テレヴィジョン受像器の同調部の真空管と回路内の雑音について

Noise in Tubes and Circuits in a Tuner of a Television Receiver

By

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1. Radio-Frequency Input-Circuit Design Considerations

In general, the bandwidth of the radio-frequency sections of a television receiver must be broad enough to enclose the entire transmitter bandwidth curve if all available picture details is to be retained. This is readily accomplished by fixed tuned intermediate amplifiers following the frequency mixer.

In principle, a radio-frequency amplifier is not necessary, and the signal could be fed directly to the tuned input circuit of the mixer. The problems of image rejection, local oscillator rediation, excessive shot noise, and conversion loss of conventional mixers, however, are such that a stage of amplification ahead of the mixer is desirable. It becomes exceedingly difficult to achieve appreciable gain ahead of the mixer for the higher ultra-frequency channels, and a radio-frequency stage is not used in most cases. Rather, a simple tuned preselecter is used ahead of a crystal diode mixer, since triode mixer noise becomes excessive owing to transit-time effects. Then it becomes very important that the first intermediate-frequency amplifier be of an extremely low noise type, since the diode mixer results in a conversion loss. Disk-seal tubes of the planer or cylindrical type such as the pencil tube triodes[19] and the more recent development of conventional-type triodes suitable for operation up to 100 mc[20,21] however, make the use of radio-frequency amplifiers at ultra-frequency practical, although the disk-seal types pose problems in economy for multi-channel coverage and the performance of radio-frequency amplifiers at ultra-high frequency (UHF) is well below that at very-high-frequency (VHF).[22]

On low-band UHF channels, conventional miniature radio-frequency amplifier pentodes such as 6AK5 and 6AG5 offer no particular problems, although the input-conductance component resulting from transit-time effects begins to be important. As the higher channels are reached, and the input conductance becomes greater, the noise performance becomes such that triode radio-frequency amplifiers with their lower noise figure are desirable.

Triodes in conventional tuned circuits introduce problems resulting from grid-plate capacitance. These problems have been attacked in the past in various ways as indicated below:

1. Push-pull twin-triode radio-frequency amplifier followed by twin-triode mixer with push pull input and single-ended output in combination with push-pull local oscillator. Such arrangements permit balancing out of grid-plate capacitance in radio-frequency and mixer stages by cross neutralization.

2. Grounded-grid radio-frequency amplifier followed by cathode-coupled twin-triode mixer with conventional triode oscillator.

Push-pull circuits as radio-frequency amplifiers are relatively cumbersome where multi-channel operation is required, and improved tubes which are more stable make possible the use of single tubes in self-neutralized circuits. The neutralizing adjustment, in this case, however, tends to be critical and unstable. The grounded-grid circuit, while not requiring neutralization, has several important disadvantages, among which are the variation of input impedance with g_m and, hence, with bias, causing a serious mismatch, particularly where agc is applied to the radio-frequency stage, and the extremely low imput impedance, which

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makes the design of a circuit with the required selectivity exceedingly difficult.

Some of these disadvantages may be overcome with careful design. In particular, if a grounded-grid stage is followed by an additional radio-frequency amplifier ahead of the mixer, agc may be applied to the added stage, which will prevent subsequent stages from overloading while eliminating the requirement that agc be applied to the grounded-grid stage. Because of the low input impedance, the tuned circuit has a low Q, and selectivity is poor. This, however, can be made up in the following stage. Furthermore, an impedance step-down ratio is normally required. Such a requirement would result in a poorer noise figure than would be obtained from a neutralized grounded-cathode stage at frequencies where transit time is not important. This, however, is not the case where transit-time noise predominates. The input admittance is a function of bias both for grounded-grid and grounded-cathode operation where transit-time is appreciable. Hence there are difficulties in applying agc in either case. Most of the forms of input circuits used in the past have largely given way to some form of the cathode circuit with attendant signal-to-noise ratio advantages. This, circuit, as well as others, will be discussed.

2. Tube Consideration

If a television receiver is to work under a large variety of conditions, each of the following factors of tuner performance is highly significant; signal-to-noise ratio, selectivity and band-pass characteristics, voltage gain, amount of oscillator radiation, amount of antenna mismatch, and degree of cross-modulation in the radio frequency tube, and mixer tube.

It has been found that triodes, such as 6J6 and 6J4, generate little noise but are unstable in neutralized circuits, or have objectionable antenna termination characteristics in grounded grid operation. The pentodes, such as 6AG5 and 6AU6, generate considerably more noise, but do not require neutralization and are more stable in tuned input circuits. Thus improvement may be sought either with triode or pentode operation.

In considering the relationship of television tuner performance to tube design, the need for high sensitivity, i.e., low noise figure, dictates the use of a triode design having high transconductance, low input loading, low input and output capacitances, and low values of lead inductance.[3,4] Furthermore, for proper antenna termination, the tube should have an input impedance that does not change with variation of the gain-control bias voltage which must be applied to the radio-frequency amplifier stage to avoid overloading with strong signals. To reduce cross-modulation in the radio-frequency amplifier tube, an extended cut off characteristic is desirable. Unfortunately, this characteristic conflicts with the sharp-cutoff grid design desired for low imput loading. The oscillator radiation attriburable to the radio-frequency amplifier tube is a function of the capacitance from the radio-frequency amplifier output terminals to the antenna terminals, and of the circuit impedance at these terminals. The low-noise features of triodes have been recognized generally, but stability difficulties and other problems associated with the use of triode tubes in the conventional circuits have limited their extensive application in television tuners. Consequently, pentodes have been used in the radio-frequency stages in most receivers despite their higher noise.

3. Grounded-Cathode and Grounded-Grid Circuits

The grounded-cathode circuits have the serious disadvantage of requiring a neutralization adjustment which is rather critical and unstable when a tuned input circuit is used. The grounded-grid circuit, while it does not require neutralization, has a very low input impedance which varies inversely with transconductance. This variation makes it impossible to maintain correct antenna termination when gain-control voltage is applied to the radio-frequency stage. Also it is difficult to produce the low-inductance input circuit required for good selectivity. It is necessary to provide gain control in the radio-frequency amplifier when strong signals are present. Variation in receiver input impedance with bias, experienced with grounded-grid operation, causes improper antenna termination and resultant reflections which impair definition and may cause ghosts. These facts, plus the lack of a moderate-cost tube suitable for grounded-grid operation, may account for the rather limited use of the grounded-grid stage in the past. Television boosters, however, do not require gain-control voltage since they are not used generally with strong input signals, and may, therefore, employ push-pull grounded-grid operation.

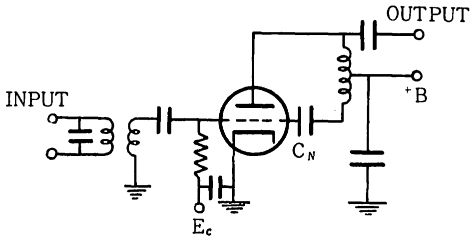


Fig. 1-a. Grounded-Cathode Circuit

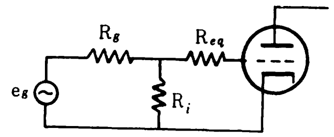


Fig. 1-b. Equivalent Grounded-Cathode Circuit

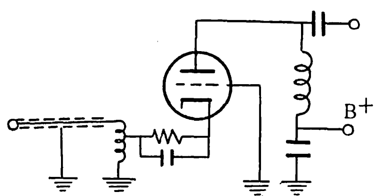


Fig. 2-a. Grounded-Grid Circuit.

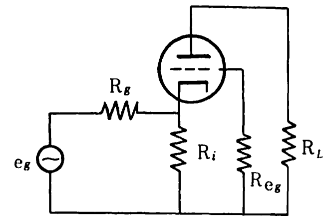


Fig. 2-b. Equivalent Grounded-Cathode Circuit.

Noise Figure of the grounded-cathode circuit is given[9] by

$$F=1+\frac{R_g}{R_i}+\frac{R_{eq}}{R_g}\left(1+\frac{R_g}{R_i}\right)^2$$

The equivalent circuit including the equivalent noise resistance R_{eq} is shown in Fig. 1-b. The noise figure is seen to depend on the input resistance R is shown in Fig. 1-b. The noise figure is seen to depend on the input resistance R_i whose value is set either by bandwith or input conductance considerations, and the equivalent noise resistance of the tube. When impedance matching is attained

$$F_0=2+4\frac{R_{eq}}{R_i}$$

Theoretically, these formulas apply for both triodes and pentodes. With triodes (which are more desirable by virtue of their low equivalent noise resistance), however, the qualifying condition is stability. Naturalized single-ended triodes are critical in adjustment, especially over a wide band. Push-Pull triodes are neutralized more easily but contribute twice the noise of a single tube.

The grounded-grid amplifier, shown in Fig. 2, is degenerative with the feedback voltage developed across the generator impedance due to the flow of plate current. The noise figure is given[9] by

$$F=1+\frac{R_{eq}}{R_i}+\left(\frac{\mu}{\mu+1}\right)^2\frac{R_{eq}}{R_g}\left(1+\frac{R_g}{R_i}\right)^2$$

When μ is much greater than unity, as it is in practice, the expression becomes identical to the expression for noise figure of the grounded-cathode circuit.

Since the plate current flows through the generator, the tube presents a cathode impedance

$$R_k = \frac{e_{kg}}{i_p} = \frac{R_p + R_L}{\mu + 1}$$

Assuming that $R_i \gg R_g$, $R_p \gg R_L$, and $\mu \gg 1$, $R_g = \frac{1}{g_m}$ for matching condition. For a triods

$$R_{cq} = \frac{2.5}{g_m}$$

and

$$F = 1 + \frac{R_{cq}}{R_g} = 3.5 = 5.5^{db}$$

voltage gain is, under the above assumptions,

$$A = \frac{1}{2} g_m R_L.$$

4. Cathode Degeneration in Amplifiers

The amplifier circuit with cathode impedance is shown in Fig. 3-a below. From the basic equivalent circuit of Fig. 3-b,

$$E_L = -I_P Z_L$$

$$I_P = \frac{\mu(E_g - I_P Z_K)}{Z_L + r_P + Z_K}$$

$$\therefore \quad I_P\left(1 + \frac{\mu Z_K}{r_P + Z_L + Z_K}\right) = \frac{\mu E_g}{r_P + Z_L + Z_K}$$

$$\therefore \quad I_P = \frac{\mu E_g}{r_P + Z_L + (1+\mu) Z_K}$$

$$\therefore \quad E_L = \frac{-\mu Z_L}{r_P + Z_L + (1+\mu) Z_K} E_g$$

$$\therefore \quad A = \frac{E_L}{E_g} = \frac{-\mu Z_L}{r_P + Z_L + (1+\mu) Z_K}$$

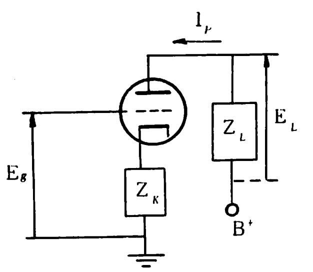


Fig. 3-a. Circuit with cathode impedance

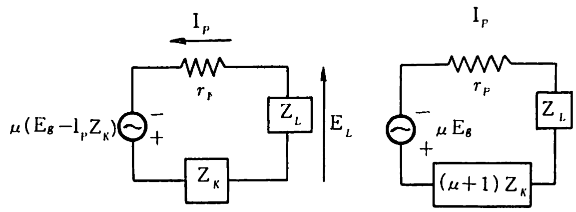


Fig. 3-b. Basic Equivalent Circuit

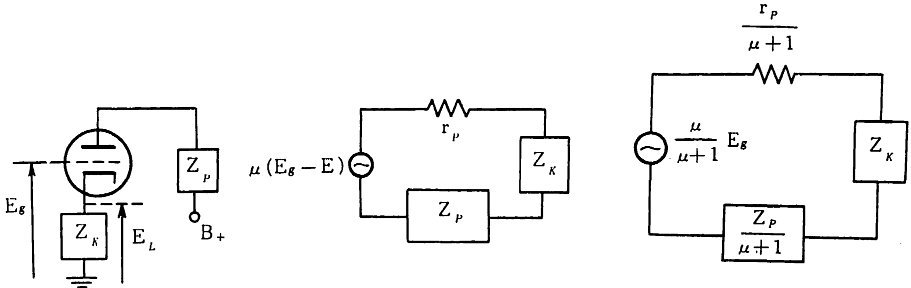
5. Cathode Follower

(Cathode-coupled amplifier, grounded-plate amplifier)

The circuit configuration of a cathode follower is shown below. As seen from the figure, the output impedance is placed between cathode and ground, instead of between the plate electrode and ground. From the basic equivalent circuit shown, voltage gain of the amplifier is

$$A=\frac{E_L}{E_g}=\frac{\mu Z_K}{r_P+Z_P+(1+\mu)Z_K}=\frac{\mu}{\mu+1}\frac{Z_K}{\frac{r_P+Z_P}{1+\mu}+Z_K}$$

When the amplification factor of tube μ is considerably greater than unity, as is normally the case, the cathod-follower system acts, as far as the load is concerned, as though the amplification of the tube was slightly less than unity and as though the effective plate resistance of the tube was approximately $r_P/\mu=1/g_m$, a relatively low resistance, when $Z=0$, or when plate is grounded.



Cathode follower Equivalent Circuit Alternative Equivalent Circuit

Fig. 4

The cathode follower circuit places one terminal of the load impedance at ground potential without the use of a blocking condenser, has an especially good response characteristics at high frequencies, which results from the capacitance shunting the load as the result of load or tube capacitance between cathode and ground, occurs at the frequency for which the reactance of this capacitance is equal to the equivalent resistance formed by equivalent plate resistance $\frac{r_p}{\mu+1}$ in parallel with the load impedance Z_k, when $Z_p=0$, and is large because of the small value of the equivalent plate resistance $\frac{r_p}{\mu+1}$. The characteristics at low frequencies can also be made very good by having the *d-c* cathode current flow through the load impedance, so that blocking condensers with resulting phase shifts are avoided.

The input admittance of a tube used as a cathode follower is somewhat less than when the cathode is grounded. In particular, the effective grid-cathode capacitance is less than the actual capacitance because the voltage between grid and cathode is less than the applied signal voltage. This causes any impedance Z_{gk} existing between cathode and control grid to appear to an applied voltage E_g to have a value greater than the actual impedance between these electrodes in accordance with the relationship

$$\text{Equivalent grid-cathode impedance}=\frac{Z_{gk}}{1-\frac{E_L}{E_g}}$$

The resulting reduction in input capacitance improves the frequency response characteristic of the amplifier driving the cathode follower stage, as compared with driving a conventional grounded-cathode system where the load impedance is in the plate circuit.

The equivalent generator impedance that the load impedance Z_L sees when looking back toward the tube in a cathode-follower stage is so low as to make it possible to obtain an impedance match between the tube and a low impedance load such as a cable or other transmission line. Such matching between gecerator impedance and load impedance is often desired in television systems to eliminate reflections at the sending end of the cable, and thereby to reduce echoes. When an impedance match is desired between a tube and a transmission line, a triode tube is selected such that with normal operation the equivalent tube impedance will approximate the characteristic impedance of the transmission line. A final adjustment can then be made either by varying grid bias of the cathode follower stage, or by adding trimming resistances in shunt with the input of the line if the tube resistance is too high, or in series with the load if the tube resistance is too low.

The chief use of the cathode-follower amplifier is as a wide-band power amplifier to deliver power to loads of relatively low impedance. In carrying out this function it will be noted that the cathode follower does not amplify voltage, since the voltage developed acrsss the load will always be less, though usually only slightly less, than the signal voltage applied to the amplifier input. Rather, the cathode-follower acts as an impedance transformer, which takes a voltage developed across a relatively high impedance, and by applying it to the grid of the cathode-follower stage transfers this voltage only slightly reduced in amplitude, to a valtage across a relatively low resistance load. Thus amplification of power is obtained, with added advantages of excellent frequency response, high input impedance, and low distortion.

The maximum power that a cathode-follower stage is capable of developing in the load impedance corresponds to the power delived to the load resistance when the plate current swings from the operating value down to virtually zero. Thus, to a first approximation,

$$\text{Maximum power output without excessive distortion} = \frac{I_b^2 R_L}{2}$$

where I_b is the *d-c* plate current at the operating point, and R is the resistance component of the load impedance. It will be noted that adjusting the operating conditions so that the output impedance of the tube will match the impedance of the load does not affect the power that can be delivered to the load, except indirectly as such an adjustment affects the *d-c* plate current.

The amplitude distortion of a cathode-follower stage tends to be small because negative feedback action is present.

The tubes used in cathode follower systems are normally triodes having high transconductance, or pentodes with high transconductance connected to operate as triodes. A high transconductance is important in order to make the equivalent output impedance of the tube small. If the rated plate current of a single tube is not sufficient to give the required power output, additional tubes can be placed in parallel. Doing so increases the input capacitance of the system somewhat, an effect that must be taken into account in designing the exciting amplifier stage, but does not cause the high-frequency response of the cathode-follower system to deteriorate. This is because the equivalent output resistance goes down in direct proportion to the increase in shunting capacitance caused by the increased number of tubes.

The noise figure of a cathode follower is the same as that of a grounded cathode amplifier.

Because of its lower gain and regenerative tendencies, depending on the nature of the cathode impedance, this circuit has found limited application in the television tunner.

6. Double-plate Grounded-Grid Circuit

A double-plate grounded circuit offers an easy way to separate two television bands. The common cathode is the input terminal for all the channels and the plates pass the two bands respectively. In the figure given, the two triode sections are assumed identical and the alternate load impedances are assumed small in comparison with the plate resistance at the respective operating frequencies.

Assuming $\mu \ll 1$, and matching the generator by the dynamic input conductance of the tube noise noise figure is 11.3 db.[9] This circuit is inferior, from a noise figure standpoint, to other triode configurations.

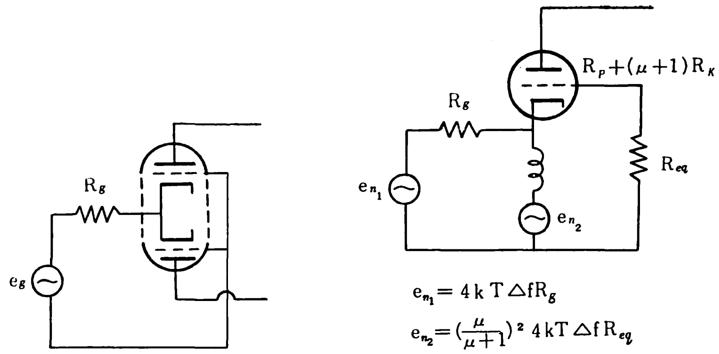


Fig. 5. Double-plate Grounded-grid Circuit

7. Cathode-Coupled Triode

The cathode coupled amplifier combines the qualities of high input impedance and high gain and essentially the favorable noise-figure of a single grounded-grid triode. In Fig. 6, it is assumed that the two triodes are identical and the input resistance of the second tube is high compared with equivalent cathode impedance of the preceding stage. The noise figure is given[9] by

$$F=1+\frac{R_g}{R_i}+2\frac{R_{eq}}{R_g}\left(1+\frac{R_g}{R_i}\right)^2$$

under conditions of impedance match

$$F_0=2+8\frac{R_{eq}}{R_g}$$

There is a decided advantage in using this circuit if R_g can be made high.

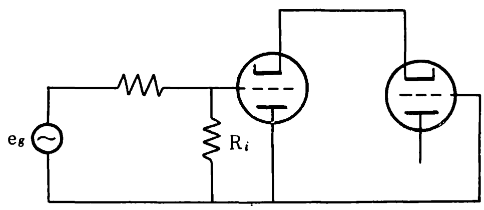


Fig. 6. Cathode-Coupled Triod.

8. Inverted Amplifier Circuit

Fig. 7 shows an inverted amplifier of C. E. Strong, which is an improvement of inverted ultra-audion amplifier of H. Romander, which in turn is a modification of Alexanderson's grounded-grid-amplifier circuit.

9. Cascade Circuit

The cascade circuit, shown in Fig. 8, consists of a normal grid-driven amplifier with its plate load consisting of the input impedance of a grounded-grid stage. This configuration is particularly useful in low-level input stages where tube noise is the limiting factor. It combines the desirable qualities of a pentode, namely, the high gain, stability, low out-put-to-input admittance and high input impedance, and the low noise quality of a triode.[7,8]

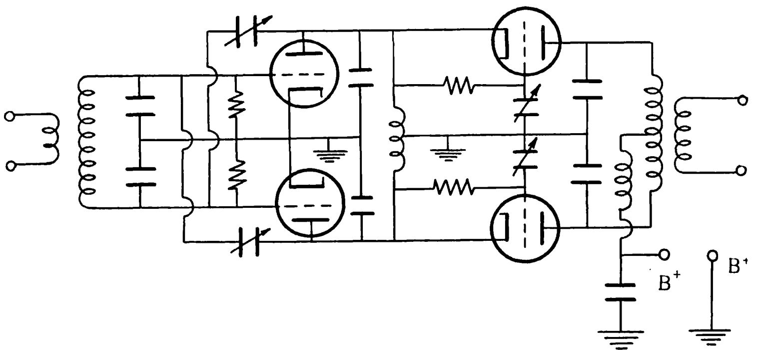


Fig. 7. Inverted Amplifier, push-pull grounded-cathode stages driving push-pull grounded grid stages.

This circuit, while well suited for intermediate-frequency amplifier use, is extremely difficult to apply to a multi-channel tuner because the neutralization, which is frequency-sensitive, is required for the input circuit, and it requirs, therefore, individual coil switching for each channel. This neutralization is not extremely critical at any one frequency and can be accomplished with a tuning coil which is effectively in parallel with the grid-plate capacitance of the first unit. The neutralization coil also serves as a radio-frequency choke returning the cathode of the second unit to ground, thus eliminating the cathode choke otherwise required. Attempts to use this circuit without neutralization have been unsuccessful, except at the lower-frequency channels, because the degenerative feedback increases with frequency. The capacitance to ground from the plate of the input triode and from the cathode of the output triode, plus the distributed capacitance to ground of their connecting leads, also causes degeneration in the higher frequency channels where the value of this capacitive reactance approaches the input impedance of the grounded-grid-section. This input impedance is approximately the reciprocal of the transconductance and is in the order of 200 ohms in a tube having a transconductance of 500 microohms. A distributed capacitance of only 2 $\mu\mu f$, because it has a reactance of only 400 ohms at 200 MC, appreciably reduces the input impedance of the grounded-grid unit. This effect reduces the gain, causes degeneration due to the capacitive phase angle, and allows the noise of the output unit to contribute to that produced by the input unit, impairing the noise figure

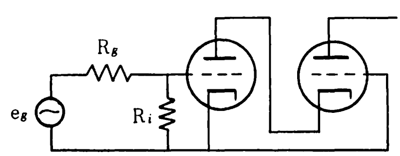


Fig. 8. Cascade circuit.

Assuming identical triodes, no cathode resistance, and $\mu \gg 1$, noise figure is[9]

$$F=1+\frac{R_g}{R_i}+\frac{R_{eq}}{R_g}\left(1+\frac{R_g}{R_i}\right)^2\left(1+\frac{1}{\mu^2}\right)$$

Under matching condition and $\mu^2 \gg 1$

$$F_0=2+4\frac{R_{eq}}{R_g}$$

The noise figure is better than that of the cathode couple arrangement, and the gain is higher by 6 db since the voltage gain of the first stage is unity and not one half as in the cathode-coupled circuit.

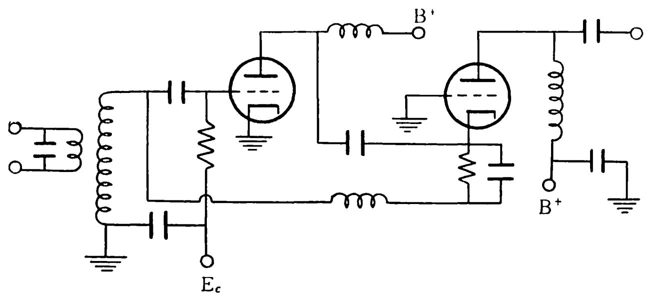


Fig. 9. Cascade Circuit

10. Driven-Grounded Grid Circuit

Fig. 10 shows a driven-grounded-grid circuit. This term is a misnomer in that the inverted amplifier and cascade circuits are driven-grounded-grid circuits. Neutralization is accomplished by means of a bridge circuit commonly employed with single triode amplifiers. This method of neutralization has the distinct advantage of being relatively independent of frequency, provided the connecting leads in series with the neutralization capacitor are short. This circuit requires less involved switching than in the cascade circuit, but requires one more switch contact than a pentode circuit.

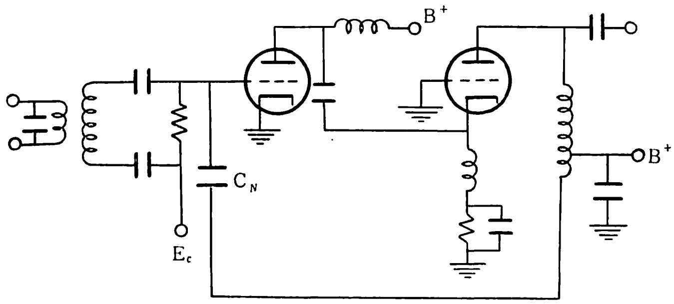


Fig. 10. Driven-grounded-grid Circuit

11. Direct-Coupled Driven-Grounded Grid Circuit

Fig. 11 shows another version of the driven-grounded-grid circuit in which the plate of the input triode is directly coupled to the cathode of the output triode. Neutralization is accomplished in the same manner as previously described. This circuit has the advantage

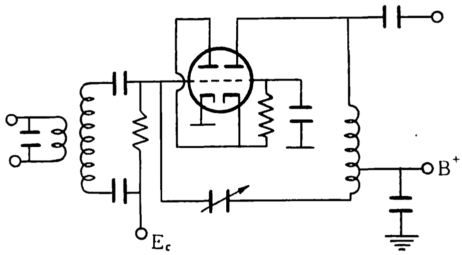


Fig. 11. Direct coupled driven grounded grid circuit.

that several components are eliminated from the coupling network between the two units, consequently, the distributed capacitance is reduced and the gain at the higher channels is increased. Another important advantage is that application of bias to the input triode causes the voltage between plate and cathode to increase, extending the cut-off of the tube. This extension reduces cross-modulation, without the use of a remote-cutoff tube. Such a tube would adversely affect the signal-to-noise ratio, either by increasing input loading or by reducing transconductance. Because of lower distributed capacitance between the plate circuit of the input triode and ground, this amplifier can give fairly satisfactory results on the low channels without neutralization. It is interesting to note that the number of components in this circuit equals the number required for a conventional pentode amplifier, the grid resistor and by-pass capacitor in the grounded-grid triode circuit being equivalent to the screen resistor and capacitor of the pentode circuit.

The foregoing driven-grounded-grid circuits have an input impedance and an admittance from output to input terminals which are dependent to a large extent on certain characteristics of the tubes employed.

12. Modified Cascade Circuit

A method which eliminates the need for actual neutralization by lowering the impedance loading the plate circuit of the input stage is shown below.

At the lower frequencies, this entire circuit is ineffective where ordinary neutralization is not needed anyway. Use of a cascade circuit does not solve the problem of variable input loading due to variable bias, since transit-time loading is a major portion of the input conductance and varies directly with application of agc thus poses a problem, but one not so difficult as for the grounded-grid amplifier, and some form of delayed agc is practical.

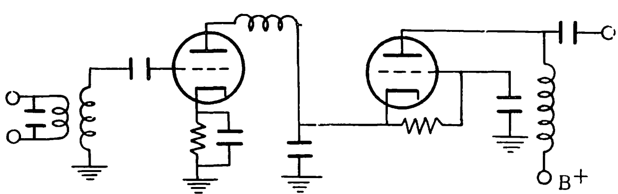


Fig. 12. Modified cascade circuit.

13. Considerations Above 100 MC and U. H. F.

Above 100 MC the cathode inductance becomes an important factor in determining the input loading. The input conductance varies directly as square of frequency. Within the range of frequencies where $(g_m \omega L_k)^2 \ll 1$, the capacitance $C+C_{gk}$ shunted by a resistance $\frac{1}{g_m}\left(\frac{f_0}{f}\right)^2$, where f is the operating frequency and f_0 the resonance frevuency of the cathode inductance and the grid-to-cathode capacitance.

Another source of loading is the electronic input conductance due to the finite transit time in the cathode to grid space, the input resistance due to transit time is

$$R_\theta = \frac{20}{g_m(\omega\tau)^2}$$

where τ is the transit time, and R_θ is the damping effect on the circuit. The noise contribution of R_θ is greater than that of an equal external resistance by the ratio of the equivalent tube temperature to room temperature. It is necessary, therefore, to separate R_θ from other input resistors.

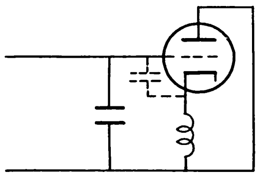


Fig. 13. The effect of Cathode inductance at high frequencies

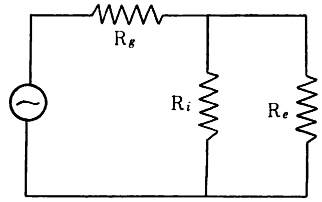


Fig. 14. Effect of transient time

The noise figure of the input circuit is[9]

$$F=1+\frac{R_g}{R_i}+\frac{T}{T_1}\frac{R_g}{R_\theta}$$

The noise figure of a tube and its input circuit is[9]

$$F=1+\frac{R_g}{R_i}+\frac{T_1}{T}\frac{R_g}{R_\theta}+\frac{R_{eq}}{R_g}\left(1+\frac{R_g}{R_i}+\frac{R_g}{R_\theta}\right)^2$$

When the ratio of $T_1/T=5$ is used, account is taken of the induced noise. This noise component being due to the induced noise voltage in the grid when the grid to the cathode transit angle is not negligible. It is directly related to R_{eg} and measurements have shown that the choice of $T_1/T=5$ gives reasonably accurate results. Tube design is steadily being improved and it is now possible to build radio-frequency amplifiers with such tubes and semi-lumped circuit techniques for frequencies up to 100 MC in grounded-grid circuits.[29] More complicated interstage tuned circuits are required in order to achieve the required selectivity and noise figure.[30] An alternative approach to the UHF tuner problem is to eliminate the radio-frequency amplifier and its problems entirely, replacing it by a tuned pre-selector in order to obtain the required selectivity, followed by a crystal mixer also fed from a conventional local oscillator. The noise figure of a crystal mixer is itself considerably lower than that of a triode mixer; however, in the absence of radio-frequency gain and because of the conversion loss of the mixer, the first intermediate-stage must be one with an extremely low noise figure.[31]

14. Broad-Band Tuning

The television front end selectivity requirements depend largely on the nature and intensity of the sources of interference in the image spectrums. If an intermediate frequency is chosen in the 41 to 47 region, the image spectrum falls well outside the television bands, as it should be possible to devise a selective radio frequency circuit broad enough to pass all the channels within one band, yet having sufficient selectivity to reject effectively all image signals. With proper design, the intermediate-frequency selectivity can be relied upon for adjacent channel rejection.

By eliminating the need for selective radio-frequency tuning, station selection can be accomplished by tuning the oscillator.

Since the bandwidth required of the radio-frequency stage in such a system is six or seven times that assigned to one channel, a proportional reduction in gain would be expected. Fortunately, however, the improvement in the figure of merit as a result of eliminating the switch and other incidental capacitances makes it feasible to obtain gains which compare favorably with those of switch-narrow-band circuits using similar tubes. Also, because only a fraction of the pass-band is being used at any one time, further increase in gain can be realized by reducing the damping from its critical value, without appreciably deteriorating the resolution.

The antenna and input radio-frequency stage set the ultimate limit of the useful receiver sensitivity. A yardstick of the quality of the radio-frequency stage is its noise figure which is defined as the ratio of the stage's actual noise power output to the noise power output due to antenna thermal agitation noise of an amplifier of identical bandwidth and gain, but introducing no noise of its own.

The over all noise figure of a multistage network can be determined by considering the noise contributed by individual stages and the gain of any preceding stages. It is apparent that, if the stages have even moderate gains, the noise figures beyond the first stage or the first two stages may be neglected with negligible error.

15. Antenna Transformers

The design of the imput transformer is predicated by the conditions of match and bandwidth. A single-tuned tranformer can be used by tapping on either the inductance or capacitance. The split-capacitance form is simpler physically and easier to adjust, but it must be remembered that by its very nature it is also a low-pass filter and sufficient low-frequency selectivity must be secured elsewhere.

The signal plate current is directly related to the available antenna power and inversely related to the product of bandwidth and input capacitance. If e and r are the antenna signal voltage and antenna radiation resistance respectively, then

$$i_p=\sqrt{\frac{e^2}{4r}}\,\frac{g_m}{\sqrt{\pi \Delta f c_i}}.$$

16. Interstage Coupling

Selectivity and figure of merit are the major considerations in the design of the coupling network from the radio-frequency amplifier to the mixer. The response to the image frequency devided by response to the signal frequency is given as approximately equal to $\Delta f/4f_i$, for a single-tuned circle and $2(\Delta f/4f_i)^2$ for a double-tuned circuit. In the high band where the sources of image interference are not particularly powerful the mean rejection figure of 5 obtainable with a double-tuned circuit, will be found sufficient, especially when an equal amount of selectivity is secured in the input transformer. A higher degree of selectivity is desirable in the lower frequency band. The lower-skirt selectivity should be high enough to reject the image signals. A simple network offering a high degree of selectivity and rejection of selected frequencies onside the pass-band is the bridged-T.

The need for selectivity not being as great at the higher frequency band, the use of simple circuits is desirable to assure uniform results with the small tuning elements employed. Stagger-tuning offers similar selectivity characteristics with greater simplicity of construction and tuning at some loss of gain.

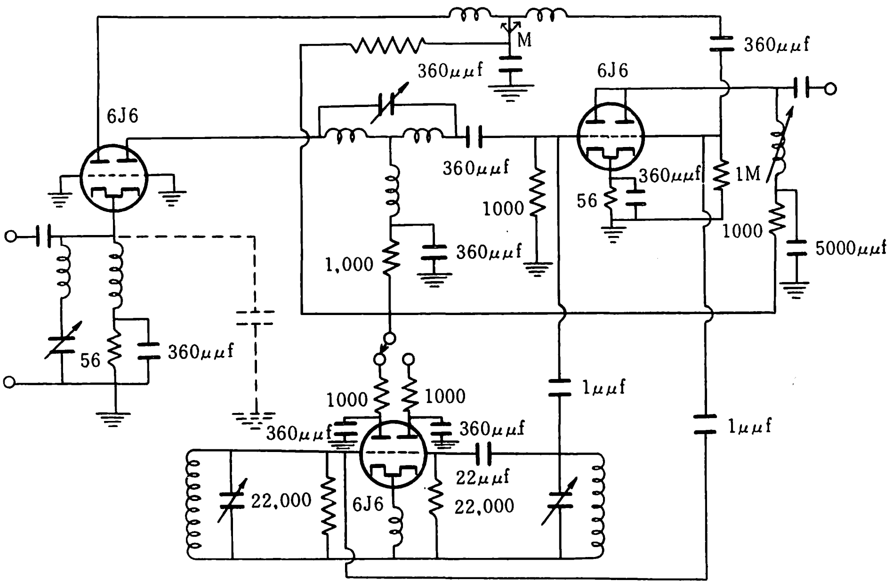
When the input and output capacitances are lumped, the tuning inductances at the upper frequency channels generally become so small as not to lend themselves to normal tuning methods. To increase the tuning inductance the series instead of parallel arrangement of the output and input capacitances may be used for tuning. The gain and band-width prcduct in this case remains the same.

Inasmuch as the radio-frequency amplifier in its wide-band application offers comparatively lower gain the noise properties of the mixer stage cannot be completely disregarded in evaluating the overall noise figure. By this token it is desirable to use a low noise-contributing tube such as triode. Due to its high grid-to-plate capacitance, however, care must be excercised in its use both at the low and high bands. At the low band the intermediate-frequency is close to the radio-frequency pass-band and it may, therefore, be necessary to include an intermediate-frequency series resonant trap in the grid of the mixer depending on the selectivity.

At the high band, resonance within the radio-frequency pass-band of the inductonces inherent in the coupling capacitor and elsewhere in the mixer plate circuit should be avoided.

17. Practical Circuits

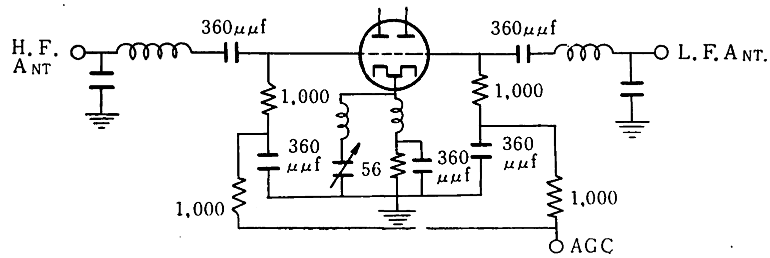
The simplest, although least satisfactory from a noise-figure standpoint, circuit is shown in the figure below. Two sections of a 6J6 are used as grounded-grid amplifiers, one for the low and one for the high band. The input matches either a 100-ohm unbalanced line or with the additions of a balun transformer can be made to match a 300-ohm balanced line. The network in the cathode has two frequencies of antiresonance, one in the midband of the low and one in the midband of the high channels and is series resonant in the f-m band.



This aids in matching the line and offers rejection f-m interference.

The plate circuit of the low-channel amplifier employs a bridged-T as the coupling network to the mixer tuned to reject the interfrequency. A double-tuned circuit is used to couple the high band amplifier. The two amplifiers couple into the respective grids of the 6J6 mixer, the plates of which are connected in parallel. By this device there is no need to switch any radio-frequency circuits. Two Colpitts osscillators for the respective bands are used, only one being operative at a time. The bands are switching the B^{+} voltage to the appropriate triode section of the 6J6 oscillator.

The circuit which is identical to the one just described except for the radio-frequency

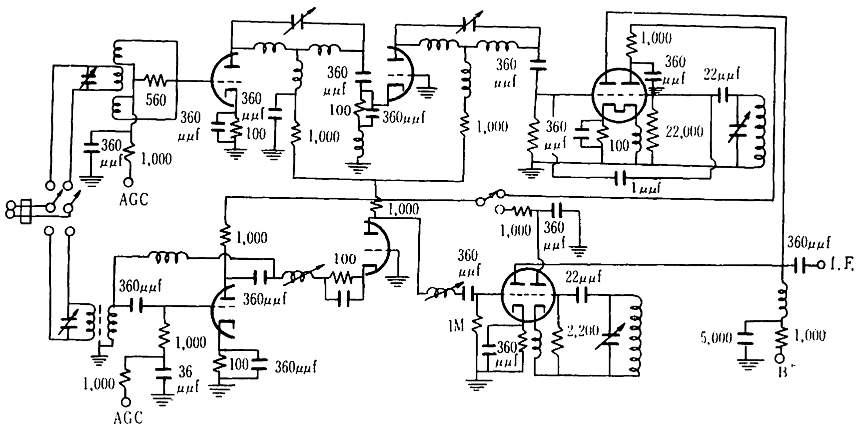


stage which uses a 6J6 in what is essentially a reflex cathode-coupled amplifier, as shown in the figure is examined.

It is assumed that in either band the tuning elements of the other band bypass the appropriate tube elements. This circuit still retains the basic simplicity of the first, but offers a better noise figure by virtue of the higher impedance at the grid input. Two inputs may be combined by using a proper dividing network.

Next, a circuit which uses four 12AT7 double triodes, one in a cascade circuit for each of the channels and two as oscillator-mixer combinations. Band switching is accomplished by switching the transmission line to the alternate inputs as well as switching the B^+ to the oscillators.

Double-tuned input tracsformers designed to match a 300-ohm line over a pass band of 40 mc are used. Two bridged-T networks are used in the low band amplifier, the one designed to reject the intermediate-frequency couples to the mixer and the other designed to reject the f-m couples the plate to the cothode of the cascade.

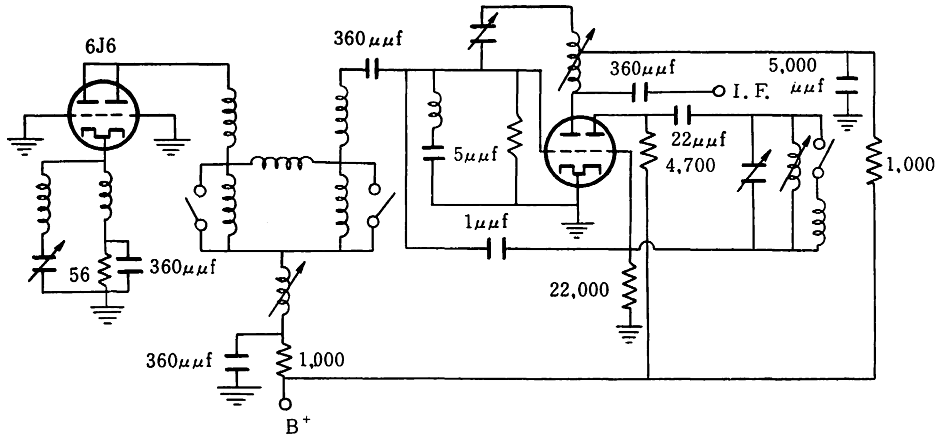


The high band employs a staggered pair in its coupling networks and an inductance is added to tune out the grid-to-plate capacitance. In the low band amplifier, the improvement resulting from tuning out the grid-to plate capacitance was not enough to warrant the addition of another tuning element. The inductances in the staggered pair, the primary antenna transformer capacitances and the bridged-T trap capacitors are made tunable; all other elements are fixed-tuned. This was found particularly justifiable in the low band amplifiers where Q is low (approximately 2).

By using one section of the tube as an oscillator and the other section in the same envelope as the asscciated mixer, more favorable and uniform injection of oscillator voltage was obtained. Little detuning is caused by the application of age.

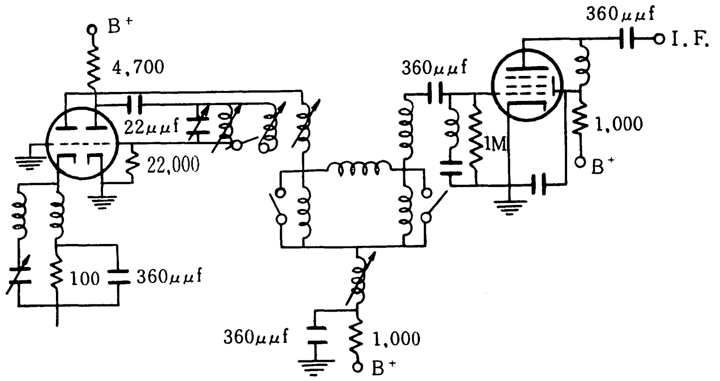
The common feature of all the circuits described thus far is that band switching is accomplished by applying B^+ to the tubes chosen for the selected band, and in one case by switching also the antenna cable at a low radio-frequency impedance level. By dispensing with switching of the radio-frequency coupling circuits, a higher figure of merit would be realized as a result of eliminating switching capacitances. Only half of the tubes, however, are used at any particular time. More efficient operation may be secured by switching the radio-frequency coupling circuit. With proper care the figure of merit at the high band need not be compromized and only a slight reduction would be suffered in the low band.

In the figure shown below, the first 6J6 functions as a grounded grid radio-frequency amplifier. The cathode network is the same as the second circuit discussed above, being antiresonant within the two television bands and resonant with the f-m band. The high-band radio-frequency coupling network is not switched and hence the switch capacitance does not affect it. The low-band circuit appears in series with that of the high band and is switched. The switch introduces an additional 1.5 to 2 $\mu\mu$f of capacitance into the low band-pass filter.

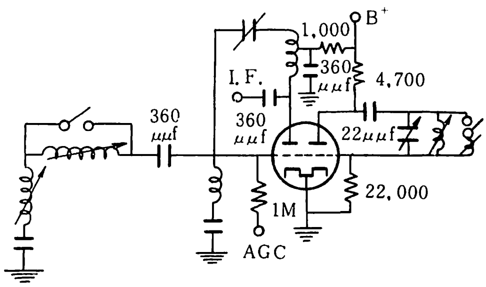


Double-tuned inductively-coupled circuits are used for interstage coupling. To suppress the response to image and intermediate-frequency signals, series resonat traps are included. The series resonant filter in the grid of the mixer is tuned to the intermediate frequency rejection, the mixer grid impedance is also reduced, making it less subject to regeneration. In spite of the low grid impedance to the intermediate-frequency, however, it is necessary to neutralize the mixer.

In order to avoid the need for neutralization which calls for an additional tuning adjustment, a pentode mixer may be employed as shown in the figure below.



Here the radio-frequency amplifier and oscillator are combined calling for a double triode with individual cathodes. Oscillator radiation under condition is higher and the gain and noise figure is less favorable. The figure below shows a tuner comprising a single dual triode, one section being used as an oscillator and the other as a mixer. In spite of the lack of radio-frequency gain preceding the mixer, a good noise figure can be obtained by virtue of the fact that the grid impedance of the triode mixer is high. The main weakness of this circuit is the prohibitive amount of oscillator radiation which is likely to result. This difficulty can be eliminated to some extent by the choice of an intermediate-frequency in the 40 to 50 region which will place the oscillator frequencies outside the television bands. The choice of higher



intermediate-frequency, however, will make neutralization more critical and lower the intermediate-frequency rejection ratio.

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テレヴィジョン受像器の同調部の真空管と回路内の雑音について (摘要)

伊 波 直 朗

本論文においては，テレヴィジョン受像器の同調部の真空管内および回路中におこる電気雑音の原因を解析し，種々の回路方式の得失を論じ，最後に低雑音高性能の回路の設計を提示した。無線周波入力回路の設計にあたって払わなければならない一般的考慮，種々の真空管の雑音特性の比較，陰極接地型回と格子接地型回路の得失，増幅器の陰極劣化，陽極接地型回路の特性，双陽極格子接地型回路の特性陰極結合三極管の特性，逆倒増幅器回路の設計，カスケイド回路の特性，被励振格子接地型回路の設計，直接結合被励振格子接地型回路の特性，100 メガサイクル以上の周波数における特性，広帯域同調，アンテナ変圧器，段間結合等も論議されている。最後に 6 種の実際的回路の設計を提起してある。

一般にテレヴィジョン受像器の無線周波部分の帯域は，実現可能なだけの絵素の鮮度を保つためには送像器の全帯域曲線を包含するものでなければならない。これは周波数混合管の次段に固定同調中間周波増幅器をおくことによって，かんたんに達成できる。

原則的には無線周波増幅器は必要不可欠のものではなく，信号を直接混合管の同調入力回路に印加してもよい。しかしながら，影響拒否局部発振器からの輻射,大きな覆射雑音,ふつうの混合管の変換損失などのもんだいがあるため，混合段以前に増幅段を入れることがのぞましい。UHF チャネルの高い方の周波数域に対しては混合管の前段に大きな利得をもたせることは非常に困難となるので，多くのばあいは無線周波段は使われていない。むしろ水晶二極混合管の前段に単純な同調プリセレクターを使うのがふつうである。これは走行時間のために三極混合管の雑音が大きいためである。このときには，最初の中間周波増幅器は非常に雑音の低い型のものでなければならない。これは二極混合管を使うと変換損失があるためである。しかし近年の特殊真空管の発達によって無線周波増幅器を 1000 メガサイクルに至る周波数帯域で使用することも可能になってきている。

VHF チャネルの低い帯域では，6AK5 や 6AG5 のようなふつうのミニアチュア無線周波増幅用五極管を使っても，走行時間効果による入力コンダクタンスが重要になりかかってくるとはいえ。まず，もんだいはない。チャネルの周波数が高くなるにしたがってこの入力コンダクタンスは大きくなり，雑音指数の小さい三極管無線周波増幅器の方が雑音特性の点からのぞましくなってくる。

ふつうの同調回路を使うときの三極管には格子―陽極間静電容量によっていろいろな問題が生じてくる。この問題を解決するには次に掲げるようにいくつかの方法がある。

1. プッシュプル双三極管無線周波増幅器の次段にプッシュプル局部発振器とプッシュプル入力単一終端出力の双三極混合管を使うこと。こうすれば，相互中和によって無線周波及び混合段において格子―陽極間静電容量を平衡させることができる。
2. 格子接地型無線周波増幅器の次段に陰極結合双三極混合回路とふつうの三極管発振器をおくこと。

プッシュプル回路を無線周波増幅器に使うことはたくさんのチャネルで動作させなければならないときにはどちらかといえば厄介である。安定度の高い改良された真空管ができたために，現在では自己中和回路で単一の管を使うことも可能である。しかしこの場合，中和調整は臨界的で不安定である。格子接地型回路は中和を必要としないが，重要な欠点をもっている。中でも重要なのは g_m によって，したがってまた偏倚電圧によって，入力インピーダンスが変化し，重大なインピーダンス不整合を（特に自動電圧制御が施されているときに）生ずるということと，入力インピーダンスが低いため必要な選択度をもった回路を設計することがいちじるしく困難であるということである。