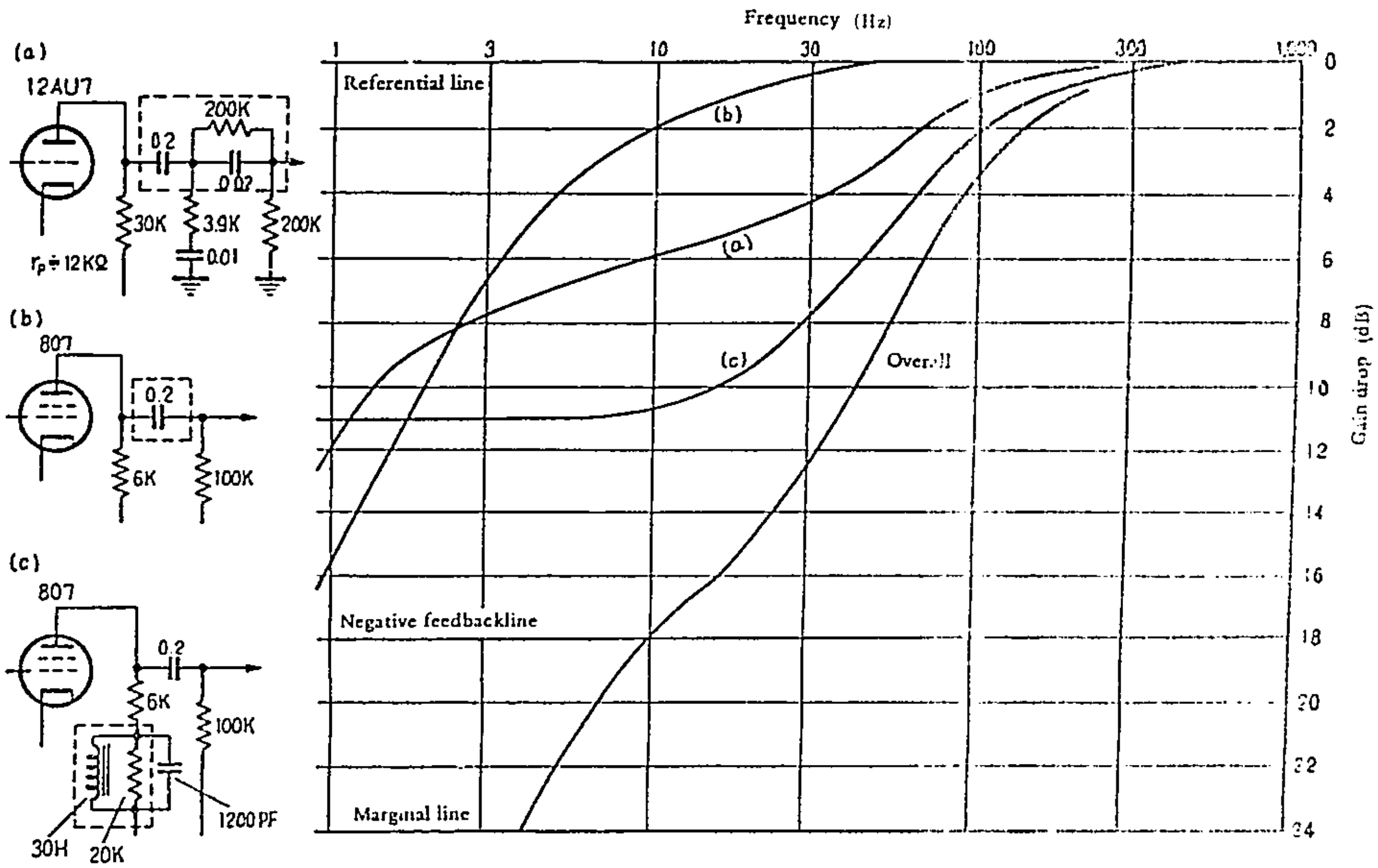


Fig. 29 Response with low frequency compensation

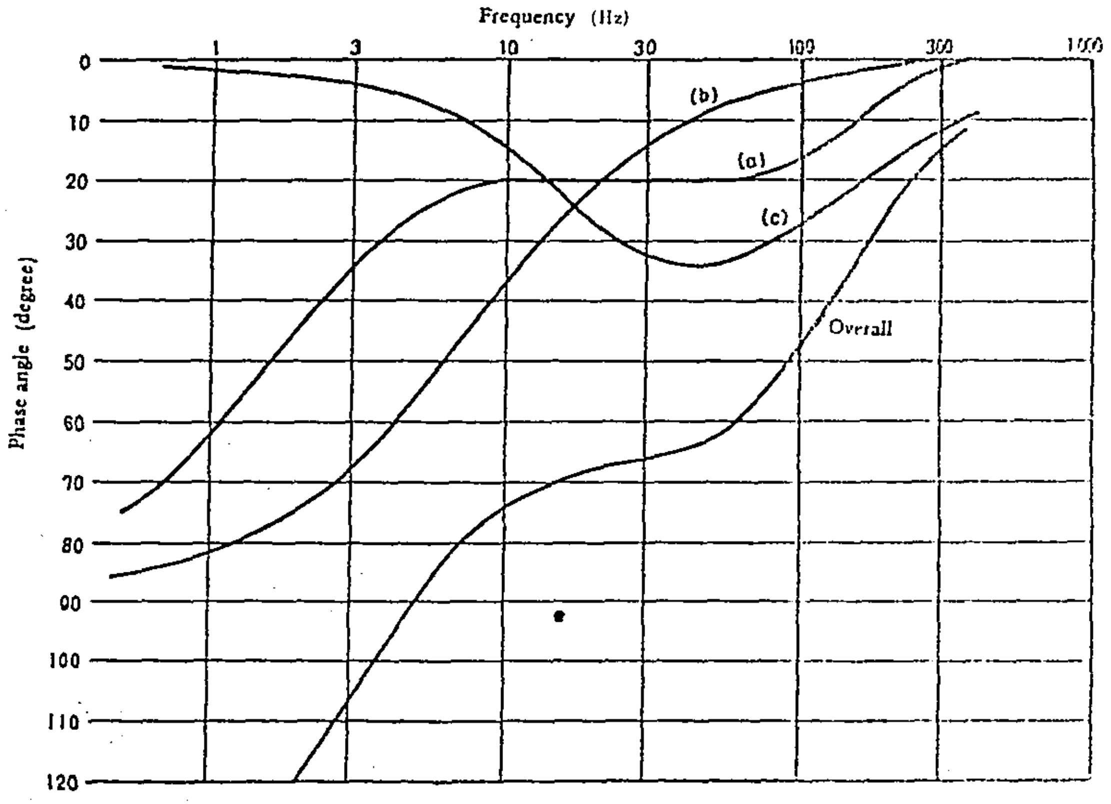


Fig. 30 Phase characteristics with low frequency compensation

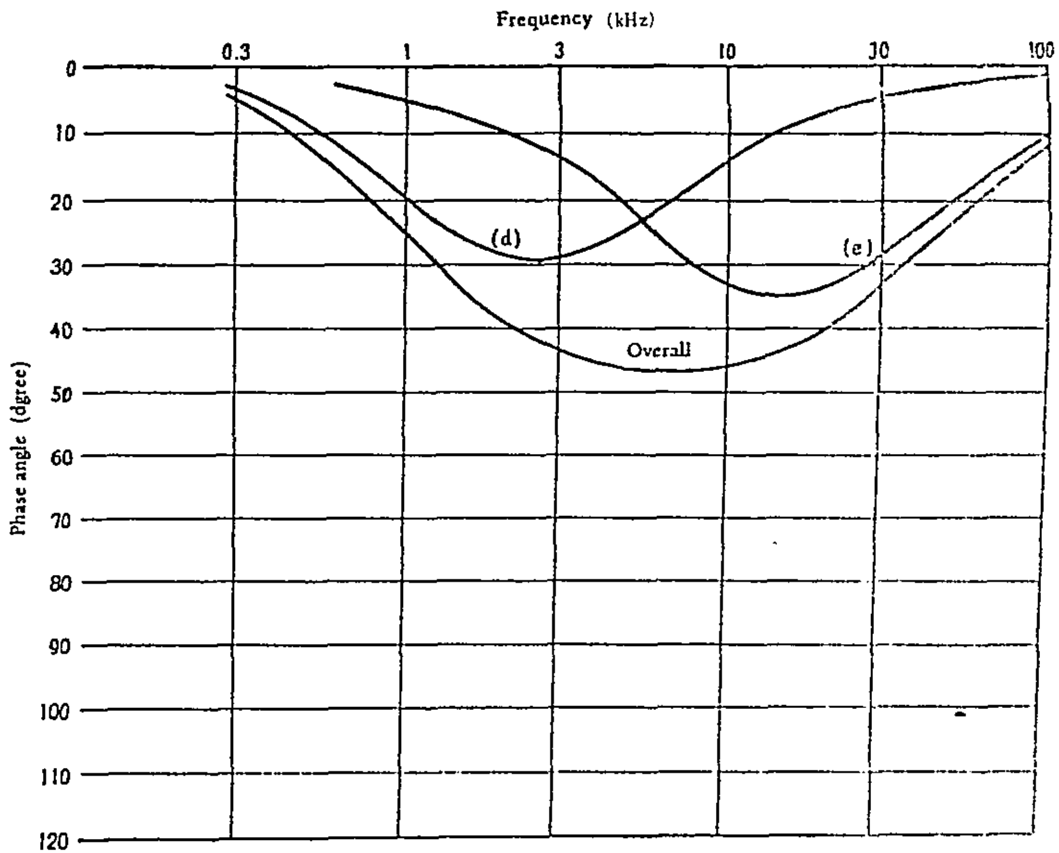
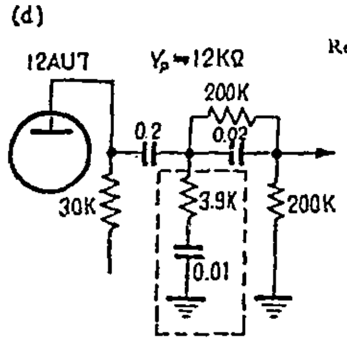


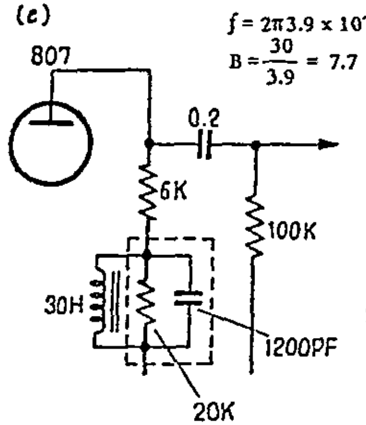
Fig. 32 Phase characteristics with high frequency compensation



$3.9K \times 0.01\mu = 3.9 \times 10^{-5}$

$f = 2\pi 3.9 \times 10^{-5} = 4{,}000$

$B = \frac{30}{3.9} = 7.7$



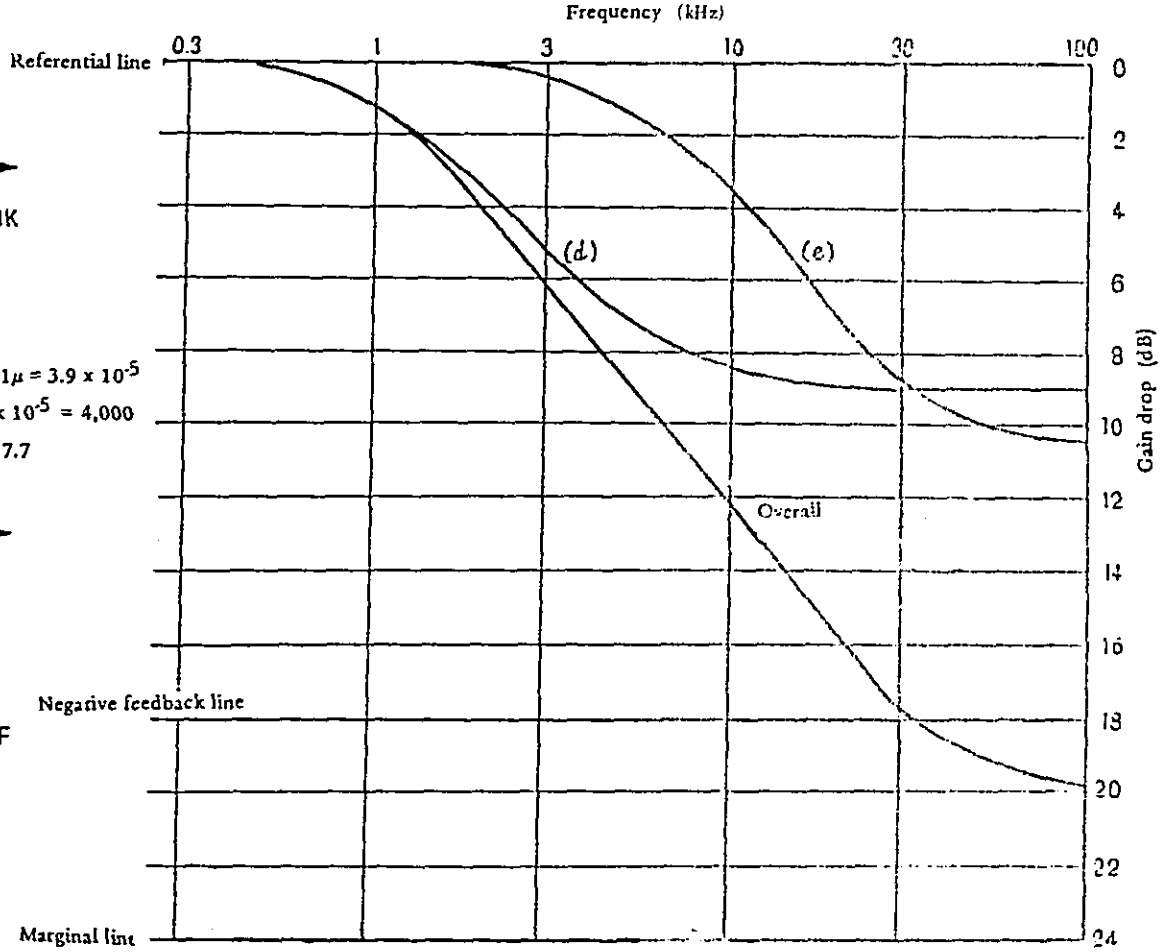


Fig. 31 Response with high frequency compensation

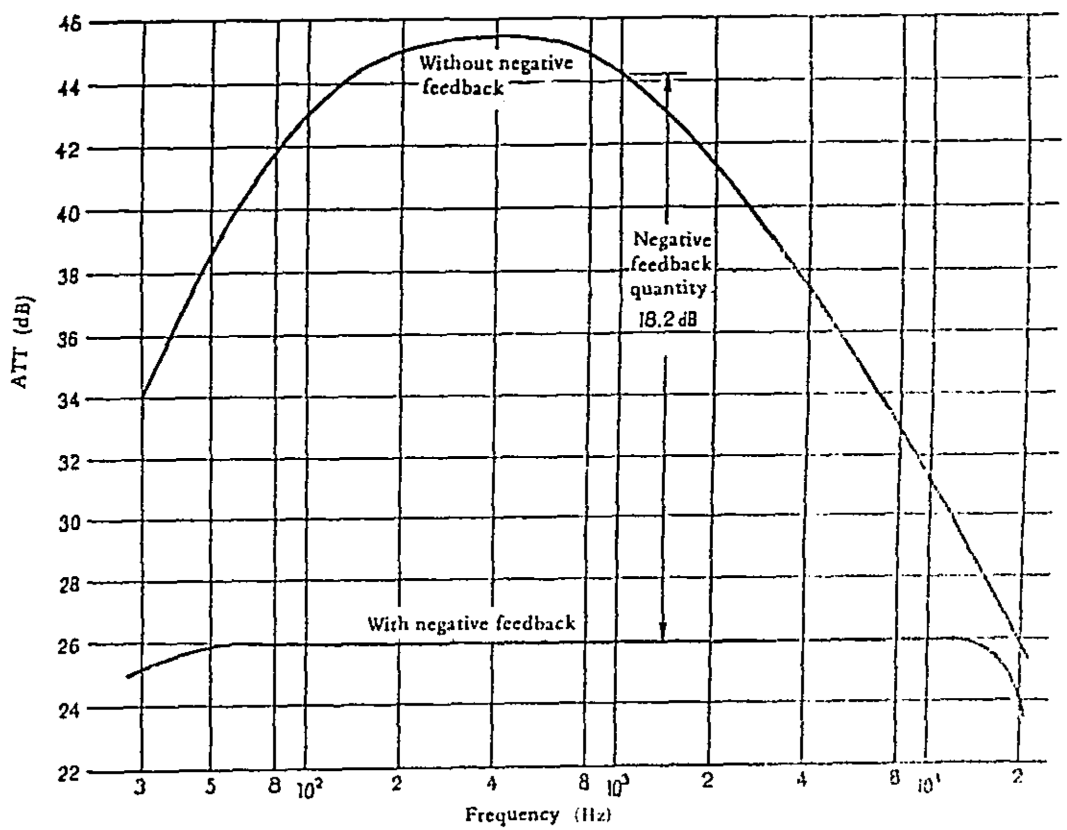


Fig. 33 Measured value of the modulation frequency response

5. Oscillator

5.1 Electrical characteristics of quartz plate

Since a rigorous specification is applied to the frequency allowance of radiation waves of broadcasting transmitters, an extremely stable frequency oscillator is required. Therefore, a crystal controlled oscillator is applied for this purpose.

Fig. 34 indicates an electrical equivalent circuit of a oscillating quartz plate, which is understood to be a kind of resonance circuit. In this figure, c_1 indicates the capacity between electrodes holding the crystal and L.C.R. indicate mass, compliance, and abrasion resistance of the quartz plate respectively.

An example of measured values of these are indicated as follows.

Frequency (oscillation in the direction of thickness)	430 kHz
Thickness	0.636 cm
Width	3.33 cm
Length	2.75 cm

$L = 3.3H$

$C = 0.042\mu\mu F$

$C_1 = 5.8\mu\mu F$

$R = 4{,}500\Omega$

$Q = 2{,}300$

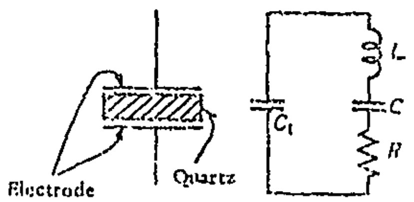


Fig. 34 Equivalent circuit of the crystal oscillator

(Refer to Terman, Radio Engineers Handbook)

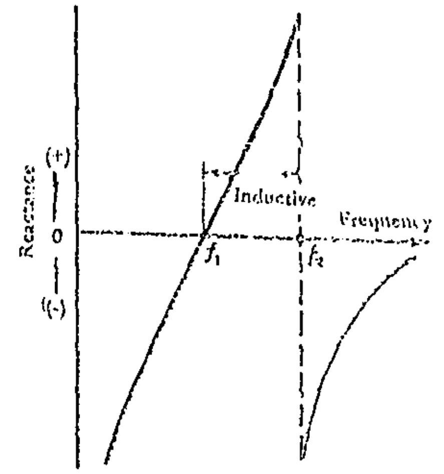


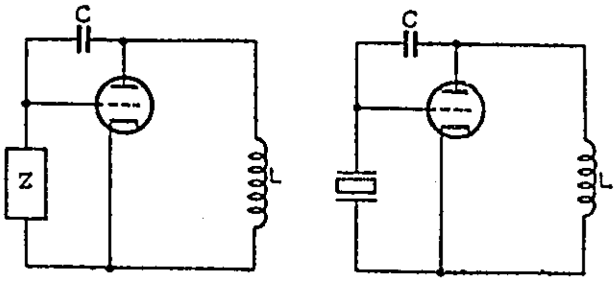
Fig. 35 Impedance variation of the crystal oscillator

According to this, the Q of the resonance circuit is very high while the crystal is oscillating, and can not be realized even by the electrical resonance circuit.

Fig. 35 indicates the variation of impedance of the quartz plate oscillator in oscillation, f_2 is the resonance frequency of the circuit of Fig. 34 in parallel resonance and f_1 is the series resonance frequency of the quartz plate itself. The resomnance circuit has a very high Q, and C, is much larger than C, so f_1 is determined by the constant of the crystal oscillator itself and f_2 is also very close to f_1. In addition, the reactance is inductive at the extremely narrow frequency range between f_1 and f_2, and at other frequencies, it is capacitive.

5.2 Generation of oscillation

In general, when the phase and amplitude of a feedback amplifier is equal to the input voltage, the amplifier will oscillate. In case of a single stage amplifier, the input voltage phase is reverse to the output voltage, so a phase-shift circuit of 180° is required for generation of oscillation.



(a) Feedback amplifier having leading phase shift circuit

(b) Crystal oscillator

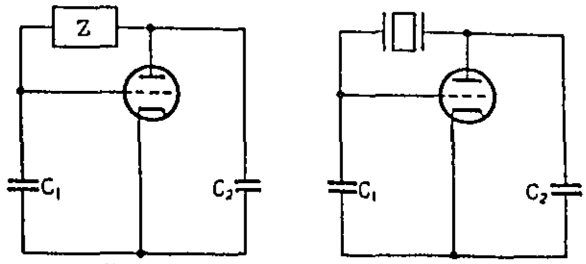
Fig. 36 Principle of GK type crystal oscillator

Fig. 36 (a) indicates a circuit of leading phase. In case that the impedance Z is capacitive the maximum variation of phase is 90° at the extreme low frequency range, so that it will not conform to the oscillation condition. If Z is purely resistive, a phase variation of 180° occurs at extremely low frequency range, but, due to the reduction of the feedback voltage, no oscillation will generate. If Z becomes inductive enphase angle will easily become 180° and, presuming that the reactance of the inductance is not so small, the magnitude of the feedback voltage will become sufficient to cause oscillation.

Fig. 36 (b) indicates an example of a crystal oscillator, where a crystal is inserted as for the impedance Z. This circuit oscillates while the impedance of the crystal oscillator is inductive. Therefore, the oscillation occurs at a frequency slightly higher than the mean value of f_1 and f_2, namely closer to f_2 of Fig. 35. In this system, as the crystal is connected between the grid and cathode, it is called GK type.

Fig. 37 (a) is a lagging phase circuit where, if Z is inductive and its reactance is not so high, the phase angle between input and output easily becomes 180° and the fe edback voltage turns out sufficient to cause oscillation. Replacing the impedance Z with a crystal, the circuit of Fig. 37 (b) is obtained.

In this circuit, the crystal impedance becomes inductive between f_1 and f_2 of Fig. 35. Therefore, it oscillates at a frequency closer to f_1. As the crystal is connected between the grid and plate, it is called the GP type.



(a) Feedback amplifier having lagging phase circuit (b) Crystal oscillator

Fig. 37 Principle of GP type crystal oscillator

The oscillation frequency exists between f_1 and f_2 with regard to either GK type or GP type and as these frequencies are expremely close, the stability of the oscillation is excellent, Since the crystal is put into a constant temperature bath to prevent the fluctuation of constants, it can cope with high grade requirements in respect to frequency deviation.

5.3 GK type crystal oscillator

Fig. 38 indicates an example of GK type crystal oscillator.

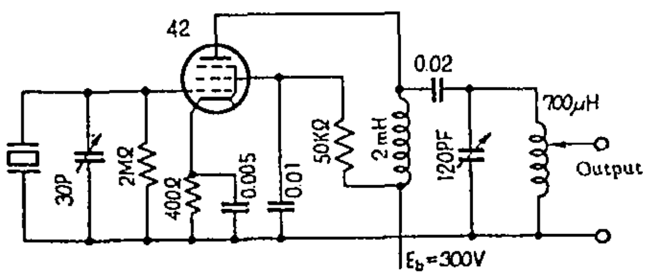


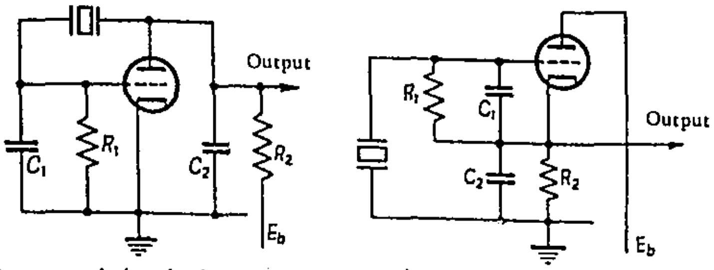
Fig. 38 Example of GK type oscillator

The feedback action from plate to grid is carried out by the internal static capacity between grid and plate of the vacuum tube 42. Therefore, no capacity is required to be connected outside. A tuning circuit is involved in the plate circuit. However, it is not for tuning with the oscillation frequency but it is to increase the inductive reactance of the load. As is widely known in a LC parallel circuit the parallel impedance becomes inductive and larger than the resctance of L when C takes a value less than that of resonance. The closer C approaches to the reśonance point, the higher this impedance becomes. Therefore, the oscillation strength correspondingly becomes stronger. If C becomes larger than the value of resonance, the parallel impedance becomes capacitive and the oscillation stops. Thus, if C is used at a value close to the resonance point, the oscillation strength is strong but the oscillation frequency is subjected to fluctuate because of slight change of LC. Therefore it is preferable for frequency stabilization to reduce C and, as the case may be, to exclude the tuning variable condenser and apply L only.

The oscillation frequency of GK type can be finely adjusted by inserting of a small variable condenser in parallel to the crystal.

GK type crystal oscillator oscillates easily and its output is large. However, in regard to the stability, GP type described below is superior, therefore, the GK type oscillator is becoming expelled from the broadcasting transmitters gradually.

5.4 GP type crystal oscillator



(a) Grounded cathode (Pierce circuit) (b) Grounded (Sabaroff circuit)

Fig. 39 Two systems of GP type oscillation circuit

In GP type oscillation circuits, there are two types, such as Fig. 39 (a) and (b), according whether the cathode or plate of the vacuum tube is grounded or not. The one in (b) is a cathode follower, a modification of (a), which is called Sabaroff circuit. High reactances are required with regard to C_1 and C_2, so that those capacities shall be small. Also, R_1 and R_2 shall be sufficiently larger than the reactances of C_1 and C_2, otherwise the oscillation becomes difficult. This oscillation circuit utilizes series resonance of the crystal, and the crystal impedance is low at the resonance point. Thus, if the frequency is to be adjusted by the parallel variable condenser, the effect will be less. Also, Sabaroff circuit incorporates predetermined constants without adjustable elements, so the oscillation frequency is not subject to fluctuation.

Fig. 40 is an example of a practical oscillator called "Sabaroff electron coupled circuit." In this circuit, the screen grid of the pentode 6CL6 plays the role of the plate of Fig. 39 (b) to constitute a Sabaroff oscillation circuit and the output is electronically coupled to the oscillator inside the vacuum tube. The features of this system is the stability of oscillation frequency, because the load is irrespective to the oscillat'on circuit, and that an large output can be obtained. This system is exclusively specified by BTS(Broadcasting Technical Standard) and is applied to NHK's broadcasting transmitters.

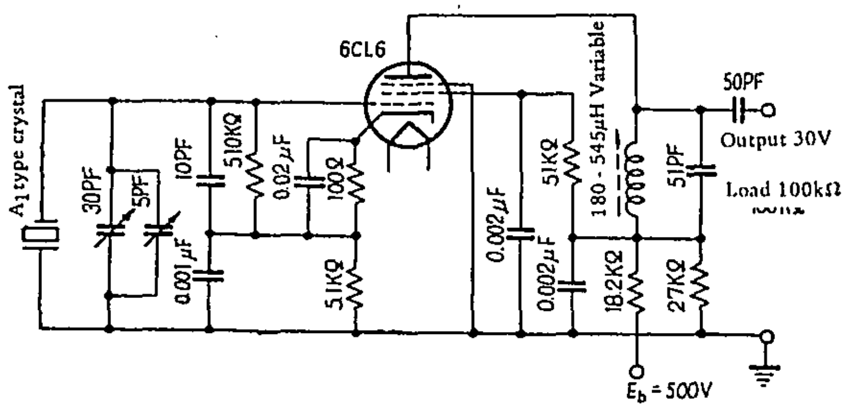


Fig. 40 BTS7102X1K-1 type crystal oscillator

6. High-frequency amplification circuit

6.1 Fundamental items

First, the symbols used in this section are described below.

E_b	Plate DC-voltage	E_p	Plate-voltage fundamental-wave amplitude
e_p	Plate-voltage instantaneous value	e_{pm}	Minimum plate-voltage
E_c	Grid DC-voltage	E_g	Grid-voltage amplitude
e_g	Grid voltage instantaneous value	e_{gm}	Maximum grid voltage
I_b	Plate DC current	I_p	Plate current fundamental-wave amplitude
i_p	Plate current instantaneous value	i_{pm}	Plate current peak-value
I_c	Grid DC current	I_g	Grid current fundamental-wave amplitude
i_g	Grid-current instantaneous value	i_{gm}	Grid current peak-value
θ_p	Plate current conduction-angle	θ_g	Grid current conduction-angle
ξ_v	Voltage utilization factor	ξ_i	Current utilization factor
R_L	Load resistance to fundamental wave	μ	Amplification factor of vacuum-tube
P_o	Plate output	P_i	Plate input
P_p	Plate loss	P_g	Grid loss
P_d	Excitation power	η_p	Plate efficiency

A high-frequency amplification circuit means an overall high-frequency system, including buffer amplifier, modulated amplifier and power amplifier of medium-wave broadcast transmitter as shown in Fig. 1. All power amplification circuits in the broadcasting transmitters can be deemed as power amplifiers. The operation thereof is of class B_2, class C_2 or its modified form. The majority of these circuits are single tube circuits and the use of pushpull tube circuits is very rare.

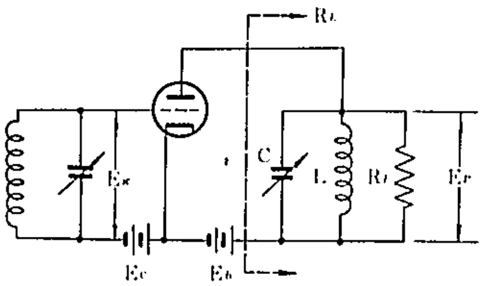


Fig. 41 High frequency amplification circuit

Fig. 41 indicates a fundamental circuit of high-frequency power amplifier. As the high-frequency amplifiers are to amplify single sinusoidal waves, a tuning circuit is usually provided in the input circuit. Since the plate current contains a large amount of harmonic components, an LC parallel resonance circuit which resonates to the fundamental wave, is inserted in parallel across Load R_L, in order to eliminate the harmonic components.
Thus, the load of the vacuum tube will become pure R_L resistance in respect of the fundamental wave. For other frequencies, it will become short-circuit by means of LC circuit and sinusoidal wave of the fundamental wave will only appear in the output.

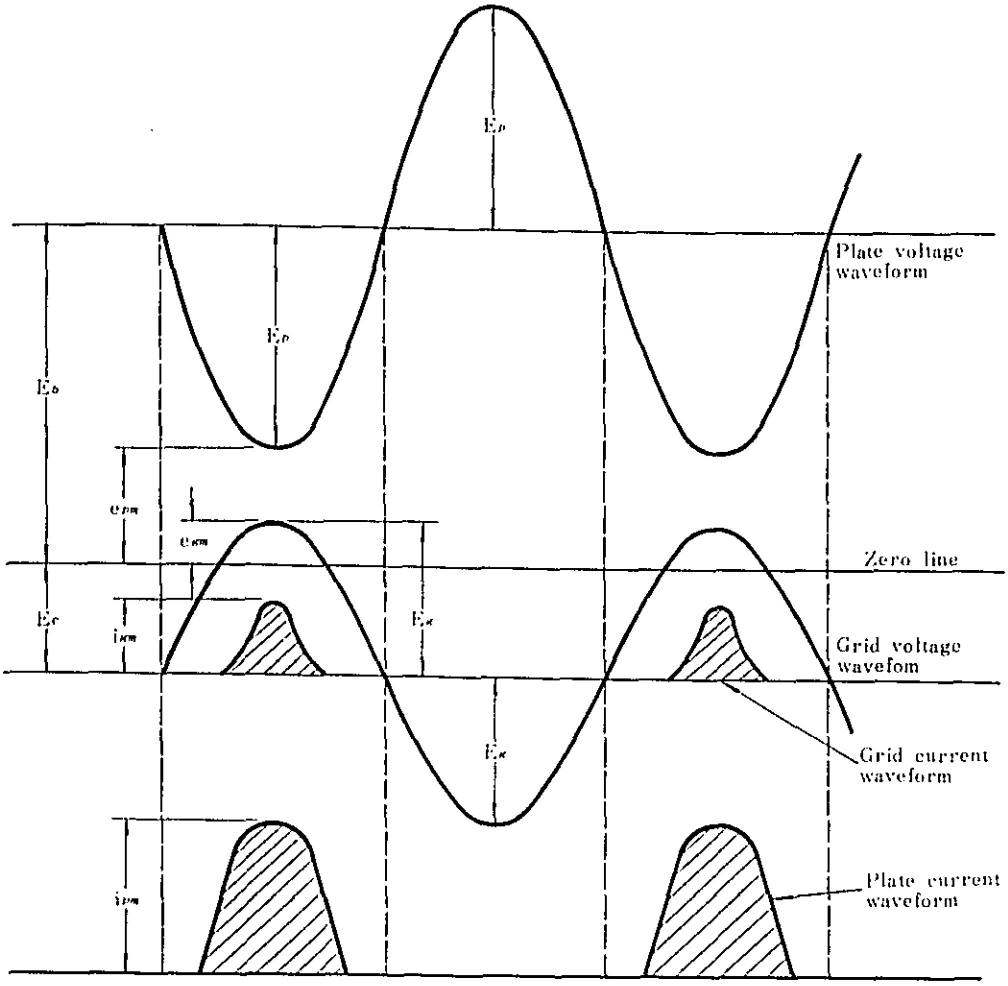


Fig. 42 Voltage and current waveform of high-frequency amplifier

Fig. 42 indicates the relation between the voltage applied to the plate, grid and the plate-current waveform of a high-frequency amplifier.

In this figure, the grid voltage will become positive potential at the positive peaks of the grid excitation, as a small amount of grid current will flow at that period.

The plate output voltage Ep and grid excitation voltage Eg are in reverse phase to each other and, when the plate potential becomes minimum e_{pm}, the grid potential will indicate maximum e_{gm}' and the plate current i_{pm} and the grid current i_{gm} will attain its maximum value. In addition to this, there exists relations as following between the voltages.

$$e_{gm} = E_g - E_c$$

$$e_{pm} = E_b - E_p$$

In order to improve the plate efficiency of actual power amplifiers, it will be necessary to raise the voltage-utilization factor indicated by E_p/E_b, but as the following relation holds,

$$\frac{E_p}{E_b} = 1 - \frac{e_{pm}}{E_b}$$

the e_{pm} becomes the element of determining the voltage-utilization factor. The value α expressed by the following equation shall be

$$\alpha = \frac{e_{pm}}{e_{gm}}$$

larger than 1, for reasonable operation of vacuum tubes. Namely, the value of e_{pm} must be kept always larger than that of e_{gm}. This means that a practical limit exists for the voltage-utilisation factor.

Referring to the equation (17) of the section 3.1, the voltage E is expressed by the following equation.

$$E = E_c + \frac{E_b}{\mu}$$

is called the DC lumped voltage. Presuming that a sinusoidal voltage E_g is applied to the grid, and by this voltage, a sinusoidal voltage E_p will appear at the plate circuit. Then, an AC voltage will be superposed onto the DC voltage and, moreover, since E_g and E_p are in reverse phase to each other, the lumped voltage will be expressed by the following equation.

$$E = (E_c + E_g) + \frac{(E_b - E_p)}{\mu}$$

$$= (E_c + \frac{E_b}{\mu}) + (E_g - \frac{E_p}{\mu})$$

$$= E_{dc} + E_{ac}$$

1; where

The first term of the above equation expresses the DC lumped voltage and the second term indicates the AC lumped voltage. When the composite value thereof becomes positive, the plate current will flow. When it becomes negative, the plate current will not flow. When it becomes exactly zero, it complies to the cutoff point. Fig. 43 indicates the above relationship graphically. As the hatched portion indicates the positive composite lumped voltage, plate current will flow within this range. Therefore, the plate current conduction-angle θ_ρ is given by the following equation.

$$\cos \theta_p = \frac{E_{dc}}{E_{ac}}$$

By substituting the former relationship to this, the following equation will be obtained.

$$\cos \theta_p = \frac{E_c + \frac{E_b}{\mu}}{E_g - \frac{E_p}{\mu}} \tag{27}$$

The above equation shows an important relationship of the plate current conduction-angle, and various modifications of this equation are practically used.

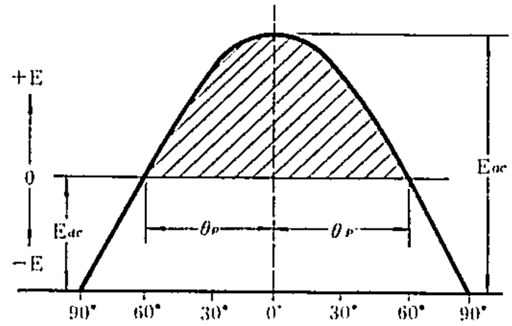


Fig. 43 Plate current conduction-angle

6.2 Operation analysis by constant-current characteristic curve

A constant-current characteristic currve is a curve of a vacuum tube with the plate-voltage in the abscissa and the grid voltage in ordinate with plate current (or grid current) taken as a parameter. Fig. 44 indicates the constant-current characteristics of triode 5T31, where the solid-line indicates the plate current and the broken-line indicates the grid current.

The features of the constant-current characteristics are as follows.

(1) The characteristics of the grid voltage at positive range are totally included in the first quadrant above $E_c = 0$.

(2) The plate current and grid current (in case of multi-electrode tubes, screen-grid current) can be indicated on the same plance.

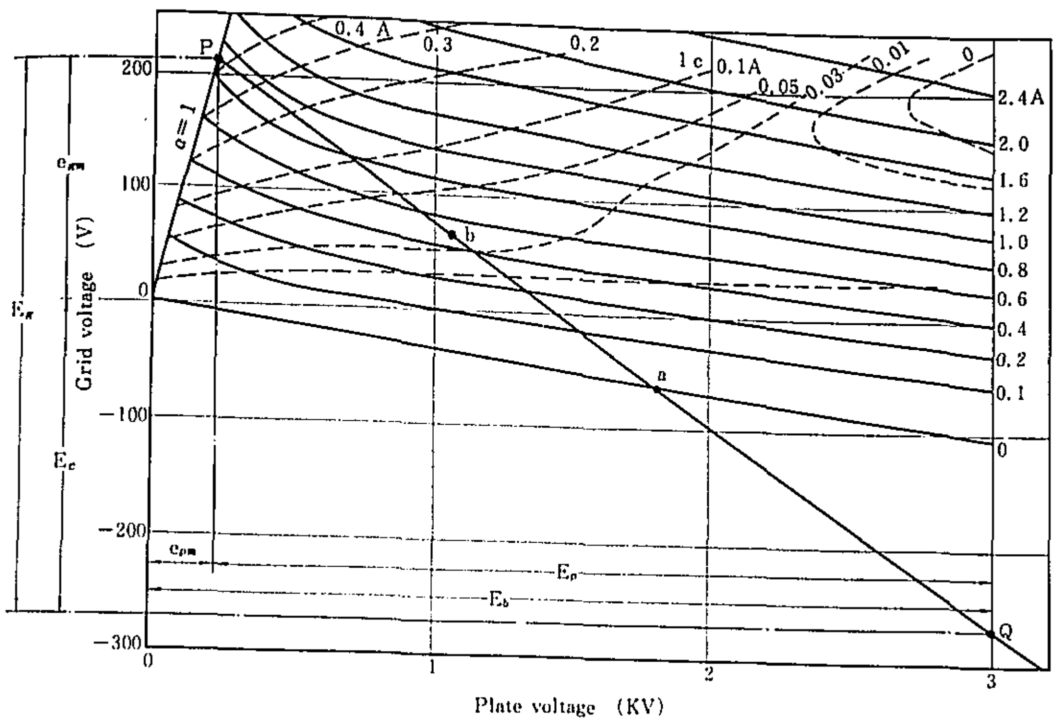


Fig. 44 Analysis of power amplifier (5T31) by constant-current characteristics

(3) The line indicating $E_c = E_b$ becomes linear. This complies to the line of $\alpha = 1$ in Fig. 44.

(4) The line of $I_b = 0$, i.e. cut-off line is clearly indicated.

(5) The gradient of the plate current curve indicates the amplication factor μ.

Since $\mu = \frac{\Delta Eb}{\Delta Ec}$ (Ib constant)., μ will become larger as the curve becomes a horizontal ine. As is obvious from the curve, the μ is constant over a considerably wide-range. However, the μ will decrease rapidly in the neighbour of $E_c = E_b$, and this is considered to be the practical limitation of application. Thus, as the magnitude of μ can be judged and the condition of its variation could be easily observed, it is called the " μ " curve.

(6) Referring to the upper right side of Fig. 44, as indicated by the broken lines of grid current zero and 0.01 A, there arises a position where E_b and E_c do not correspond to 1 : 1. The is because of the secondary electron emission properties of the grid, i.e. "Dynatron characteristics". This will become evident by converting the normal E_c - I_c characteristic curve (E_b = constant) to constant-current characteristics.

Referring to Fig. 44, when excitation is not applied to the amplifier, the point "Q" determined by E_b = 3kV and E_c = 270 V is a stationary point. Since the grid of vacuum tube is sufficiently biased for class C operation, no plate current will flow. Next, apply a sinusoidal wave excitation to the grid, and as the grid voltage increases, the plate voltage will decrease, due to the drop of voltage in the load and, when the grid voltage reaches maximum value of ℓ_{gm} = 210 V, the plate voltage will become to decrease to minimum value ℓ_{pm} = 210 V. This point is indicated by P. The locus indicating the instantaneous grid voltage and the instantaneous plate voltage coincides with the linear line connecting P and Q, because the AC grid voltage and AC plate voltage are sinusoidal and the phase difference therebetween is 180°. The current-value where the line PQ intersects the curve of the plate-current and the grid-current, is the instantaneous values of the plate-current and the grid-current, corresponding to the instantaneous grid-voltage at that time.

If the conduction-angle of the plate-current is measured to the front and rear direction, from the center of the plate current waveform as shown in Fig. 43, P point in Fig. 44 becomes the center of the waveform and the "a" point where PQ intersects the line of $I_b = 0$ indicates the cut-off point. Therefore, the conduction-angle of the plate current can be given by the following equation.

$$\cos\,\theta_p = \frac{aQ}{PQ}$$

Then, to obtain the current value of an arbitrary phase-angle, for instance θ, presume a point "b" on the PQ line that will fulfill the following condition and, then read the current of that point correspondingly.

$$\cos\,\theta = \frac{bQ}{PQ}$$

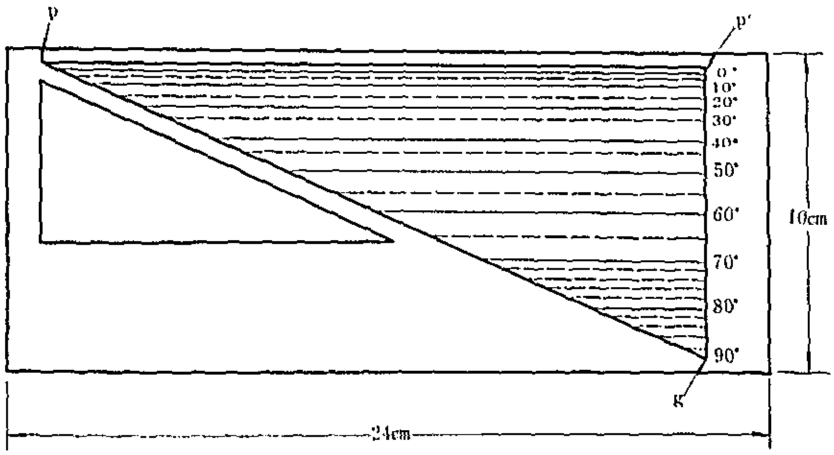


Fig. 45 Cosine scale

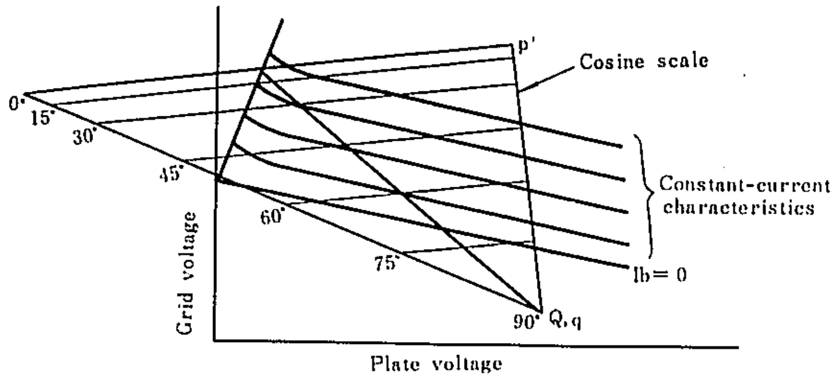


Fig. 46 How to use the cosine scale

In order to read out the current easily, the application of a cosine scale is convenient. The cosine scale is a transparent celluloid plate, as shown in Fig. 42, where a triangle qpp' is drawn and the distance between qp' is divided into degrees from 0° to 90°, in proportion to the cosine of the angle. In using this, as shown in Fig. 46, the summit q of the cosine scale is to be placed on the stationary point Q of the constant-current characteristics and, the line pp' is to be superposed on the upper point P of the working line. The intersecting point of the working line PQ and the angle sacle line will give the electrical angle that position. Then, read the plate current of that point. The plate current conduction-angle θ_p can easily be obtained by the intersection point of lines $I_b = 0$ and PQ.

When the relation between the electrical angle and the plate current is thus known, the mean value of the plate current, i.e. DC component, and the fundamental wave amplitude of the AC component can be calculated by the method described in the following section. In addition, the grid current can also be obtained by the same method.

Referring to Fig. 44 again, if the working line PQ is determined, it is obvious that following relations will hold from this drawing.

$$\left.\begin{aligned}
&\text{Output AC voltage} \quad E_p = E_b - e_{pm} \\
&\qquad\qquad\qquad\qquad\quad = 3{,}000 - 210 \\
&\qquad\qquad\qquad\qquad\quad = 2{,}790\text{V} \\
&\text{Excitation voltage} \quad E_g = |E_c| + e_{gm} \\
&\qquad\qquad\qquad\qquad\quad = 270 + 210 \\
&\qquad\qquad\qquad\qquad\quad = 480\ \text{V}
\end{aligned}\right\}$$

In this operation example, the point P is on the line of $\alpha = 1$ i.e. $e_{gm} = e_{pm}$. However, this relation is for the case when the voltage utilisation factor is taken as maximum, and therefore, it is not always like this.

Further, the working line extends downwards to the right from the stationary point Q, at the negative half-wave of excitation voltage, however as no plate current or grid current will flow within this range, this matter is unnecessary to be considered.

6.3 Analysis of current waveform by coefficient method

It is a well known fact that a periodic waveform even if it is deeply distorted, can be extended into an infinite series of triangular functions, i.e. Fourier Series, as shown hereunder.

$$\left.\begin{aligned}
&f(x) = a_0 + \sum_{n=1}^{\infty} (a_n \cos nx + b_n \sin nx) \\
&\text{Where} \quad a_0 = \frac{1}{2\pi}\int_{-\pi}^{\pi} f(x)\,dx \\
&\qquad\quad\ \ a_n = \frac{1}{\pi}\int_{-\pi}^{\pi} f(x) \cos nx\,dx \\
&\qquad\quad\ \ b_n = \frac{1}{\pi}\int_{-\pi}^{\pi} f(x) \sin nx\,dx
\end{aligned}\right\} \quad (28)$$

If f(x) = f(-x) holds, i.e. symmetrical with regard to 7-axis,

$$
\left.\begin{aligned}
a_o &= \frac{1}{\pi}\int_0^{\pi} f(x)\,dx \\
a_n &= \frac{2}{\pi}\int_0^{\pi} f(x)\cos nx\,dx \\
b_n &= 0
\end{aligned}\right\} \qquad (29)
$$

is obtained.

Now let us presume an ideal vacuum tube (a vacuum tube having a constant μ and g_m throughout the characteristic curve). If the grid exitation voltage is a sinusoidal waveform, the plate current waveform will become a partial sinusoidal waveform as shown in Fig. 47. Presuming that the amplitude of a sinusoidal wave is A, the plate current i_p will be given by the following equation.

$$i_p = A(\cos\theta - \cos\theta_p)$$

Since this waveform is symmetry with regard to the y-axis, the following equation can be obtained by applying equation (29).

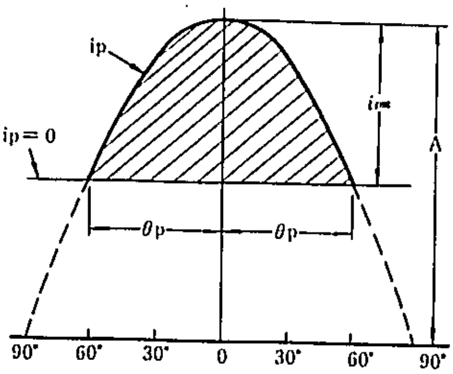


Fig. 47 Part of plate current waveform of sinusoidal wave

Mean plate current

$$I_b = \frac{1}{\pi}\int_0^{\theta_p} A(\cos\theta - \cos\theta_p)d\theta$$

$$= \frac{A}{\pi}(\sin\theta_p - \theta_p\cos\theta_p)$$

Plate current fundamental wave amplitude

$$I_p = \frac{2}{\pi}\int_0^{\theta_p} A(\cos\theta - \cos\theta_p)\cos\theta\, d\theta$$

$$= \frac{A}{\pi}(\theta_p - \sin\theta_p\cos\theta_p)$$

Also the peak value of the plate current i_{pm} is given by the following equation.

$$i_{pm} = A(1 - \cos\theta_p)$$

Therefore, the ratios of i_{pm} and I_b, I_p, etc. are given by the following equations.

$$\left.\begin{aligned} \frac{I_b}{i_{pm}} &= \frac{\sin\theta_p - \theta_p\cos\theta_p}{\pi(1 - \cos\theta_p)} \\ \frac{I_p}{i_{pm}} &= \frac{\theta_p - \sin\theta_p\cos\theta_p}{\pi(1 - \cos\theta_p)} \end{aligned}\right\} \quad (30)$$

Utilization factor of plate current

$$\left.\begin{aligned} \xi_i &= \frac{I_p}{I_b} \\ &= \frac{\theta_p - \sin\theta_p\cos\theta_p}{\sin\theta_p - \theta_p\cos\theta_p} \end{aligned}\right\} \quad (31)$$

As, the grid current increases rapidly when the grid voltage becomes positive, presume that it is proportional to the square of the sinusoidal partial wave, as given in the following equation.

However, this presumption is not accurate as to the case of the presumption of plate current characteristics.

$$i_g = B\,(\cos\theta - \cos\theta_g)^2$$

where B is a constant and θ_g is the grid current conduction-angle $= \frac{E_c}{E_g}$. By expanding this into Fourier Series, the following equations are obtained.

DC component of grid current I_c

$$= \frac{B}{2\pi}\left(\theta_g - \frac{3}{2}\sin 2\,\theta_g + 2\,\theta_g\cos^2\theta_g\right)$$

Fundamental-wave component of grid current I_g

$$= \frac{B}{\pi}\left(\frac{3}{2}\sin\theta_g + \frac{1}{6}\sin 3\,\theta_g - 2\,\theta_g\cos\theta_g\right)$$

Also the grid current peak value i_{gm} is given by

$$i_{gm} = B\,(1 - \cos\theta_g)^2$$

Therefore, the ratios of i_{gm} and I_c, I_g, etc. are given by the following equations.

$$\left.\begin{aligned} \frac{I_c}{i_{gm}} &= \frac{\theta_g - \frac{3}{2}\sin 2\,\theta_g\cos^2\theta_g}{2\pi\,(1-\cos\theta_g)^2} \\ \frac{I_g}{i_{gm}} &= \frac{\frac{3}{2}\sin\theta_g + \frac{1}{6}\sin 3\,\theta_g - 2\,\theta_g\cos\theta_g}{\pi\,(1-\cos\theta_g)^2} \end{aligned}\right\} \quad (32)$$

The equations (30), (31) and (32) are determined only by the current conduction -angle, the results calculated are shown in Fig. 48 and 49. As such calculations are based on the presumption of vacuum tube characteristics, they are called the coefficient method. The results calculated are somewhat different from the actual results, but as they can be calculated easily, it is a convenient method for rough designing of high-frequency amplifiers.

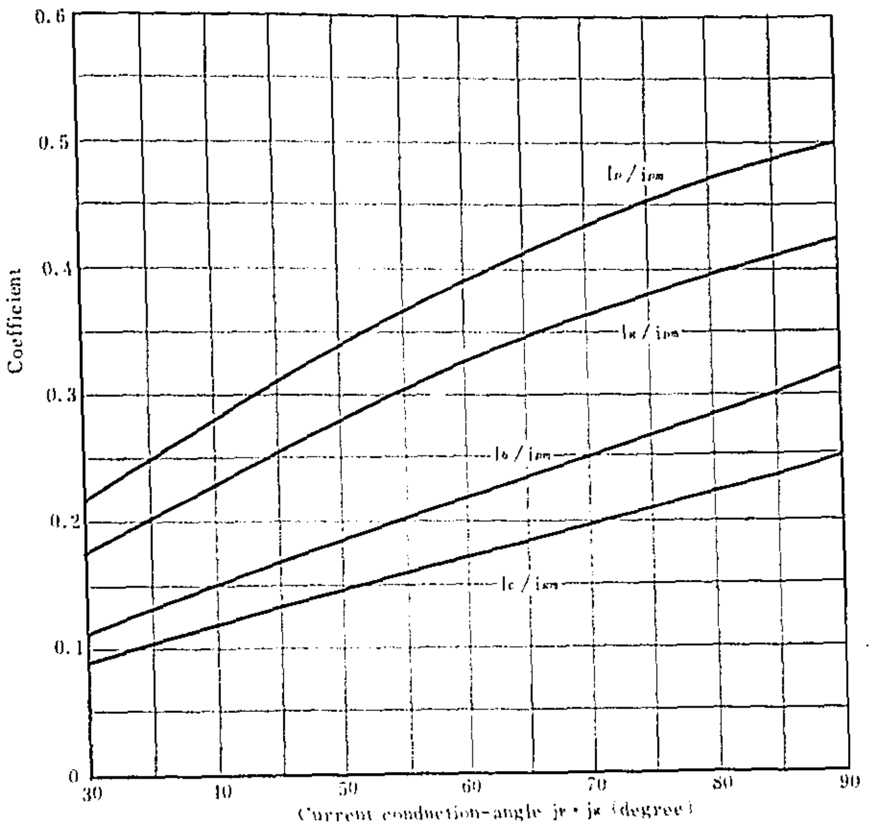


Fig. 48 Coefficient giving $I_b \cdot I_p$ and $I_c \cdot I_g$ in correspondence with i_{pm} and i_{gm} respectively.

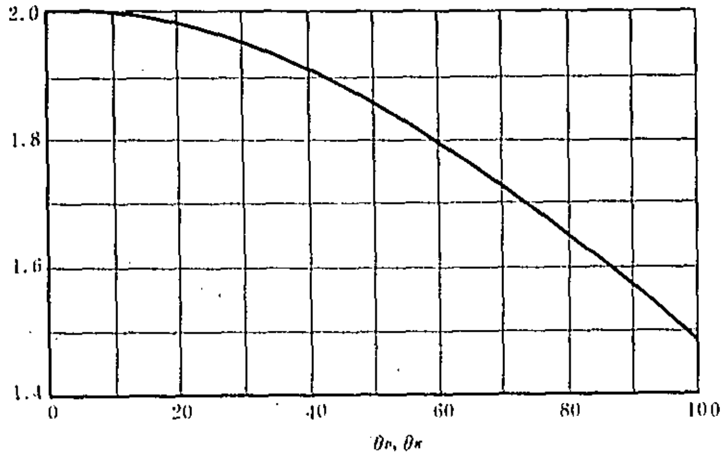


Fig. 49 Relation between current conduction-angle and current utilization factor

6.4 Analysis of current waveform by approximate integral method

The actual plate current and grid current waveforms are not so simple to be analysed mathematically, as in case of the coefficient method in the aforementioned section. Therefore, this approximate integral method is applied to analize such waveforms.

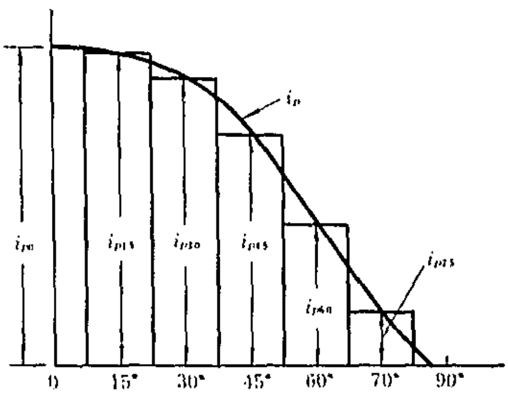


Fig. 50 Approximate integral method

Referring to Fig. 50, presume that the curve i_p is the plate current waveform under analysis. Divide the space between the electrical angle from 0° to 180° equally into 12 sections, i.e. each 15° and, that ip(0), ip(15), ip(30) ... etc. be current values at 0°, 15°, 30° ... etc., respectively.

Then the mean plate current and the amplitude of the fundamental wave of the plate current can be obtained by following equations in virtue of equation (29).

$$\left.\begin{aligned} I_b &= \frac{1}{12}\{\frac{1}{2}\,i_{p(0)} + i_{p(15)} + i_{p(30)} + i_{p(45)} + i_{(60)} + i_{p(75)}\} \\ I_p &= \frac{1}{6}\{\frac{1}{2}\,i_{p(0)} + i_{p(15)}\cos 15° + i_{p(30)}\cos 30° + \\ &\qquad i_{p(45)}\cos 45° + i_{p(60)}\cos 60° + i_{p(75)}\cos 75°\} \end{aligned}\right\} \quad (33)$$

(With regard to the equation to calculate the grid current, the suffix p and q are only to be replaced. Since the conduction angle of class B and class C amplifiers is less than 90°, it will be sufficient to calculate up to ip (75)).

(Example of calculation)

Fig. 51 gives the waveforms of ip and ig, corresponding to the electrical angle θ derived from the working line PQ of the constant current characteristics of Fig. 44.

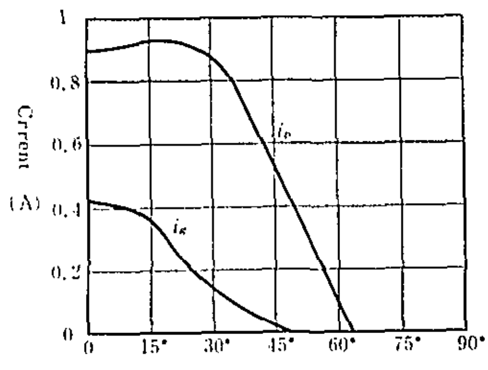


Fig. 51 Current waveform derived from the constant current characteristics of Fig. 44

In analyzing this, it will be convenient to make a table as shown below.

θ (Degree)	0	15	30	45	60	75	Total
i_p (A)	$0.9\times\frac{1}{2}$	0.92	0.88	0.52	0.08	0	2.85
$\cos\theta$	1	0.966	0.866	0.707	0.5	0.259	
$i_p \cos\theta$	0.45	0.89	0.76	0.37	0.04	0	2.51

$$I_b = \frac{1}{12}\left(\frac{0.9}{2} + 0.92 + 0.88 + 0.52 + 0.08\right) = \frac{2.85}{12} = 0.237(A)$$

$$I_p = \frac{1}{6}\left(\frac{0.9}{2} + 0.89 + 0.76 + 0.37 + 0.04\right) = \frac{2.51}{6} = 0.418(A)$$

The current utilization factor obtained from the above result is $\left(\frac{0.418}{0.237}\right. = 1.76$. On the other hand, the plate current conduction angle θ_p obtained from Fig. 44 is 65° and the corresponding current utilization factor derived from Fig. 50 is 1.76, which is equal to the above value despite of different way of approaching.

θ (Degree)	0	15	30	45	60	75	Total
i_g (A)	$0.42\times\frac{1}{2}$	0.35	0.15	0.04	0	0	0.75
$\cos\theta$	1	0.966	0.866	0.707	0.5	0.259	
$i_g \cos\theta$	0.21	0.34	0.13	0.03	0	0	0.71

$$I_c = \frac{1}{12}\left(\frac{0.42}{2} + 0.35 + 0.15 + 0.04\right) = \frac{0.75}{12} = 0.0625(A)$$

$$I_g = \frac{1}{6}\left(\frac{0.42}{2} + 0.34 + 0.13 + 0.03\right) = \frac{0.71}{6} = 0.118\ (A)$$

6.5 Calculation of operation condition of power amplifier

In power amplification, if the four conditions of E_b, E_c, e_{pm},

e_{gm} (or i_{pm}) are given, the working line can be determined on the constant current characteristic. Therefore, the necessary current value can be obtained by the waveform analysis according to the foregoing method. Thus, all operating conditions can be calculated by the following equation.

$$\left.\begin{aligned}
E_p &= E_b - e_{pm} \\
E_g &= E_c + e_{gm} \\
&I_b,\ I_p,\ I_c,\ I_g \text{ - Obtained from waveform analysis} \\
R_L &= E_p \,/\, I_p \\
P_o &= E_p\ I_p\ /2 \\
P_i &= E_b\ I_b \\
P_p &= P_i - P_o \\
\eta_p &= P_o/P_i \\
\xi_v &= E_p/E_b \\
\xi_i &= I_p/I_b \\
P_d &= E_g\ I_g/2 \\
P_g &= P_d - E_c\ I_c
\end{aligned}\right\} \qquad (34)$$

On the contrary, in designing a power amplifier, use the given values of E_b and P_o, with the presumed, values of ξ_v and θ_p to determined from experience for the calculation. As determining ξ_v means to determine e_{pm}, and determining θp means to determine ξ_i, therefore,

$$\begin{aligned}
e_{pm} &= E_b(1 - \xi_v) \\
\eta_p &= \xi_v \cdot \xi_i/2 \\
P_i &= P_o/\eta_p \\
I_b &= P_i/E_b
\end{aligned}$$

If I_b is known, i_{pm} can be obtained by the coefficient method in Fig. 48. If I_b is known, i_{pm} can be obtained by using the coefficient method in Fig. 48. If i_{pm} and ℓ_{pm} are known, P point can be determined on the constant current characteristic. After the P point is set, the stationary point P can be determined by using the cosine scale, so that the conduction angle will become θ_p.

Thus the temporary working line P Q will be obtained, the operation condition can be calculated by the method described above. If the results are far apart from the normal values of power amplifiers, the presumption of ξ_v and θ_p were inadequate, they should be reconsidered and the calculation should be done again, until the desirable condition is obtained.

In case α approaches 1, and the grid voltage rises, the plate current will not increase, but, instead, it will often decrease. As a result, a depression will occur in the plate current waveform, as shown in Fig. 51. Under this condition, the power will not increase, even if the grid excitation is increased, but exessive grid current will flow, thus resulting in surplus consumption of excitation power.

(Example of calculation)

Calculation will be carried out with regard to the working line PQ in Fig. 44.

$$\left.\begin{array}{ll} E_b = 3{,}000 \text{ V} & E_c = -\ 270 \text{ V} \\ e_{pm} = \ 210 \text{ V} & e_{gm} = \ 210 \text{ V} \end{array}\right\} \text{Given condition}$$

Then the calculation is continued by applying equation (34).

$$E_p = 3{,}000 - 210 = 2{,}790 \text{ V}$$

$$E_g = 270 + 210 = 480 \text{ V}$$

$$\left.\begin{array}{ll} I_b = 0.237 \text{ A} & I_p = 0.418 \text{ A} \\ I_c = 0.0625\text{A} & I_g = 0.118 \text{ A} \end{array}\right\} \text{Derived from the example of section 6.4.}$$

$$R_L = 2{,}790/0.418 = 6{,}680\Omega$$

$P_o = 2{,}790 \times 0.418/2 = 584$ W

$P_i = 3{,}000 \times 0.237 = 711$ W

$P_p = 711 - 584 = 127$ W

$\eta_p = 584/711 \times 100 = 82$ %

$\xi_\upsilon = 2{,}790/3{,}000 \times 100 = 93$ %

$\xi_i = 0.418/0.237 = 176$

$P_d = 480 \times 0.118/2 = 28.3$ W

Derived from the example of section 6.4.

$P_g = 28.3 - 270 \times 0.0625 = 11.5$ W

As for the equation to calculate the grid excitation power P_d, the following approximation may be applied in place of, $P_d = E_g I_g/2$.

$$P_d \fallingdotseq E_g\, I_c$$

The reason is that, since the grid current conduction angle j_g is generally small the, I_g/I_c obtained from (32) can be approximated to 2. By applying this equation, the results will be approximated as,

$$P_d \fallingdotseq 480 \times 0.0625 = 30 \text{ W}$$

Let us consider the variation of plate current and grid current current of class C high-frequency amplifiers, when the plate tank-circuit tuning is adjusted. Refer to the working line of Fig. 44. When the tuning is not completed, E_p is low and the upper end of the working line P exists on the right hand side. In accordance with the progress of tuning, P point moves to the left hand side. If point P moves to the left side, i_{pm} decreases and i_{gm} increases. Therefore, under the condition of tuning, DC plate current becomes minimum and the DC grid current becomes maximum. Thus, in practical, the minimum indication of plate current meter on the broadcasting transmitter gives the tuning point.

6.6 Class C grid modulation

In class C amplifiers, when the grid excitation voltage is kept constant, and the grid bias is varied, the high-frequency plate voltage will vary as shown in Fig. 52, with a considerable with range of linear portion. Then, by superposing an audio signal onto the grid bias corresponding to the middle part of the linear portion, modulation will be available. As obvious from Fig. 52, the plate voltage utilisation factor becomes maximum at positive peak of modulated wave, and the minimum plate voltage e_{pm} decreases to the allowable limit of amplifiers. Therefore it is impossible to deepen the modulation degree.
Therefore, in order to provide modulation up to 100 % modulations in the grid modulation method, the voltage utilisation factor at 100% modulation should beselected to a reasonable maximum value of class C amplification at 100 % modulation. Therefore, at carrier wave, it will become one half. If the voltage utilisation factor at 100 % modulation is 0.92, it should be 0.46 at carrier wave.

If we presume a plate current conduction angle of 50° at carrier wave, the plate efficiency will be given by the following equation, since the current utilisation factor is 1.85.

$$\eta_p = 0.46 \times 1.85/2 = 43 \%$$

In class C plate modulation, the plate efficiency is high when modulation is not provided. However, in class C grid modulation, the efficiency is low as described above. If adjustment is provided so that the efficiency at carrier wave is to exceed this value, the peak of the waveform will become collasped. Therefore, care should be taken in selecting load resistance and the working point at carrier wave.

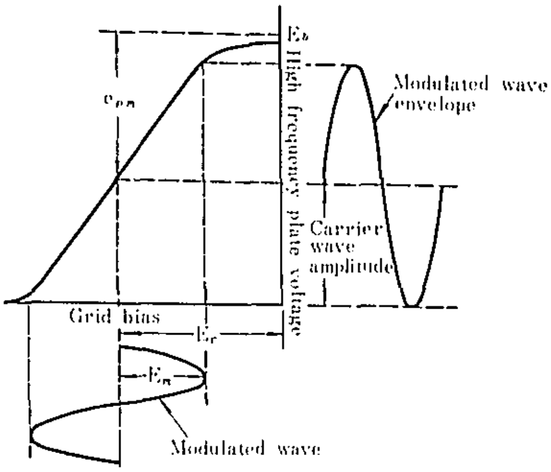


Fig. 52 C class grid modulation

Fig. 53 indicates and operating characteristics of class C grid modulation incorporating SN205C. The output at carrier wave is 185 W, and by combining three in parallel a rating output of 550 W is obtained. The operating condition obtained from this characteristic is shown in Table 3. PQ in the figure is a working line without modulation. At the positive peak of 100 % modulation, the bias becomes less in correspondence with the peak value E_m of the modulated wave, therefore point Q moves to Q'. By the reduction of the bias, the plate conduction angle θ_p as well as ip increase, so that the AC fundamental component I_p also increases. As a result, the plate AC voltage increases too, resulting in a movement of working point from P to P'. Therefore, P'Q' indicates the working line of 100 % modulation, of the positive side.

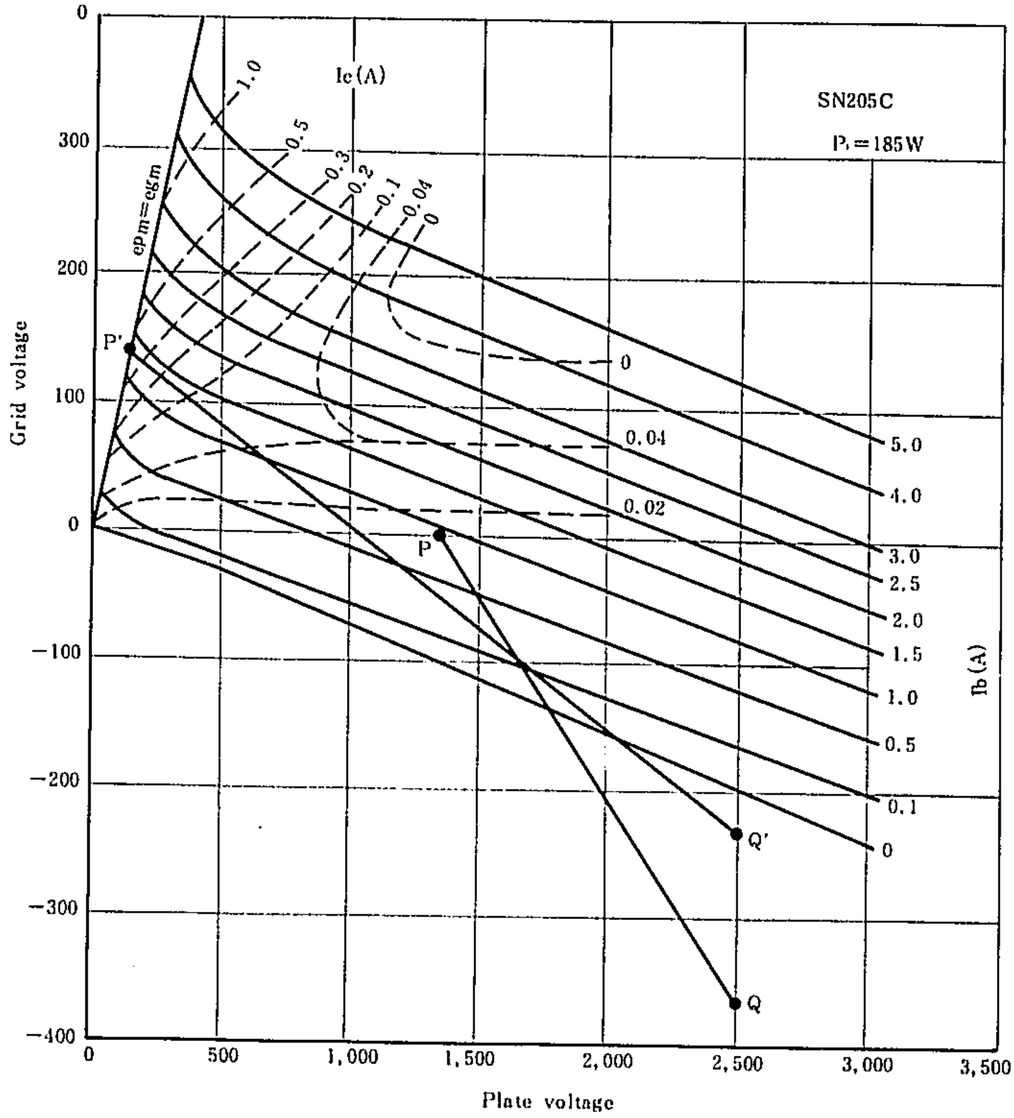


Fig. 53 Example of operation of class C grid modulation

Table 3 Analysis of SN205C grid modulation (Fig. 53)

			at carrier wave	100 % modulation positive peak	100 % modulation negative peak
Plate voltage	DC	E_b (V)	2,500	2,500	2,500
	Fundamental component	E_p (V)	1,180	2,360	0
Plate current	Peak	i_{pm}(A)	0.90	1.25	0
	DC	I_b (A)	0.17	0.38	0
	Fundamental component	I_p (A)	0.31	0.62	0
Plate current conduction angle		θ_p(Degree)	52	85	0
Output		P_o (W)	183	366	0
Input		P_i (W)	425	925	0
Plate efficiency		η_p (%)	43	40	0
Grid voltage	DC	E_c (V)	-370	-230	-570
	Fundamental component	E_g (V)	370	370	370
	Modulated wave	E_m (V)	-	140	200

In class C grid modulation, the voltage utilization factor is maximum at 100 % modulation positive peak, therefore, the procedure of design is to first set a resonable working condition and then determine the working condition at carrier wave. The distinctive feature recognised in analysis of this working condition is shown in Table 3. The amplitude of the modulated wave to obtain 100 % modulation is 140 V for the positive side and 200 V for the negative side. Fig. 54 indicates a graphical presentation of this condition, where the positive side of the modulated envelope extends and the negative side does not extend, thus causing vertical asymmetry, i.e. distortion.

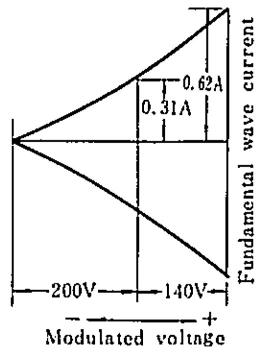


Fig. 54 Distortion due to grid modulation

This analysis, however, is obtained on the presumption of the excitation voltage is constant. Actually, since the grid current of the modulated tube flows during the positive side of modulation, the excitation voltage and the modulated voltage tend to decrease, resulting in reduction of distortion to some extent. However, even if the distortion is not compensated ideally by the grid current, there will be residual modulation distortion. For this compensation, the rectified negative feedback is effective.

6.7 Class C plate modulation

In class C amplifiers, if the plate DC voltage is changed while the grid excitation voltage and grid bias are kept constant, there will be a considerable linear portion in the high frequency plate voltage as shown in Fig. 55. Therefore, by superposing an audio signal onto the plate DC voltage, corresponding to the middle part of the linear portion, modulation will be available.

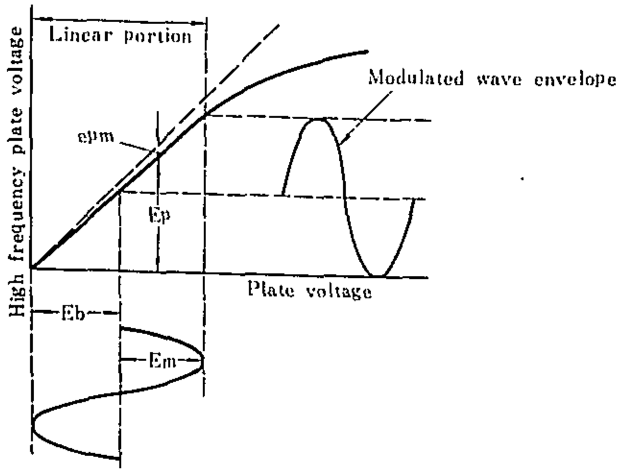


Fig. 55 Class C plate modulation

In class C plate modulation, excitation is provided sufficiently under no modulation state and adjusted so that the relationship between output voltage to excitation voltage will be in condition of saturation. Therefore, the plate voltage utilisation factor attains the maximum reasonable value as for class C power amplifiers. At the positive peak of 100 % modulation, both plate DC voltage and high frequency plate voltage becomes two times, so that the voltage utilisation remains the same as that of non-modulation state. The feature of class C plate modulation is that the voltage utilisation factor is high irrespective of the presence of modulation. For instance, it can be as high as 0.92. Therefore, the plate efficiency is excellent regardless of modulation. If we presume a plate current conduction angle of 60° during non modulation period, and expect a current utilization factor of 1.8, and plate utilization factor of 0.92, the plate efficiency factor will be,

$$\eta_p = 0.92 \times 1.8/2 = 83\ \%$$

which is approximately 2 times the efficiency of the grid molulation at carrier wave.

Fig. 56 indicates an example of operation of class C plate modulation. The modulator, is of a plate modulated 1 kW broadcasting transmitter specified by BTS, having two 5G31 tubes in parallel to generate a rated output of 1.1 kW at carrier wave. The working line PQ in the figure is completely identical to Fig. 44. At the positive peak of 100 % modulation, both DC plate voltage and high frequency plate voltage will be approximately two times of those at carrier wave therefore the output must become four times, resulting in working line P'Q'. From these characteristics, the result of analysis of the operating condition is given in Table 4.

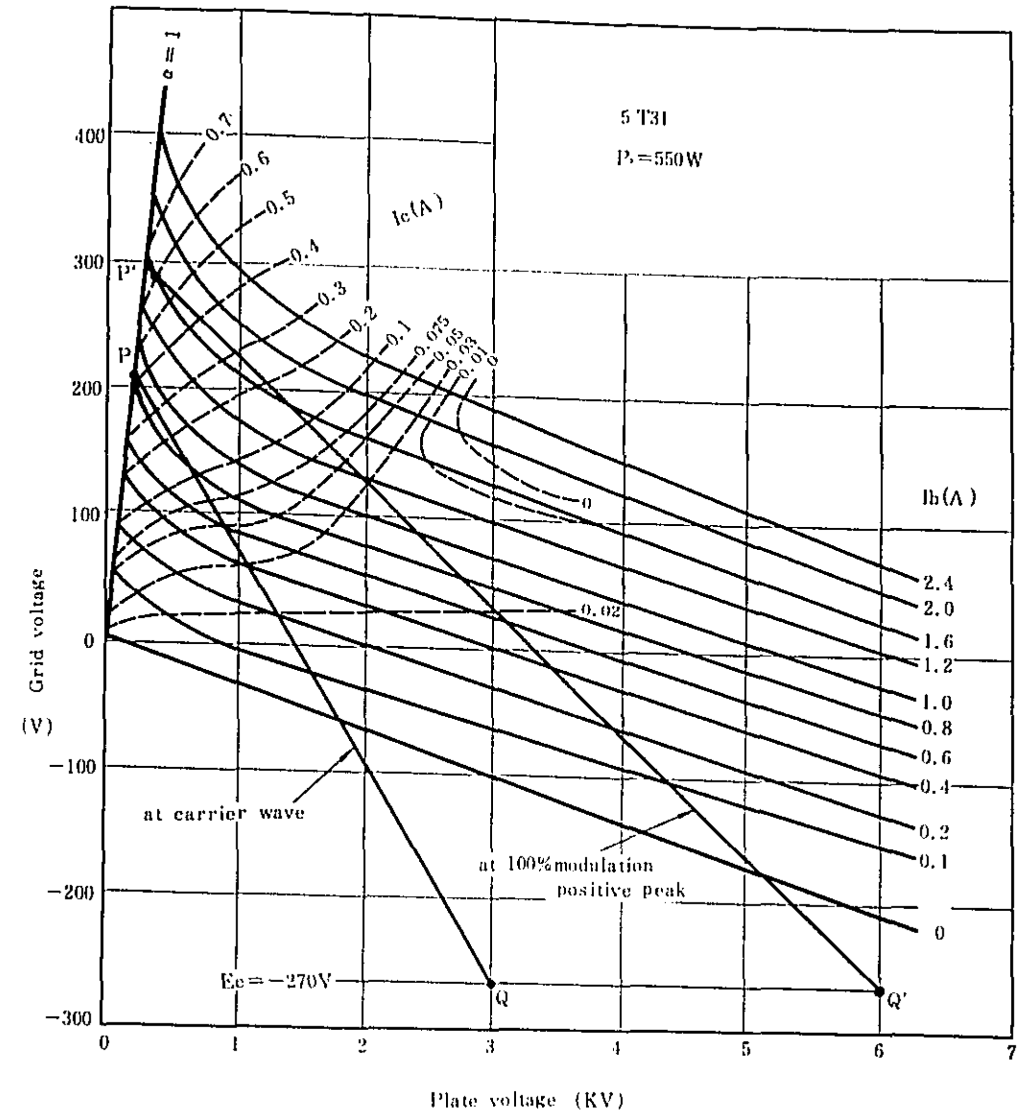


Fig. 56 Example of operation of class C plate modulation

Table 4 Analysis of 5T31 plate modulation (Fig. 56)

			at carrier wave	100% modulation positive peak
Plate voltage	DC	E_b (V)	3,000	6,000
	Fundamental component	E_p (V)	2,790	5,580
Plate current	Peak	i_{pm}(A)	0.9	1.6
	DC	I_p (A)	0.237	0.49
	Fundamental component	(A)	0.418	0.83
Plate current conduction angle		θ_p (Degree)	65	80
Output		P_o (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 component	E_g (V)	480	555
Grid current	DC	I_c (A)	0.06	0.09
	Fundamental component	I_g (A)	0.12	0.17
Excitation power		P_d (W)	28	48
High frequency load resistance		R_L (Ω)	6.680Ω (3,340 Ω per 2 tubes)	
Modulating transformer load resistance		R_m (Ω)	12,600Ω(6,300 Ω per 2 tubes)	

In determining the working line P'Q', it can be easily understood that, if the excitation voltage is left at the state of carrier wave, a four times output cannot be generated, resulting in difficulty of expansion of positive modulation side. This can be understood by analizing the characteristics. Therefore, as for point P ' P', it is necessary to make the grid voltage higher than that at carrier wave, as shown in Fig. 56.

There are two methods to realise this, as follows.

(1) To provide an exciation voltage initially, necessary for 100 % modulation positive peak.

(2) To increase the exctation voltage at 100 % modulation positive peak.

The first method is called "excess excitation". However, as grid curre

current flows drastically at carrier wave, limitation arises with regard to the grid loss and output of preceding stage. Thus the improvement of characteristics can not be deemed as the best. The second method is called the "preceding stage modulation" which is widely used. The preceding stage is modulated in the same phase of the modulated tube, but lower in degree. In this maner the excitation will increases at the positive peak of modulation. Although, the excitation will decrease at the negative peak of the modulation, but this does not cause any problem.

In the example of Table 4, excitation voltage is 480 V at carrier wave, and 555 V at positive peak, so the modulation degree for the preceding, stage will be as follows.

$$\frac{555 - 480}{480} \times 100 = 15.6 \%$$

As shown above, the preceding stage modulation reduces the modulation distortion prominently, and NHK widely applies this system.

The drawback of the class C plate modulation system is the necessity of large power for modulation. As already described in chapter 2, the mean high frequency power at 100 % modulation of a sinusoidal wave is 1.5 times of that without modulation. This increment must be supplied from the modulator. For instance, if a broadcasting transmitter of an output of 1.1 kW is 100 % modulated with sinusoidal waves, the mean-output will become 1.65 kW, the increment of 0.55 kW is to be supplied by the modulator. If we presume the plate efficiency of a modulated amplifier as 78 % (including losses of tank circuit etc.), then, to obtain a power of 0.55 kW, the modulator must supply the following amount of power to the modulation amplifier.

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

The reason why plate modulation is adopted, inspite of such drawback, is the excellent plate efficiency at carrier wave together with economy of power consumption.

6.8 Class B modulated wave amplification

There is a system in broadcasting transmitters where modulation is provided at the low-power stage and the modulated waves are amplified by high-frequency power amplifiers and then supplied to the antenna. In order to prevent distortion of the modulated waves envelope in this case, the method of amplification should be class B. In class B modulated wave amplification, the voltage utilisation factor increases in accordance to the modulation degree. Therefore, the voltage utilisation factor shall be set to a reasonable maximum value, so that it will not saturate even at the positive peak of 100 % modulation. Therefore, in case of non-modulation, the value becomes one half of it. If the voltage utilization factor at 100 % modulation is 0.92, it becomes one half at non-modulation, i.e. 0.46.

Since class B operation is involved, the plate current conduction angle is always 90° and the current utilization factor is constantly 1.57. Therefore, the plate efficiency at non-modulation is,

$$\frac{1.57 \times 0.46}{2} = 36 \%$$

which is lower than that of grid modulation.

In class B modulated wave amplification, distortion inherent to the amplification is relatively small. However, it is preferable to apply rectified negative feedback to cope with the modern technical requirements.

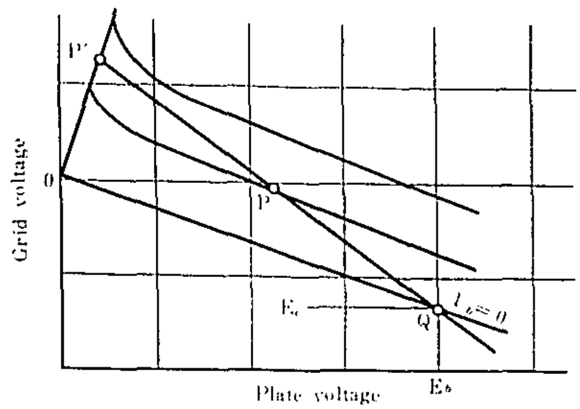


Fig. 57 Class B modulated wave amplification

Fig. 57 indicates an operation characteristic of class B amplification. Working line PQ relates to the case of non-modulation and P'Q to 100 % positive modulation. Because of class B operation, stationary point Q is located at the intersection between I_b = 0, E_b and E_c, and P' point is positioned on the extended line of QP.

In addition to the modulated wave amplification, the case of class B pushpull audio amplifier can be analysed by virtue of the constant current characteristics. This method is also more convenient than the E_b - I_b characteristics.

7. High-Frequency Output Circuit and Associated Circuits

7.1 High-Frequency Output Circuit

high-frequency power obtained by the power amplifier of the broadcasting transmitter is fed to the feeder or antenna through a high-frequency output circuit. Fig. 58 shows an example of the high-frequency output circuit consisting of a coupling circuit and a tank circuit.

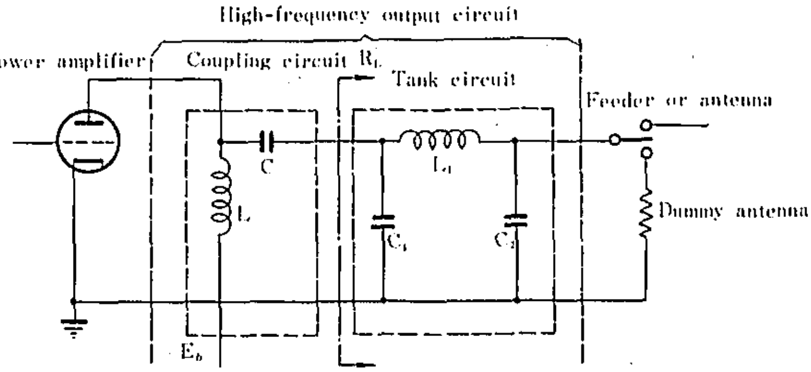


Fig. 58 Composition of a broadcasting transmitter output circuit

(1) Coupling Circuit

The function of the coupling circuit is to supply DC plate voltage to vacuum tubes and feed the high-frequency output power to the tank circuit. The coupling choke coil L becomes a DC circuit and prevents the high-frequency current from flowing into the power supply circuit. The coupling condenser C becomes a path for the high-frequency current and prevents the DC voltage from entering into the tank circuit.

A larger L and C can be considered to be entering advantageous, in view of the purpose of their functions, but if they are excessively large, inconvenience will occur. The L and C and the tank circuit will form a resonant circuit and if the resonance frequency approaches the audio frequency band, the frequency response will become deteriorate, the phase-angle will increase and affect the negative feedback function. Furthermore, it may cause unfavourable phenomena, such as modulation distortion and parastic oscillation. The C will resonate with the leakage inductance of the modulator transformer and deteriorate its characteristics at high frequencies. Therefore, the value adopted for L is 2 mH; and C is 1,000 pF for frequencies less than 1,000 Hz and 5000 pF for frequencies over 1,000 kHz.

(2) Tank Circuit

The function of the tank circuit is to convert the impedance of a feeder or antenna into the optimum load resistance required to the power amplifier and to transmit the high-frequency output of power amplifier to the feeder or antenna. In this case, the frequency-band to transmit should include not only the carrier frequency but also the side-bands caused by modulation.

Another function of the tank circuit is to suppress the frequency components effectively existing outside the band to be transmitted, i.e. higher harmonic occurring in the power amplifier. Generally, in the tuning circuits, the transmission of frequency band-width and suppression of harmonics are contrary to each other. In low power broadcasting transmitters this is of no problem, but in high-power transmitters, L C direct resonant circuits tuned to the second and third harmonics are often inserted in parallel to the output circuit of the tank circuit, to short-circuit the harmonics.

7.2 Design of Tank Circuit

When there exists a network N consisting of pure reactances as shown in Fig. 59, there is a general law that, if the impedance viewed from terminals 1 - 2 to the right is R_1, the impedance viewed from 3 - 4, to the left will be equal to R_2. In this case, impedance matching is performed at the terminals 1-2 and 3-4. Under this condition, if the network is divided at an arbitrary point A A', and the right and left side impedance measured, the resistance and reactance component will be equal in value, but the reactance will be contrary in sign to each other, i.e. series tuning. This condition is called "conjugate impedance matching".

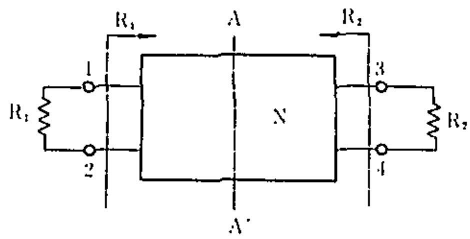


Fig. 59 Impedance matching

The tank circuit of broadcasting transmitters are utilising the nature of reactance networks described above. For the analisis of tank circuits, it will be very convenient to use the following conversion formula of series-parallel equivalent circuit.

Assuming that the parallel circuit of resistance R and reactance X shown in Fig. 60(a) is equivalent to the series circuit of resistance R' and reactance X' shown in Fig. 60(b), then relationship of these values will be given by the following equation;

$$\left.\begin{aligned} R' &= \frac{R}{1 + S^2} \\ X' &= R'S \end{aligned}\right\} \qquad (35)$$

wherein,

$S = \frac{R}{X}$ in the series circuit.

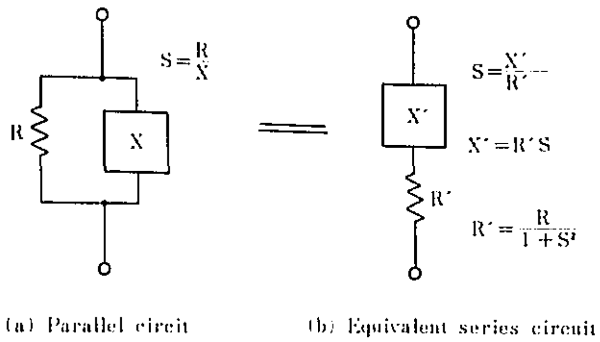


Fig. 60 Conversion from parallel circuit to series circuit

If

$$S = \frac{X'}{R'}$$

is promised in the equivalent series circuit, the value S will be equal to that of the S in the parallel circuit, i.e. will not be varied at all by the equivalent conversion.

Fig. 61 is the so-called L type circuit usually used for impedance matching, and it can be analized as follows by applying the above-mentioned parallel and series conversion equations;

(When resistance R_1 is connected to the input terminal, the impedance viewed on left hand of AA' can be replaced to an equivalent series circuit as shown in Fig. 61(b). To obtain conjugate matching at AA', the following equation should hold:

$$R' = R_2$$

wherein,

$$X' = X_L$$

The equation can be transformed as follows:

$$\left.\begin{aligned} S &= \sqrt{\frac{R_1}{R_2} - 1} \\ X_L &= R_2 S \\ X_C &= \frac{R_1}{S} \end{aligned}\right\} \qquad (36)$$

This matching circuit is used when R_1 is larger than r.

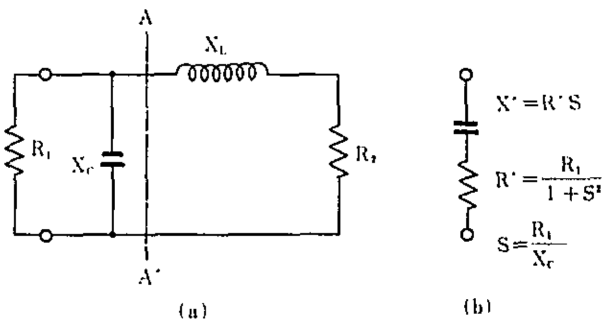


Fig. 61 L type matching circuit (R_1 > r)

The circuit of Fig. 62 is for the case that r is smaller than R_2, but as it will be the same to Fig. 61 with R_1 and R_2 replaced with each other, the following relationship will be obtained by replacing R_1 and R_2 with each other in equation (36).

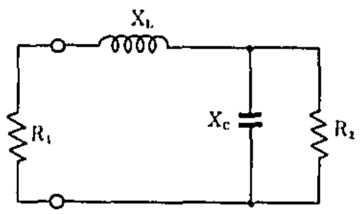


Fig. 62 L type matching circuit (r < R_2)

$$S = \sqrt{\frac{R_2}{R_1} - 1}$$

$$X_L = R_1 S$$

$$X_C = \frac{R_2}{S}$$

(In short, the values of circuits shown in Figs. 61 and 62 can be figured out by using equation (36), presuming that the one with a condenser is R_1 and the one without a condenser is R_2.

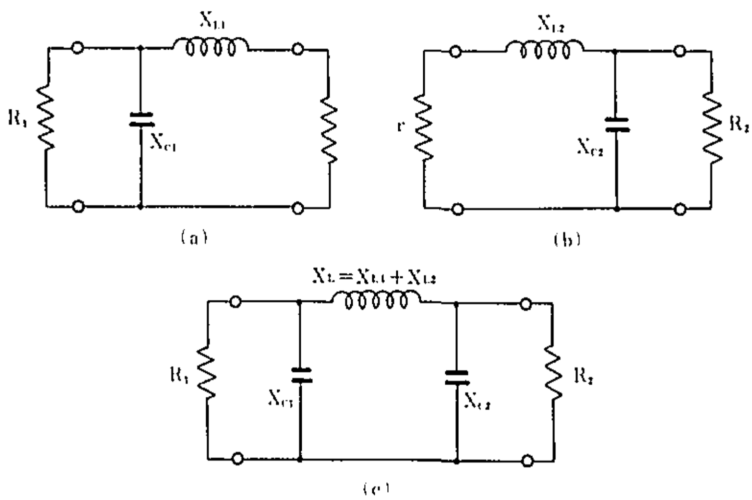


Fig. 63 π circuit

Fig. 63(a) and (b) show two L type matching circuits, assuming that R_1 matches r_1 and R_1 matches R_2, respectively. This matching condition is maintained even if the two circuits are connected with each other directly, as shown in Fig. 63(c). This circuit called the π type circuit and is used in the final-stage tank circuit of broadcasting equipment substantially without exception.

If $$S_1 = \frac{R_1}{X_{c1}} \qquad S_2 = \frac{R_2}{X_{c2}}$$

are assumed in the π circuit of Fig. 63 the relationship:

$$\gamma = \frac{R_1}{1 + S_1^2} = \frac{R_2}{1 + S_2^2}$$

exists wherein:

$$X_{L1} = \gamma S_1 \qquad X_{L2} = \gamma S_2$$

Therefore, if R_1 and R_2 are given first and by selecting a proper value for S_1, the necessary values for designing the circuit can be calculated as follows,

$$\left.\begin{aligned}
\gamma &= \frac{R_1}{1 + S_1^2} \\
S_2 &= \sqrt{\frac{R_2}{\gamma} - 1} \\
X_{c1} &= \frac{R_1}{S_1} \\
X_{c2} &= \frac{R_2}{S_2} \\
X_L &= X_{L1} + X_{L2} = \gamma(S_1 + S_2)
\end{aligned}\right\} \quad (37)$$

Current following to L = $\sqrt{\frac{P_0}{r}}$
Voltage of X_{c1} = $\sqrt{R_1 P_0}$
Voltage of X_{c2} = $\sqrt{R_2 P_0}$
Herein, P_0 is high-frequency output.

Then, what does the S mean which was used in the heretofore described calculation? Assuming that a certain voltage is applied to a resistor and reactive parallel circuit, or a certain current is conducted in a series circuit, then it can be understood that the following relationship will hold.

$$S = \frac{\text{Reactive power}}{\text{Active power}}$$

When S is large in a tuning circuit, the tuning will be sharp and harmonic suppression will be good, but the band-width will become narrow and cause attenuation to side-bands at higher range of modulation frequencies. The reactive power grows excessively large so that the condenser and inductance are required to be large in kVA. Therefore, S is usually fixed to about 5 at low frequency band and to about 10 at high frequency band (Refere to Chapter 1, Table 1).

Impedance matching is possible even when a tank circuit and a load is coupled electromagnetically, not with π circuit. (the π circuit, however, is superior to electromagnetic coupling in removing higher harmonics.) The π circuit is extensively employed in broadcasting equipment since its advantage is such that it can be coupled directly with an unbalanced feeder such as coaxial cables.

[Examples of Calculation]

When the load resistance, load and output of a high-frequency amplifier tube are respectively 3,340Ω, 350Ω and 1.1 kW, the values of tank circuit can be calculated as shown below, assumin its angular frequency as $w = 5 \times 10^6$ (f = 800 kc)

The influence of the coupling condenser between the vacuum tube and the tank circuit is slight. If this is neglected, then, in Fig. 63, the relation of R_1 = 3,340Ω, R_2 = 350Ω will hold. Assume the S_1 of this circuit to be 8, then according to equation (37) the calculation will be as follows,

$$\gamma = \frac{3,340}{1 + 8^2} = 51.4\Omega$$

$$S_2 = \sqrt{\frac{350}{51.4} - 1} = 2.41$$

$$X_{c1} = \frac{3,340}{8} = 417\Omega \qquad C_1 = \frac{1}{5 \times 10^6 \times 417} = 480 \text{ pF}$$

$$X_{c2} = \frac{350}{241} = 145\Omega \qquad C_2 = \frac{1}{5 \times 10^6 \times 145} = 1,380 \text{ pF}$$

$$X_L = 51.4\ (8 + 2.41) = 536\Omega \qquad L = \frac{536}{5 \times 10^6} = 107\ \mu\text{H}$$

Current flowing in L $= \sqrt{\frac{1,100}{51.4}} = 4.63\text{A}$ (at wave carrier)

Voltage of $C_1 = \sqrt{3,340 \times 1,100} = 1.92$ kV (at wave carriage)

" $= 1.92 \times 2\sqrt{2} = 5.43$ kV (100% modulation peak value)

kVA of $C_1 = \frac{(1.92 \times 10^3)^2}{417} = 8.85$ kVA (at wave carriage)

" $= 8.85 \times 1.5 = 13.3$ kVA (at continuous 100% modulation)

Voltage of $C_2 = \sqrt{350 \times 1,100} = 0.621$ kV (at wave carrior)

" $= 0.621 \times 2\sqrt{2} = 1.76$ kV (100% modulation peak value)

kVA of $C_2 = \frac{(0.621 \times 10^3)^2}{145} = 2.66$ kVA (at wave carrier)

" $= 2.66 \times 1.5 = 4$ kVA (at continuous 100% modulation)

In determining the actual voltage of condensers, the endurable voltage and kVA at 100 % modulation, as indicated above, should be adopted. In the calculation of C_1 and C_2, the stray capacity was not taken into account. Therefore, in practical 1 a proper amount should be substracted from the calculated values.

7.3 Neutral Circuit

The most distinctive difference between triodes and multielectrode tubes are the amount of electrostatic capacity between the grid and plate electrodes. The capacity of multielectrode transmitter tubes are in the order of 0.1 pF, while the triode transmitter tubes are in the range from several pF to scores of pF.

In tube amplifiers, the amplified output voltage is fed back to the grid through the electrostatic capacity between grid and plate. The input impedance of the amplifier is therefore influenced by the amplification degree and the capacity between grid and plate. The relationship is expressed in the following equation:

Input resistance $$R_g = \frac{1}{A \sin\theta} \cdot \frac{1}{WC_{gp}}$$

Input capacitance $$C_g = C_{gk} + C_{gp}\,(1 + A\cos\theta) \qquad (38)$$

Herein, $$A = |E_p/E_g|$$

θ: leading angle between Eg and (-)Eg

(θ is positive at inductive load.)

The above equation (38) obviously shows that the input impedance will be infinite when the load is pure resistance, and become negative when the load is inductive and become positive when the load is capacitive. The meaning that input resistance is positive is that energy is transferred from grid to plate, and negative resistance is that the amplified energy is transferred from plate to grid, i.e. positive feedback. If this negative resistance becomes lower than the positive resistance of the grid circuit, it will match the condition of oscillation.

When a tuning circuit is used for the plate load, the relationship between frequency and input impedance is shown in Fig. 64.

When adjusting the tuning of plate tank circuit, if the tuning frequency is lower than the excitation frequency, the excitation load is heavy because the input resistance is positive, and if the tuning frequency is higher than the excitation frequency, the excitation load is light, because the input resistance is negative, and in some cases, it may oscillate. In case the tuning of tank circuit is re-adjusted, the input capacity will change and preceding stage will be detuned.

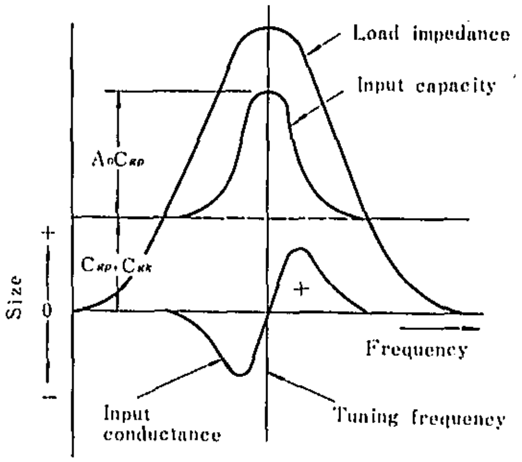


Fig. 64 The impedance of a vacuum tube loaded with a plate parallel tuning circuit. Ao represents amplification degree at tuning frequency.

In order to eleminate this influence, a neutral circuit is necessary for triode amplifiers.

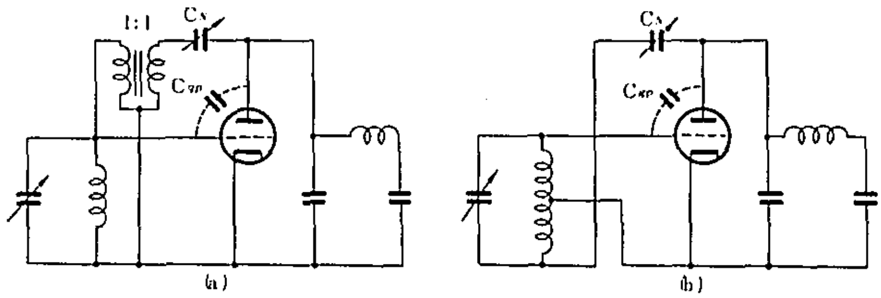


Fig. 65 Neutral circuit

Fig. 65(a) shows a method of feeding the output voltage in reverse phase to the grid, by using an iron-core high-frequency transformer, with winding ratio of 1 : 1. Fig. 65(b) shows a method of feeding the output voltage to the preceding stage tank circuit, in reverse phase to the grid. In both methods, neutrality will be accomplished by adjusting the neutralizing condenser C_N, so that it will be equivalent to C_{gp}. In comparing the method (a) with (b), the former is advantageous because the primary and secondary coupling factor is large, and complete reverse phase voltage can be induced stably in the grid, covering a wide frequency range.

In order to adjust the neutrality, apply excitation to the grid under condition of plate DC current supply off, and observe the high-frequency voltage induced in the plate with a peak voltmeter or cathode-ray tube, as adjusting the neutrality condenser to the minimum voltage point.

In case the neutrality is not adjusted completely, at the modulated tube, a partial of the excitation voltage will appear in the output at positive 100 % modulation peak, and the waveform will become distorted.

7.4 Prevention of Parasitic Oscillation

When a broadcasting transmitter oscillates at a frequency other than that of the normal transmission frequency, it is called "parasitic oscillation". It is an oscillation which will occur at various frequencies, dependent on condition of occurence.

In some cases, parasitic oscillation becomes so violent that it ignites the vacuum tube or destroys insulation due to abnormal high voltage or interrupts the normal operation of broadcasting transmitter. In some cases, it occurs only at the modulation grid period, and the cause will be very difficult to investigate.

To avoid the parasitic oscillation, a prevention circuit is furnished at places where it is liable to occur. A neutralising circuit is one of the prevention circuits. In general, however, a small L and R is inserted in parallel with the grid and plate terminals, to prevent parasitic oscillation in the VHF band.

With broadcasting transmitters provided with negative feedback, if the design of feedback is inappropriate, or the amount of feedback is excessive, super-audiable frequency oscillation will occur and radiate side-bands beyond the designated wave-band. Care must be taken because they will be very difficult to discover.

7.5 Monitor-detector

A monitor detector is a device to detect and pick up the DC and audio component out from the final stage of transmitter, by a pickup coil coupled to the tank circuit of the final stage. The DC component is used for supervision of carrier shift, sensor of carrier-breakdown, indication of modulation meter and measurement of modulated characteristics (distortion, SN ratio, etc.). Therefore, the device should have an excellent linearity characteristic. Its frequency response, distortion and SN ration characteristics, should be much better than those of the broadcasting transmitters.

As for linear detectors, full-wave rectifiers of diode-tube or germanium diodes are used. Fig. 60 shows an example of the monitor detector employing germanium diodes.

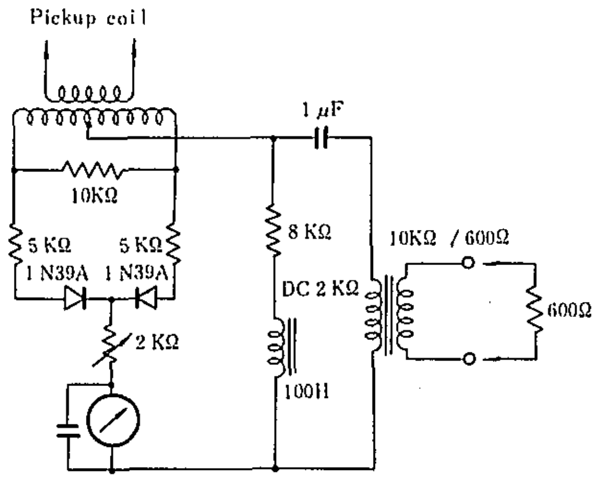


Fig. 66 An example of monitor detector

The high-frequency input circuit of this example is an untuned type, but it is available to make it a tuned type by providing a varaiable consdenser in it. In this case, a resistor should be added to reduce the Q of the circuit, sufficiently, otherwise the sidebands will be attenuated. Further, the circuit should be tuned accurately to the center frequency, otherwise there will be distortion occuring, due to the um-balance of sidebands. The reason why a resistor is inserted in series to the detector is to improve its linearity.

Next, the DC resistance of the load must not be higher than the impedance of the AC component, in respect to distortion. In case the DC resistance is higher, the top of waveform will be cut when modulation degree is increased and become the cause of distortion. This kind of distortion is called "clipping distortion". The same phenomenon occurs in audio amplifiers when it is over loaded and waveform is cut-off due to the cut-off of plate current. In the example of Fig. 66, both impedance are arranged so that they will be equivalent.

8. Proper Degree of Modulation

When testing a broadcasting transmitter with a modulation of continuous sinudoidal wave, the modulation degree will be determined in accordance to the input level. However, as the leve of an actual programme is always varying, how could the modulation level be set.

In broadcasting transmitters, the limit of modulation is 100 %. If the input exceeds this level, distortion will occur. If the maximum level during the programme is set at 100 % modulation to avoid this distortion, the average modulation degree of the whole programme will become considerably low and will result in low volume at receiver set and the SN ratio will be deteriorated extremely.

Further, if the input volume control of the transmitter is adjusted so that the high and low programme levels will be always 100 % modulation, the programme output will become smooth. But, as a programme consists of high and low levels and the programme staff in the sub-control room are adjusting the programme level, intentionally, other person are not allowed to adjust the level. Therefore, only adjustment of the whole programme level is permitted at places out-side of the sub-control room.

To deepen the modulation degree of a broadcasting transmitter by 3 dB is to increase the antenna power 3 dB, namely the similar effect of two folds. Therefore, it can be realised that deepening of modulation degree has a significant meaning to increase the output power, compared with the difficulty of increasing the antenna power.

NHK has established a standard reference level, for adjusting the input and output levels of sub-control room, master control room and outgoing transmission lines, and equipment, etc. This standard level is represented by the time-signal(880 Hz) or 1,000 Hz sine wave at the morning test. When VU meters are to indicate this level, they are adjusted by the attached attenuator to (-) 2 VU.

For actual programme levels, the peak level is read by this VU meter. For musical programmes, the meter is to exceed the 0 VU point, in an average, once or twice a minute, and for speech programmes, several times a minute.

However, even if the meter indication of programme level may be "0" or (-) 4 VU, it can not instantaneosly follow thd current peaks because of the inertia force. The actual level of peak current is to be considerably higher than the reading of the meter. This difference is called the "peak factor", and is estimated as 15 dB, in respect to statistical data. This means that the overall sound system should be designed so that it will not overloaded at levels exceeding the standard level by 15 dB.

In considering the above, the modulation degree of NHK's broadcasting transmitters are set as following,

(1) Transmitter output power exceeding 10 kW.

Adjust the limiting point of the limiting amplifier to the standard level, and set the modulation degree according to the following;

3 ML-1 limiter	80 %
2 ML-1 limiter	90 %

(2) Transmitter output power less than 10 kW.

(a) When using limiting amplifier
Regardless of the type of limiting amplifier, regulate the modulation degree to 90 %, an increment of 4 dB.
Adjust the limiting point of the limiter to 90 % modulation, and set the input volume control so that the standard input level will be 4 dB higher than the limiting point.

(b) When not using limiting amplifiers.
Set the standard level to 80 % modulation. Fig. 67 shows the relationship between input level and modulation degree, of a limiting amplifier of which the modulation degree is set in accordance with the above stated rule.

Curve (a) corresponds to transmitters of power exceeding 10 kW. Assume that the compression ratio of the limiting amplifier is set at 20/1.5 dB, then it can be seen that there will no over modulation at most of the part, but slight over-modulation, at levels exceeding the standard level by 15 dB.

Curve (b) corresponds to transmitters of power less than 10 kW. As the limiting action will start at a level 4 dB lower than the standards level, there will be a little more over-modulation than the above case(a).

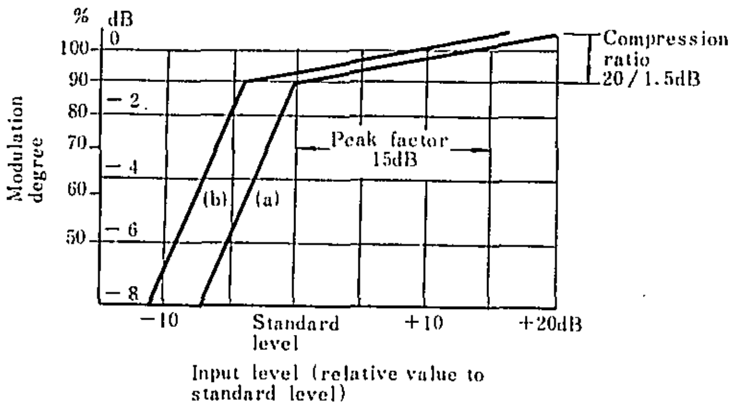


Fig. 67 Modulate on degree obtained when a limiting amplifier is used

Fig. 68 shows the modulation degree when a limiting amplifiers is not used. Assume that the modulation is 80 % at the standard level, then, if the input level is raised only 2 dB, it will become 100 %, and most, of the peak factor component will become over-modulated and result in severe distortion. If the modulation at peak level is to be strictly suppressed, the modulation of standard level should be 20 %. Comparing this with Fig. 67, the modulation can be deepened by 13 dB if a limiting amplifier is used.

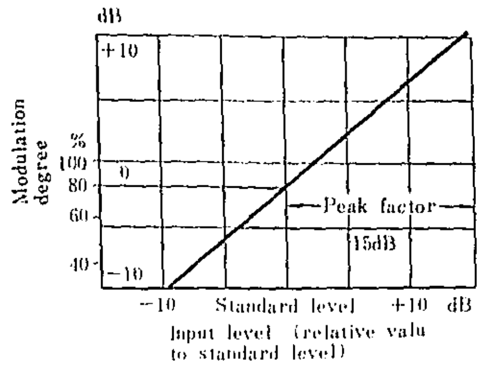


Fig. 68 Modulation degree when no limiting amplifier is used

9. Transmitter Tubes

The vacuum tubes used for power amplification of transmitters are named transmitter tubes. The principle of operation is not substantially different from receiver tubes, but, as the power dealed with is large, the construction, way of handling differs.

9.1 Vacuum Tubes

The transmitter tubes consist of triodes, tetrodes, beam tubes and pentodes, similar to receiver tubes. Triodes are used for power up to a very large extent because of the simple structure. There are water-cooled tubes of allowable plate-loss exceeding several scores of kW, which the plate is cooled by water. As for tetrodes and tubes with more electrodes, the structure becomes complicated, and the power to deal with becomes low. The upper power limit of pentodes is about 1 kW.

The material used for the cathode depends on the working voltage of the tube. For plate voltage exceeding 10 kV, pure tungsten is used, and for voltage under 10 kV, thorium tungsten is used. Oxcide covering is used for voltage under 10 kV.

The structure of filaments are made so that the magnetic field caused by their current will not induce hum. There is a middle tap in the filament so that it can be heated with current in different phases, or a double-spiral-winding to cancell out the magnetic field.

Grids employ molybdenum or tungsten wires. In the case of thorium tungsten filaments and oxide-covered cathodes, grid wire is plated with platinum to prevent emission of thermions therefrom and baked with zirconium, so that it will not obtain dynatron characteristic, i.e. secondary electron emission, when grid has positive electricity potential and grid current flows.

Plate loss and heat value increase in proportion to the size of tubes. Therefore, special consideration is required for cooling them. Natural-air-cooled tubes diffuse most of the plate-loss as heat radiation,' through their walls. Heat radiation is increased by attaching radiation fins to plate, blackening the surface, applying sun-blast treatment or zirconium baking. Nickel molybedenum and other materials are used for the plate. The utmost limit of natural air-cooling is for tubes with aelowable plate-loss up to 1 kW. For tubes exceeding this limit, forced air cooling is required.

Forced cooling type tubes have a copper-made plate which constitutes a part of the vacuum container. Water-cooled tubes have such a structure that the cooling water passes round the plate inserted into the jacket as shown in Fig. 69(a). Forced-air-cooling tubes have many radiation fins so that cooling may be performed by an air blower as shown in Fig. 69(b). The copper plate and the glass part, as a vacuum container, are required to be conglutinated hermetically. Because of the difference in thermal expansion coefficient, a metal called koval which is equal to glass in expansion coefficient, is used at the joint part or a copper-plate made extremely thin like a knife-edge is used to set the mechanical force free caused by the expansion and contraction of glass.

Such a metal as magnesium is usually deposited on inside of tube wall, as a getter, for adsorbing the gas particles, to maintain the vacuum well after evacuating the vacuum tubes. But, this method is used only for vacuum tubes having oxide covered cathode and thorium tungsten filaments. In the case of pure tungsten filaments, no getter is employed since the tungsten itself evaporates and functions as getter. As for water-cooling and forced- air-cooling tubes, it is difficult to adhere getter to the glass wall, therefore, a metallic getter such as tantalum and zirconium is used.

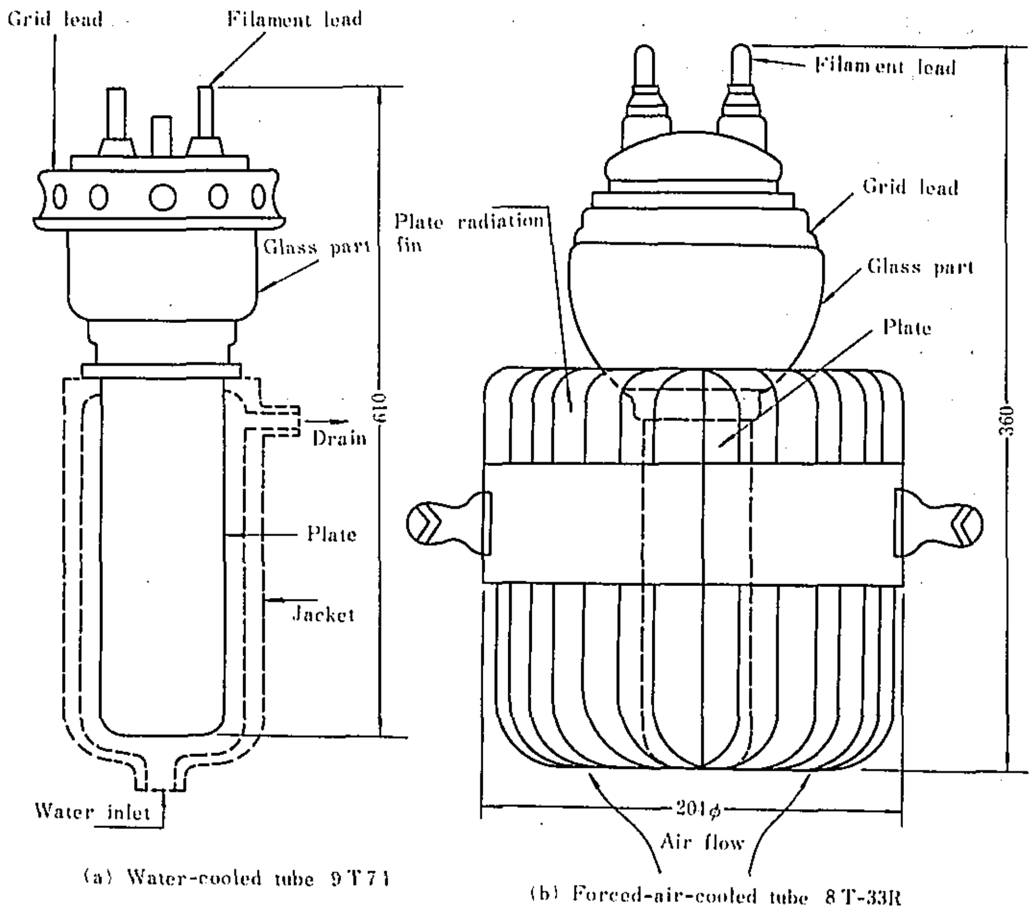


Fig. 69 Transmitter tube cooling systems

9.2 How to Use Transmitter Tubes

The details of the procedures for using and maintaining vacuum tubes are described in the "Vacuum Tube Maintenance Procedure" of the Regulations. Herein stated, therefore, are only their important points requiring special attention. Transmitter and special receiver tubes are required to be collectively ordered and supplied by the Head Office according to the "Vacuum Tube Supply Procedure". The stations, therefore, should take the prescribed procedures for securing the required number, thereof.

Vacuum tubes in hand should be used in principle until their life guaranteed period expires at the earliest if such a guarantee has been given and, thereafter, in the order of their manufacture as long as they remain usable.

Vacuum Tubes

(1) Starting

To start vacuum tubes, first heat their cathodes sufficiently and then impress voltage on grid, plate and screen-grid in the order mentioned. To stop them, reverse this procedure.

In the case of water-cooled and forced-air-cooled tubes, feed water (air) before switching on filament power supply and, when stopping the tubes, cut filament power and after there is no residual heat, switch off water air power supply.

(2) Filament Voltage

The filament voltages of oxide-covered cathodes are required to be used at 90 - 100 % of their ratings. Do not leave the filament ignited without conducting plate current, as a semiconductive intermediate layer will be formed between core metal and oxide substance and deteriorate the cathodes.

The thorium tungsten filament is carbonized for reducing thorium evaporation. If the tube is used for a long time, the carbonation degree will gradually become low and reduce the filament resistance. If the filament voltage is constant, the heating power will increase in proportion to the decrease of the carbonation degree and more thorium will be evaporated and result in rapid reduction of radiation. Therefore, it is desirable to use the filament voltage at a lower voltage within the allowable range of output power and distortion of the broadcasting transmitter, since there exists surplus radiation current. In large sized tubes, when the filament is heated, a current several times of the normal value will flow, and this current will provide an electromagnetic force and deform the filament or break the filament or become the cause of inter contact of electrodes. Therefore, to prevent this sudden flow of current, a limiting resistance or leakage transformer is used. The latter is convenient because constant current heating can be performed to avoid current increase due to de-carbonization.

(3) Cooling

The plate loss of vacuum tube is converted into heat and raises the temperature. If cooling is improper, the vacuum degree deteriorates, since the gas inside will be discharged due to the excessive rise of plate temperature or slow-leak may occur from the sealing part because of the difference of expansion coefficient between glass and metal. Special consideration is required also in cooling large-size tubes, because their grid-loss is large,

As for water-cooled tubes, their jacket is at a high potential Therefore, a snake pipe of porcelain, acryl or other material are inserted between the jacket and the water-pipe for insulation. At the joint-point of the snake pipe, metal will be electrolysed and carroded by the leakage current. To prevent this, a lead or stain-less rode is attached to the joint-point. Where the quality of water is bad, fur accumulates and leakage current will become excessive. In this case, the water is to be distilled and cooled by circulation, and then futher cooled by the water of a well. The flow of water must be maintained as specified, and the temperature should be held less than 60°C at the outlet. If necessary, the place where electrodes are drawn out must be air-cooled to maintain the temperature of each part less than the maximum specified value.

In the case of forced-air cooled tubes, the air-flow must be as the prescribed rate (2.2 - 3.5^3/min, per 1 kW of plate loss) and the maximum temperature of each part must not exceed the prescribed value. If the cooling air contains dust, equipment will be fouled and cooling effect will be reduced by the dust adhered to the radiation fins. To prevent this, filter is to be mounted on the air inlet.

In the natural-air cooled tubes, the large sized ones are used under a condition until the plate gets ignited. Therefore, the cooling should be provided not only to the tube but also to the parts in the vicinity.

(4) Flash

A flash occurs mainly between anode and cathode, due to low vacuum degree, and excessive large current flows occasionally to the grid, and often melts the grid. The flash may occur accidentally and the tube may remain usable and in some cases, the flash may continue and the tube will become unusable. In many cases, the vacuum tube will be lefined by the clean-up function of the flash.

(5) Protective Circuit

When the load is substantially shorted or the vacuum tube flashes, an excessive large plate current will flow. In order to prevent the tube and power supply from damaging, an overload relay is inserted in the cathode circuit of the vacuum tubes to cut off the power supply imediately. In water-cooled and forced-air cooled tubes, when the water-flow is interupped during operation, a relay activated by water air cooling system, cuts off the plate and filament power supply immediately.

10. Power Supply

10.1 AC Power Supply

The power source of broadcasting transmitters are required to be stable in frequency and voltage. Particularly, the voltage variation has direct relation with the antenna power and distortion, therefore, an automatic voltage regulator is inserted in the incoming electric line to maintain the output voltage variation of (±) 10 % within the range of (±) 1 %.

In plate modulation transmitters employing class B pushpull modulation, the power consumption of transmitters vary in proportion to the strength of audio signal. Therefore, in order to keep the carrier wave variation in an excellent condition, it is desirable to maintain the voltage variation at 100 % modulation to a value less than 1.5 % of the non-modulation variation state.

With regard to the filament power source for vacuum tubes of broadcasting transmitters, the adoption of indirect heating tubes and thorium tungsten filaments for large power tubes, remarkably reduced the hum level and by the application of negative feedback, majority of the filament tubes can be heated with AC power.

10.2 Rectifiers

The DC for energizing grid-bias, screen-grid and plate is all supplied from rectifiers. For the rectifiers, silicon elements are used and various rectifying circuits are employed. The circuits are compared in Table 5.

(1) Single-Phase Full-Wave Rectifier Circuit with Central Tap

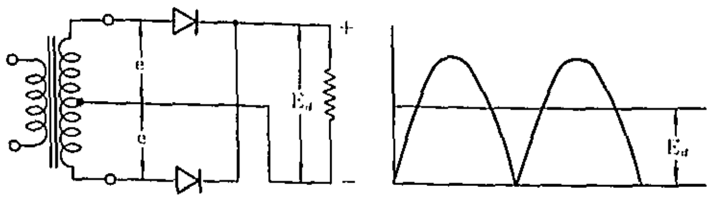


Fig. 70 Single-phase full-wave rectifier circuit with central tap

This circuit consists of two sets of half-wave rectifier circuits connected in parallel, as shown in Fig. 70 (a), and one of them rectifies the positive half-cycle while the other one rectifies the negative half-cycle, and produces an output voltage waveform as Fig. 70 (b), when the load is pure resistance. If the voltage between the secondary side terminal and neutral point of the transformer is e (effective value), the DC output voltage will be E_d and the inverse peak voltage applied to one rectifier is E_i, then the following formula holds,

$$\left.\begin{array}{l} E_d = 0.9e \\ E_i = 3.14E_d \\ \text{Transformer secondary kVA} = 1.57 \times \text{DC output} \\ \text{Transformer primary kVA} = 1.11 \times \text{DC output} \end{array}\right\} \quad (39)$$

This circuit is not suitable for rectifying high voltage large output power. Since the inverse voltage applied to the rectifier is very high, the kVA of transformer will become large and a large amount of AC voltage is required in comparison to the output voltage. However, as the structure is simple, it is used for low-voltage, low-output power applications. It requires a larger smoothing device as the frequency is low and the amount of ripple frequencies contained in output is considerably high.

(2) Single-Phase Bridge Full-Wave Rectifier Circuit

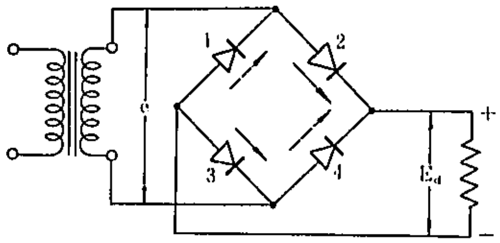


Fig. 71 Single-phase bridge full-wave rectifier circuit

The rectifier elements are connected into a bridge as shown in Fig. 71. At the positive half-cycle, element 2 and 3 work in series and at the negative half-cycle, element 1 and 4 work in series. The output waveform is equivalent to the full-wave with center tap. The voltage relationship can be expressed by the following formula,

$$\left.\begin{array}{l} E_d = 0.9e \\ E_i = 1.57E_d \\ \text{Transformer secondary kVA} = 1.11 \times \text{DC output} \\ \text{Transformer primary kVA} = 1.11 \times \text{DC output} \end{array}\right\} \quad (40)$$

The advantage of this rectifier circuit in comparison to the center tap full-wave type is that, as two rectifier elements work in series, tap full-wave type is that, as two rectifier elements work in series, the inverse voltage becomes half, and the transformer secondary-voltage becomes low and the kVA value of transformer could be reduced.

(3) Three-Phase Half-Wave Rectifier Circuit

As shown in Fig. 72, the single phase half-wave rectifiers are connected in three-phase form with output side connected in parallel

$$\left.\begin{array}{l} E_d = 1.17e \\ E_i = 2.09E_d \end{array}\right\} \quad (41)$$

$$\left.\begin{array}{l}\text{Transformer secondary kVA} = 1.48 \times \text{DC output} \\ \text{Transformer primary kVA} = 1.21 \times \text{DC output}\end{array}\right\}$$

The pulsation frequency of this circuit is three times of the power supply frequency, and as the degree of ripple contained is low, the smoothing circuit is simple. However, the inverse voltage applied to the rectifier element is high and as the secondary voltage of transformer is also high, it is used only for low power.

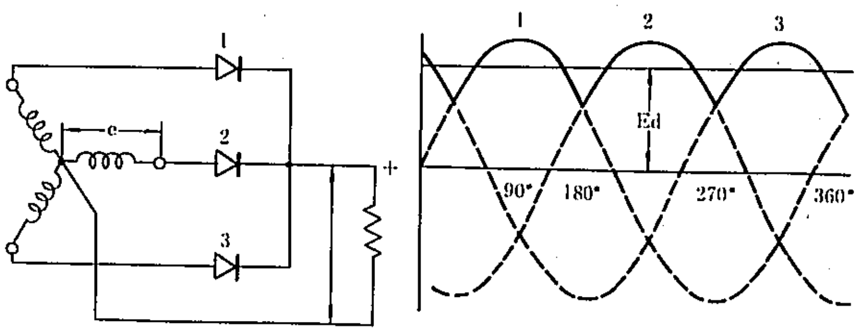
(4) Three -phase Full-wave Rectifier Circuit

If six rectifier elements are connected as shown in Fig. 73(a), the output voltage wave will be as shown in Fig. 73(b). The capital letters, such as A and B on waveform, indicate that the voltage between A and B is rectified and the figures 2 and 4 indicate that current flows through element 2 and 4, in series.

$$\left.\begin{array}{r}E_d = 2.34e \\ E_i = 1.047E_d \\ \text{Transformer secondary kVA} = 1.05 \times \text{DC output} \\ \text{Transformer primary kVA} = 1.05 \times \text{DC output}\end{array}\right\} \quad (42)$$

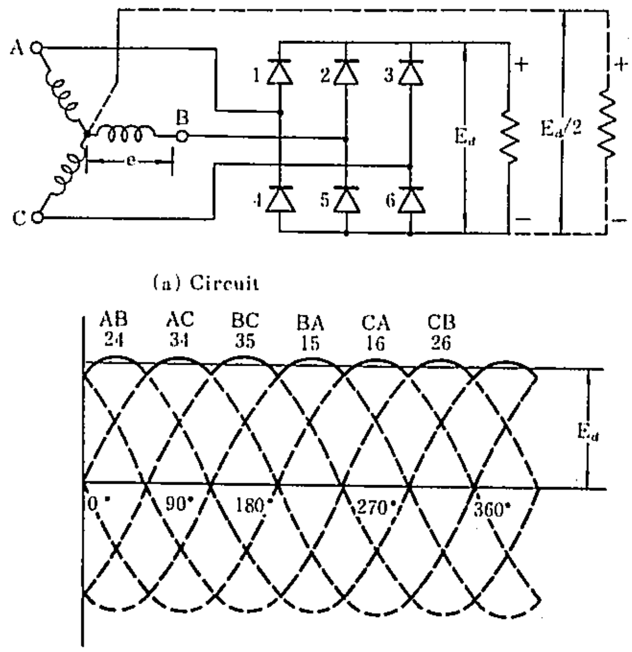
The pulsation frequency of this circuit is six times of the power source frequency, and as the pulsation voltage is low, the smoothing circuit is very simple. With regard to inverse voltage, the value will be half of that of half-wave rectifiers, since two rectifier elements are connected in series. The kVA required for the transformer is the lowest than any other system existing.

The three-phase full-wave rectifier circuir thus having many advantageous features, is employed for output power exceeding 1 kW. As shown in the broken line of Fig. 73(a), a three-phase half-wave rectifier circuit is composed between the neutral point of transformer and (-) terminal, and as it is able to draw out voltage half of the full-wave rectifier, this voltage is usually used for the plate voltage of the preceding stage.



(a) Circuit (b) Output voltage

Fig. 72 Three-phase half-wave rectifier circuit



(a) Circuit

(b) Output voltage

Fig. 73 Three-phase full-wave rectifier circuit

10.3 Smoothing Circuit

The smoothing circuit employed in broadcasting transmitters is not only for smoothing pulsation voltage but the impedance viewed from the load side must be low and also capacities. This is because to avoid audio component to be contained in the load current, when modulation is provided. It is therefore necessary to insert a condenser in the load circuit.

In the smoothing circuit, there is a condenser input type, which a condenser is inserted in the input side, and a choke input type, where an input condenser is negelected and directly connected to the choke coil.

(1) Condenser Input Type

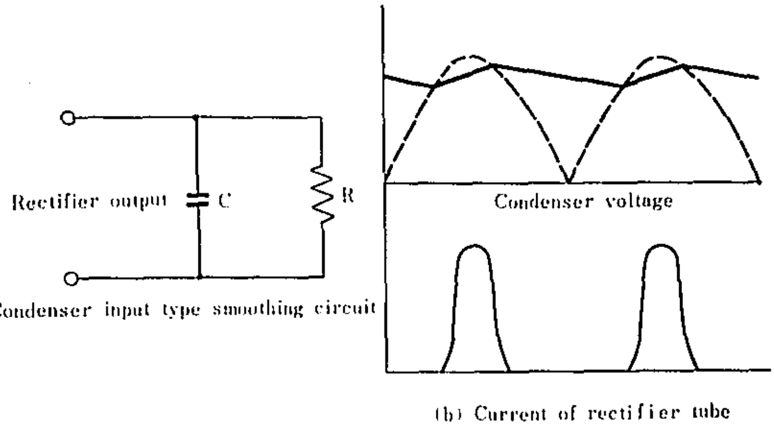


Fig. 74 Condenser-input smoothing circuit and voltage and current

As shown in Fig. 74(a), a smoothing condenser is connected in parallel to the input terminal of the smoothing device, and its terminal voltage is given in Fig. 74(b). Namely, the plate current will flow only during the period that the supplied AC voltage exceeds the terminal voltage of the condenser, and the condenser will be charged to a voltage of a value close to the peak value of the AC voltage.

When the flow of the plate current stops, the condenser will be discharged through the load, therefore, the voltage will decrease linearly, until the next charge action starts.

Assuming that the ratio between the pulsation fundamental effective value of the input condenser terminal and DC value, namely, the pulsation factor is δ_1, and the load resistance R (Ω), condenser capacity C(F), and angular velocity of pulsation fundamental frequency ω, then the following formula will hold,

$$\delta_1 = \frac{\sqrt{2}}{\omega CR}$$

Namely, the pulsation factor becomes smaller in proportion to the increase of value of C or R.

In case of the condenser input type, the condenser will be charged to a value close to the peak value of the supplying voltage, thus enabling a high voltage output at the same voltage supplied, but the voltage variation factor is low, as shown in Fig

(2) Choke Input Type

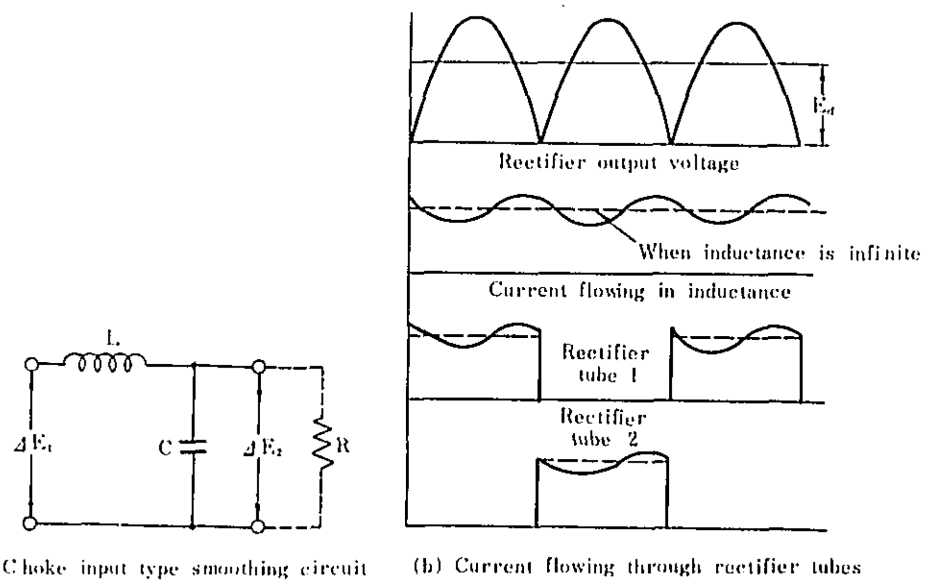


Fig. 75 Choke input smoothing circuit

In this circuit, the rectified current is averaged due to the choke coil, but as the inductance is definite, the current flow of the choke coil is delaying and pulsating than the output voltage of the rectifier as shown in Fig. 75(b).

If the pulsation voltage at input and out ut of the smoothing circuit is assumed as E_1 and E_2, the attenuation factor per stage will be,

$$\delta_2 = \frac{\Delta E_2}{\Delta E_1}$$

The value can be calculated by the following equation:

$$\delta_2 = \frac{1}{\omega^2 LC - 1}$$

Where, L is inductance (H), C is capacitance (F) and is the angular velocity of pulsation voltage fundamental frequency.

Next, let us proceed to the voltage variation factor of a rectifier of an input-choke smoothing circuit. As shown in Fig. 76, while the load current is small the output voltage will decreases rapidly, according to the increase of load current, but to a certain extent, the voltage variation factor is superior to the condenser input type. This means that when the load current is small or when the inductance of the choke is low, the smoothing circuit will operate like a condenser input-type. Therefore, the input inductance is necessary to be larger than a certain value. This is called the "critical inductance" and is given by the following formula.

$$\omega L \geqq R_{eff} \frac{E_1}{E_d}$$

Herein,

Reff = Value of the DC resistance of choke coil, rectifier and power supply transformer added to load resistance

$\frac{E_1}{E_d}$ = ratio between pulsating fundamental wave component and DC voltage contained in rectifier output (refer, Table 5)

The inductance of single-phase full-wave rectifier can be calculated as follows:

At 50 Hz, $$L \geqq \frac{R_{eff}}{942}$$

At 60 Hz, $$L \geqq \frac{R_{eff}}{1130}$$

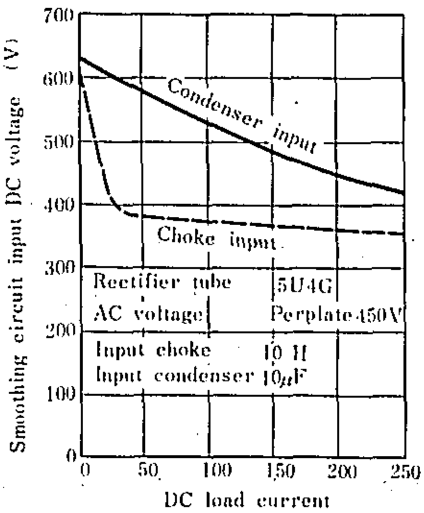


Fig. 76 Types of smoothing circuit and voltage variation factor

In general, the pulsation factor required for power source supply of vacuum tube is less than 0.1 %, and the product of L X C will be determine, according to the circuit.

In broadcasting transmitters, the inherent frequency of smoothing circuit should be taken into consideration, so that it will not resonate with the lower modulation frequencies. The inherent frequency is usually in the range of 10 - 15 Hz, thus L X C should be over 70 (H.μF). The combination of LZ and C may be considered to be freely selected, as far as the product meets the required value, but, in view of lowering the impedance viewed from the load side, it is desirable to use a large C. Although, the value of inductance should exceed the critical value.

The voltage variation factor of the choke input type smoothing circuit is low and the peak current is also small.

Table 5 Comparison of Various Types of Rectifier Circuits of Choke Input Type Smoothing Circuit

		Rectifier circuits			
		Single-phase full-wave with central tap	Single-phase full-wave bridge-type	3-phase half-wave	3-phase full-wave
Relation of voltage (relative value of DC output voltage is 1.0)					
a Effective value of transformer secondary voltage (per phase)		(one side of central tap)	1.11	0.855	0.428
b Max. inverse voltage		3.14	1.57	2.09	1.05
c Min. ripple freq. (f: power supply freq.)		2f	2f	3f	6f
d Ripple voltage	Fundamental wave component	0.667	0.677	0.250	0.057
	Secondary harmonic component	0.133	0.133	0.057	0.014
	Third harmonic component	0.057	0.057	0.025	0.006
e Smoothing device for reducing fundamental wave pulsation factor to 0.1 % L(H)× C(μF)		1,600	1,600	280	16
Values relating to current					
f $\frac{\text{Average current per plate}}{\text{Plate peak current}}$		0.500	0.500	0.333	0.333
g $\frac{\text{Average current per plate}}{\text{Load DC current}}$		0.500	0.500	0.333	0.333
h $\frac{\text{Peak current per plate}}{\text{Load DC current}}$		1.000	1.000	1.000	1.000
Relation of transformer kVA (relative value of DC output is 1.0)					
i Primary kVA		1.11	1.11	1.21	1.05
j Secondary kVA		1.57	1.11	1.48	1.05
k Average of primary and secondary kVA values		1.34	1.11	1.35	1.05

11. Performances Required of Radio Broadcasting Transmitters

To guarantee high-quality broadcast, radio transmitters are required to have the following performances according to the provisions of BTS and other regulations.

(1) To be able to radiate a designated electric wave within the frequency range of 535 - 1605 kHz, allocated to the mediumwave broadcast service.

(2) Frequency deviation

The tolerance of the radiated frequency deviation under the regulations is (±) 20 Hz, but NHK has laid down more stringent value of (±) 10 Hz, for its self-imposed control. The Japanese Government is to enforce its regulation on this matter to the same, effective on January 1, 1964.

(3) Frequency characteristics

The overall frequency characteristics of NHK's broadcasting facilities are controlled under the regulations, not to exceed (±) 2 dB between 100 - 7,500 kHz, at 1,000 Hz 50 % modulation. NHK has set a somewhat stringent value for self-impose control.

In our BTS standards, the frequency deviation is requested to be kept within (+) 1 dB from 50 - 7,500 Hz and (+) 1 dB and (-) 1.5 dB from 7,500 - 10,000 Hz, and the frequency characteristic should be declining for frequencies exceeding 10,000 Hz. The reference frequency is 1,000 Hz and degree of modulation is 50 %.

(4) Modulation degree linearity Government Regulations require linear modulation of transmitters up to 95 % at least. The transmitters of NHK are capable of 100 % modulation. In the BTS Standards, it is prescribed that 100 % linear modulation should be obtained between 50 - 5,000 Hz.

(5) Carrier variation factor It is provided by Government Regulations that at 1,000 Hz modulation, the variation factor of

carrier current amplitude shall be less than 5 %. In the BTS Standards, the variation factor is to be less than 4.5 %, for 0 - 100 % modulation at 1,000 Hz when the variation of the reception power voltage is maintained at 1.5 %.

(6) Non-linear distortion Overall distortion factor under Government Regulation is required to be less than 5 % for 80 % modulation, at frequencies of 200, 1,000 and 5,000 Hz. NHK 's maintenance standards are somewhat more strict.

BTS Standards are adopting the following distortion factors ;

Frequency range	Modulation degree	Distortion factor
100 - 5,000 Hz	20 - 80 %	2.5 % less than
	80 - 100 %	3 % less than
50 - 100 % and 5,000 - 10,000 Hz	20 - 80 %	5 % less than

(7) Noise The overall noise-ratio under Government Regulations is required to be less than (-) 50 dB for 80 % modulation at 1,000 Hz. NHK is adopting a little bit strict value for maintenance.

In the BTS Standards, the signal -to-noise-ratio is less than (-) 60 dB.

(8) Antenna power The tolerable deviation prescribed in Government Regulations is 5 % for the upper limit and 10 % for the lower limit. If these two limits are to be converted into the deviation of antenna current, the upper limit will be 2.5 % and the lower limit will be 5 %. It can be realized that a meter of high accuracy will be required to confirm the current deviation of the antenna current.

(9) Spurious radiation The strength of unnecessary radiation of electric waves, i.e. the harmonics or parasitic waves, besides the

required frequency -bands, should be less than 50 mW in terms of feeder power and, the level is limited to be 40 dB lower than that of the fundamental wave.

(10) Others In addition to the heretofore-described requirements, radio broadcasting transmitters are required to be easy to maintain and operate, with high stability, and the power consumption should be small.

Written in December, 1962 By Hiroshi Suzuki
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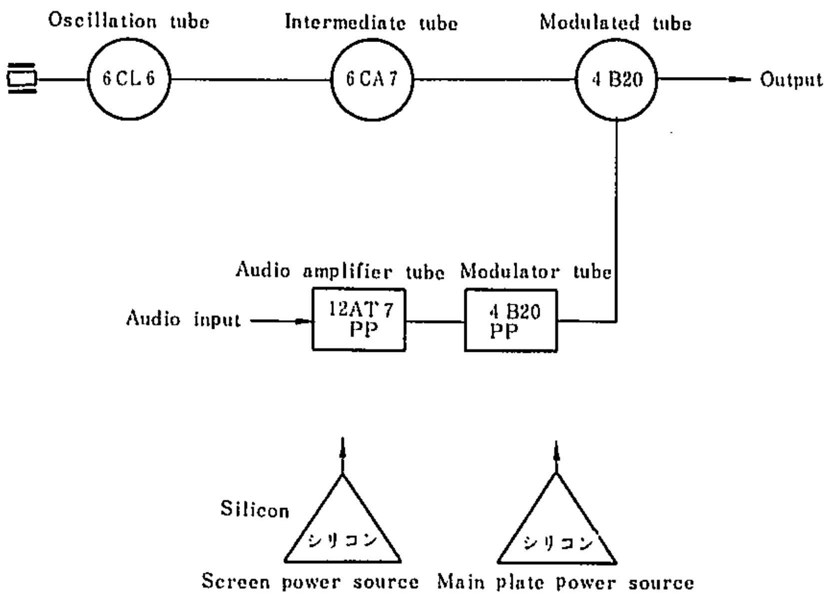
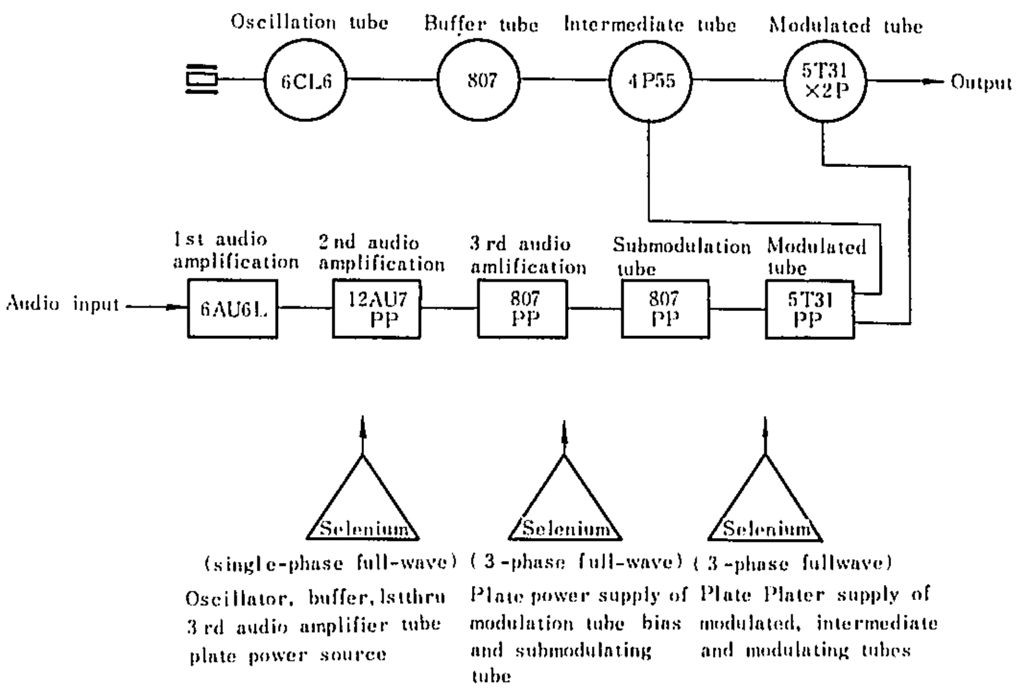
Revised in November, 1969 " P.C(250)

Annexed Table   Principal Transmitter Tubes for Radio Broadcasting Transmitters

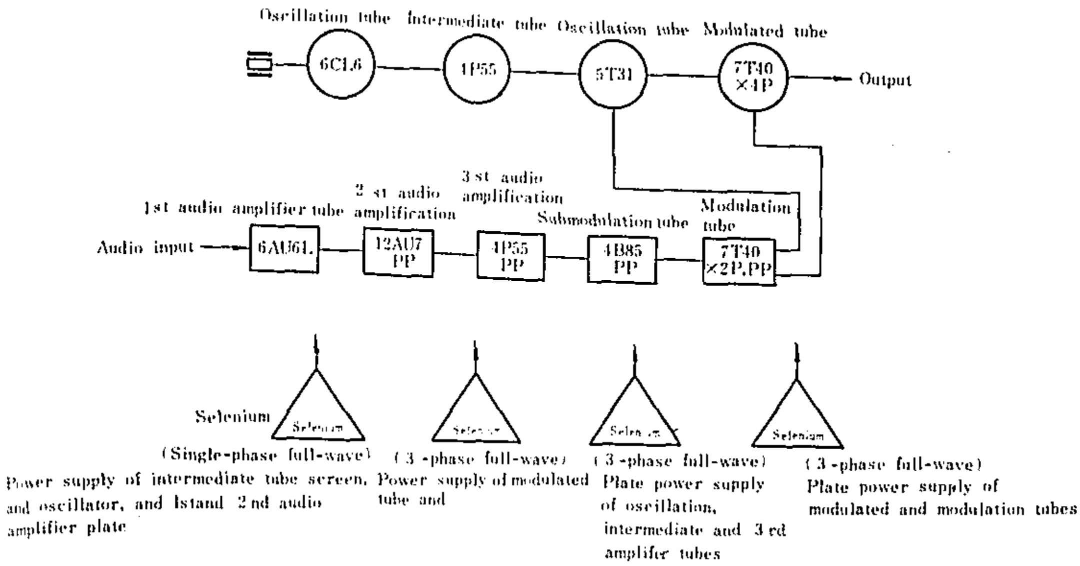
Type	UY807	4B13	4B20	4P55	4P60	4B85	SN205C	5T31	7T40	8T30	8T33	9T71
1. Kind	Beam	Beam	Beam	Pentode	Pentode	Beam	Triode	Triode	Triode	Triode	Triode	Triode
2. NHK's uses	Buffer and audio amp.	100W modulation, and modulated stage	100W modulation, and modulated stage	Intermediate and audio amp.	Intermediate	Submodulation	500W modulation, and modulated stage	1kW modulation, Modulated 5,10kW excitation	5kW modulation, modulated stage	10kW modulation, modulator 100kW excitation	10kW modulation, modulator 100kW excitation	100kW modulation, modulated stage
3. Type of cooling	Pentode	Pentode	Pentode	Pentode	Pentode	Pentode	Pentode	Pentode	Forced-air-cooled	Water-cooled	Water-cooled	Water-cooled
4. Cathode	Indirectly heated	Directly heated	Indirectly heated	Indirectly heated	Directly heated	Indirectly heated	Directly heated	Directly heated	Directly heated	Directly heated	Directly heated	Directly heated
Kind	Oxide	Tritane	Oxide	Oxide	Tritane	Oxide	Tritane	Tritane	Tritane	Tritane	Tritane	Tritane
Voltage (V)	6.3	10	6.3	6.3	10	6.3	11	7.5	7.5	12	7.5	11
Current (A)	0.9	5	3.9	3.2	3.25	4.8	12	12	16	40	60	285
5. Mutual conductance (mv) (I_b corresponding thereto)	6 (70mA)	3.75 (50mA)	(100mA)	6.5 (60mA)	2.6 (60mA)	20 (300mA)	7.5 (250mA)	4.4 (150mA)	7 (300mA)	9 (1A)	18 (1A)	30 (2A)
6. Amplification-factor	μ_2 7.5	μ_2 8.5	μ_2 5.7	μ_2 5.5	μ_2 6	μ_2 4.8	14	33	35	50	40	39
7. Internal electrostatic capacity (pF)												
Between G and P	0.2	0.16	0.9	0.4	0.07	1.5	3.4	5	5	18	26	53
Input	12	16	37	25	11	48	13	8	9	20	49	89
Output	7	13	17	21	13	20	3	0.5	0.5	0.2	0.7	1.2
8. Max. anode loss rating	25W	100W	100W	120W	125W	150W	500W	450W	1kW	10kW	10kW	50kW
9. Remarks		Old type			Old type		Old type			Old type Forced-air-cooled tube 8T30R is same in general ratings. Anote loss 6kW	Forced-air-cooled tube 8T33R is same in general retings. Anode loss 6kW	Forced-air-cooled tube 8T71R is same in general ratings. Anote loss 25kW



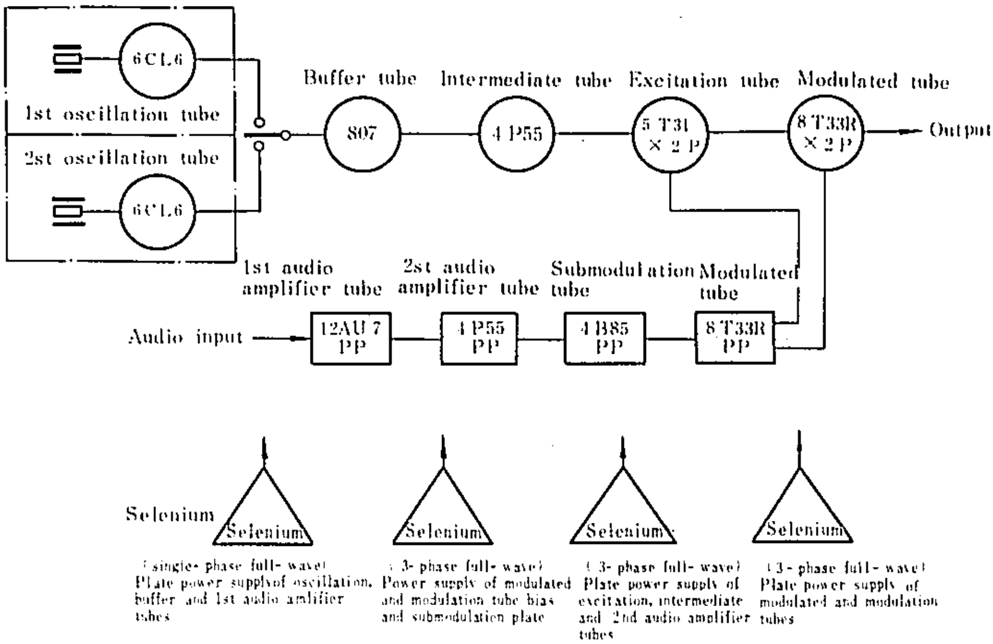
Annexed Chart 1 Block Diagram of 100W Broadcasting Transmitter (Model B100-8)



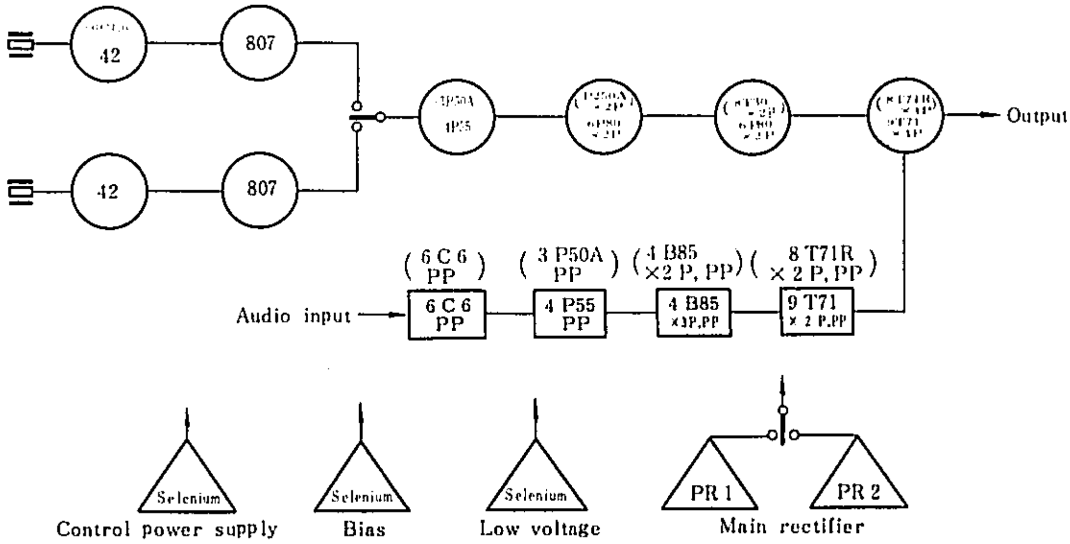
Annexed Chart 2 Block Diagram of 1kW Broadcasting Transmitter (Models B1K-5 and 6)



Annexed Chart 3 Block Diagram of 5kW Broadcasting Transmitter (Model B5K-1)



Annexed Chart 4 Block Diagram of 10kW Broadcasting Transmitter (Model B10K-10)



Annexed Chart 5 Block Diagram of 100kW Broadcasting Transmitter (Kasuga Broadcasting Station)