CHAPTER 25

FREQUENCY CONVERSION AND TRACKING

SECTION 1 : THE OPERATION OF FREQUENCY CONVERTERS AND MIXERS

(i) Introduction (ii) General analysis of operation common to all types (iii) The oscillator section of converter tubes (iv) The detailed operation of the modulator or mixer section of the converter stage (v) Conclusion (vi) Appendix.

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(i) Introduction

The better modern radio receivers are almost universally designed to use the superheterodyne circuit. In such a circuit, the received signal frequency is heterodyned with the frequency of a local oscillator to produce a difference frequency known as the intermediate frequency. The resultant signal is amplified by a selective, fixed-tuned amplifier before detection. Since the heterodyne action is usually accomplished by means of a suitable vacuum tube, it is the purpose of this paper to discuss the chief similarities and differences among the tubes which might be used, as well as to explain their behaviour.

The combination of signal and local-oscillator frequencies to produce an intermediate frequency is a process of modulation in which one of the applied frequencies causes the amplitude of the other to vary. Although this process was originally called heterodyne detection and, later, first detection, it is now called frequency conversion. The portion of the radio receiver which produces conversion may, therefore, be identified as the converter. If conversion is accomplished in a single vacuum tube which combines the functions of oscillator and modulator, this tube may logically be termed a converter tube. When separate tubes are used for the oscillator and the modulator portions of the converter, respectively, the tube for the latter purpose is conveniently called a modulator or mixer tube. This terminology will be used in this paper.

Although in some of the earliest superheterodynes, frequency conversion was accomplished by a triode oscillator and triode modulator (Ref. 1) other circuits used a single triode which served as both modulator and oscillator (Ref. 2). A triode used in the latter way could, therefore, be called a converter tube. The introduction of two-grid tubes (i.e. tetrodes) permitted a wide variety of modulator and converter arrangements which frequently gave superior performance to that possible with triodes (Ref. 3-7).

When indirectly heated cathodes became more common, conversion circuits in which the oscillator voltage was injected in the cathode circuit were used. These circuits reduced considerably the interaction between oscillator and signal circuits which would otherwise be present. (Ref. 8). When tetrodes and pentodes became available, the use of the triode was dropped except as the local oscillator. It was not long, however, before the desirability of more complete separation of oscillator and signal circuits became evident. Multigrid converter tubes were, therefore, devised to permit this separation in a satisfactory manner, at least for the frequencies then in common use. (Refs. 9-14). In some of these it was also possible to control the conversion gain by an automatic-volume-control voltage, a decided advantage. The most satisfactory of the earlier multigrid tubes was known as the pentagrid converter, a type still widely used. A similar tube having an additional suppressor grid is used in Europe and is known as the octode.

When it became desirable to add high-frequency bands to superheterodyne receivers which also had to cover the low broadcast frequencies, the converter problem became more difficult. The highest practicable intermediate frequency appeared to be about 450 to 460 kilocycles, a value which was only about 2 per cent of the highest frequency to be received. Its use meant that the oscillator frequency was separated from the signal frequency by only 2 per cent and the signal circuit, therefore, offered appreciable impedance at the oscillator frequency. A phenomenon known as "space-charge coupling," found in the pentagrid converter, indicated that signal and oscillator circuits were not separated as completely as would be desirable. (Ref. 15). In addition, the permissible frequency variations of the oscillator had to be held to less than the intermediate-frequency bandwidth, namely, 5 to 10 kilocycles ; at the highest frequency to be received, the oscillator frequency was required therefore to remain stable within 0.05 per cent. In the pentagrid converter, the most serious change in oscillator frequency occurred when the automatic-volume-control voltage was changed, and was sometimes as much as 50 kilocycles. Economic considerations have led to the use of at least a three-to-one frequency coverage for each band in the receiver. With capacitance tuning, the circuit impedance is very low at the low-frequency end of the high-frequency band so that failure to oscillate was occasionally observed in the pentagrid converter.

In Europe, where converter problems were similar, a tube known as the triode-hexode (Ref. 16) was developed to overcome some of the disadvantages of the pentagrid converter. In the pentagrid tube, the oscillator voltage is generated by, and therefore applied to, the electrodes of the assembly closest to the cathode (i.e., the *inner* electrodes). In the European form of triode-hexode, the oscillator voltage is generated by a separate small triode section mounted on a cathode common to a hexode-modulator section. The triode grid is connected internally to the third grid of the hexode section. In this way, by the application of the oscillator voltage to an *outer* grid and the signal to the inner grid of the modulator, space-charge coupling was greatly reduced and automatic-volume-control voltage could be applied to the modulator section of the tube without seriously changing the oscillator frequency. In some European types, a suppressor grid has been added so that such tubes should be called triode-heptodes.

The first American commercial development to provide improved performance over that of the pentagrid converter also utilized oscillator voltage injection on an outer grid but required a separate tube for oscillator (Ref. 17). This development, therefore, resulted in a modulator or mixer tube rather than a converter. There were many advantages accompanying the use of separate oscillator tube so that such a solution of the problem appeared to be reasonably satisfactory.

The demand arose shortly, however, for a one-tube converter system with better performance than the original pentagrid type for use in the standard all-wave receiver. A tube, the 6K8, in which one side of a rectangular cathode was used for the oscillator and the other side was used for the mixer section, was developed and made available. (Ref. 18). This tube used inner-grid oscillator injection, as with the pentagrid converter, but had greatly improved oscillator stability. Another solution, also introduced in the United States, was a triode-heptode which is an adaptation of the European triode-hexode. This type used outer-grid injection of the oscillator voltage generated by a small auxiliary triode oscillator section. A recent converter (the SA7 type) for broadcast use is designed to operate with oscillator voltage on both cathode and first-grid electrodes. (Ref. 19). This tube, in addition to having excellent performance, requires one less connecting terminal than previous converter tubes.

This paper will present an integrated picture of the operation of converter and modulator tubes. It will be shown that the general principles of modulating or mixing by placing the signal on one grid and the oscillator voltage on another, or by placing both voltages on the same grid, are the same for all types of tubes. The differences in performance among the various types particularly at high frequencies are due to a number of important secondary effects. In this paper, some of the effects such as signal-grid current at high frequencies, input impedance, space-charge coupling, feedback through interelectrode capacitances, and oscillator-frequency shift will be discussed.

(ii) General analysis of operation common to all types

A. Conversion transconductance of modulator or mixer tubes

The basic characteristic of the converter stage is its conversion transconductance, i.e., the quotient of the intermediate-frequency output current to the signal input voltage. The conversion transconductance is easily obtained by considering the modulation of the local-oscillator frequency by the signal in the tube and, as shown in another paper (Ref. 17) is determined by the transconductance of the signal electrode to the output electrode. The general analysis of a modulator, or mixer tube, is applicable to all mixers no matter how or on what electrodes the oscillator and signal voltages are introduced.

Under the assumption that the signal voltage is very small and the local oscillator voltage large, the signal-electrode transconductance may be considered as a function of the oscillator voltage only. The signal-electrode-to-plate transconductance g_m may, therefore, be considered as periodically varying at the oscillator frequency Such a periodic variation may be written as a Fourier series

$$g_m = a_0 + a_1 \cos \omega_0 t + a_2 \cos 2\omega_0 t + \ldots$$

where ω_0 is the angular frequency of the local oscillator. Use of the cosine series implies that the transconductance is a single-valued function of the oscillator electrode voltage which varies as $\cos \omega_0 t$. When a small signal, $e_s \sin \omega_s t$, is applied to the tube, the resulting alternating plate current to the first order in e_s may be written

$$\begin{aligned} i_p &= g_m e_s \sin \omega_s t \\ &= a_0 e_s \sin \omega_s t + e_s \sum_{n=1}^{\infty} a_n \sin \omega_s t \cos n\omega_0 t \\ &= a_0 e_s \sin \omega_s t + \tfrac{1}{2} e_s \sum_{n=1}^{\infty} a_n \sin (\omega_s + n\omega_0)t \\ &\qquad + \tfrac{1}{2} e_s \sum_{n=1}^{\infty} a_n \sin (\omega_s - n\omega_0)t. \end{aligned}$$

If a circuit tuned to the frequency $(\omega_s - n\omega_0)$ is inserted in the plate, the modulator tube converts the incoming signal frequency ω_s to a useful output at an angular frequency $(\omega_s - n\omega_0)$ which is called the intermediate frequency. Since n is an integer, it is evident that the intermediate frequency, in general, may be chosen to be the difference between the signal frequency and any integral multiple of the local-oscillator frequency ; this is true even though a pure sine-wave local oscillation is applied to the tube. The harmonics of the local-oscillator frequency need only be present in the time variation of the signal-electrode transconductance. The ordinary conversion transconductance is simply a special case when $n = 1$. The conversion transconductance at the nth harmonic of the local oscillator is given by

$$g_{cn} = \frac{i_{\omega_s - n\omega_0}}{e_s} = \frac{a_n}{2}.$$

Substituting the value of the Fourier coefficient a_n it is found that

$$g_{cn} = \frac{1}{2\pi}\int_0^{2\pi} g_m \cos n\omega_0 t d(\omega_0 t).$$

When n is set equal to unity, this expression becomes identical with the one previously derived. (Ref. 17).

Thus, the conversion transconductance is obtained by a simple Fourier analysis of the signal-electrode-to-output-electrode transconductance as a function of time.

Fig. 25.1. Signal - electrode transconductance versus oscillator-electrode voltage for a typical mixer tube. The applied oscillator voltage is shown at A and B is the resulting time variation of transconductance.

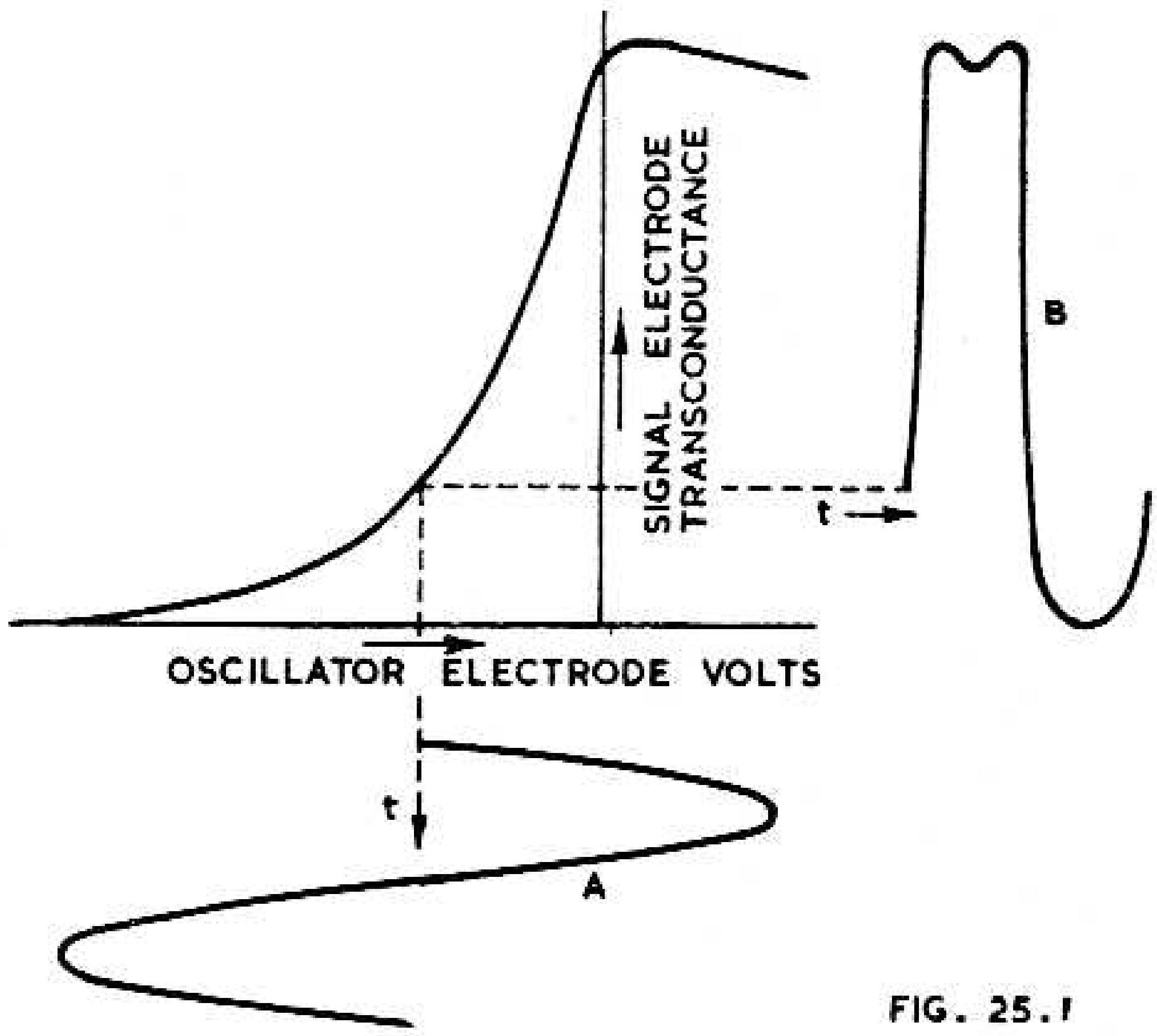


FIG. 25.1

Such an analysis is readily made from the tube characteristics directly by examination of the curve of signal-electrode transconductance versus oscillator-electrode voltage. The calculation of the conversion transconductance at the nth harmonic of the oscillator is made from this curve by assuming an applied oscillator voltage and making a Fourier analysis of the resulting curve of transconductance versus time for its nth harmonic component. The analysis is exactly similar to the one made of power output tubes, except that, in the latter case, the plate-current-versus-control-electrode-voltage curve is used. Fig. 25.1 shows a curve of signal-grid transconductance versus oscillator-electrode voltage for a typical modulator or mixer tube. In the usual case, the oscillator voltage is applied from a tuned circuit and so is closely sinusoidal in shape as at A in the figure. The resulting curve of transconductance versus time is shown at B. Any of the usual Fourier analysis methods may be used to determine the desired component of curve B. Half of this value is the conversion transconductance at the harmonic considered. Convenient formulas of sufficient accuracy for many purposes follow. Referring to Fig. 25.2a, a sine-wave oscillator voltage is assumed and a seven-point analysis is made (i.e., 30-degree intervals). The conversion transconductances g_{cn} are

$$g_{c1} = \frac{1}{12}[(g_7 - g_1) + (g_5 - g_3) + 1.73\,(g_6 - g_2)]$$

$$g_{c2} = \frac{1}{12}[2g_4 + \tfrac{3}{4}(g_3 + g_5 - g_6 - g_2) - (g_7 + g_1)]$$

$$g_{c3} = \frac{1}{12}[(g_7 - g_1) - 2(g_5 - g_3)].$$

The values g_1, g_2, etc., are chosen from the transconductance characteristic as indicated in Fig. 25.2a. The values computed from the above formulas are, of course, most accurate for g_{c1} and of less accuracy for g_{c2} while a value computed from the formula for g_{c3} is a very rough approximation.

Simple inspection of the formula for g_{c1}, the conversion transconductance used for conversion at the fundamental, is somewhat instructive. It is evident that highest conversion transconductance, barring negative values, as given by this formula, occurs when g_1, g_2 and g_3 are all equal to zero, and g_5, g_6, and g_7 are high. These requirements mean that sufficient oscillator voltage should be applied at the proper point to cut off the transconductance over slightly less than the cycle as pictured in Fig. 25.2b. For small oscillator voltages optimum operation requires the differences $(g_7 - g_1)$, $(g_5 - g_3)$ and $(g_6 - g_2)$ to be as large as possible ; this is equivalent to operation at the point of maximum slope. It should be noted that the minimum peak oscillator voltage required for good operation is approximately equal to one half the difference between the oscillator-electrode voltage needed for maximum signal-grid transconductance and that needed to cut off this transconductance. Thus, inspection of the curve of transconductance versus oscillator-electrode voltage gives both a measure of the fundamental conversion transconductance which will be obtained and the amount of oscillator excitation required. Conversion at a harmonic, in general, requires considerably greater oscillator excitation for maximum conversion transconductance.

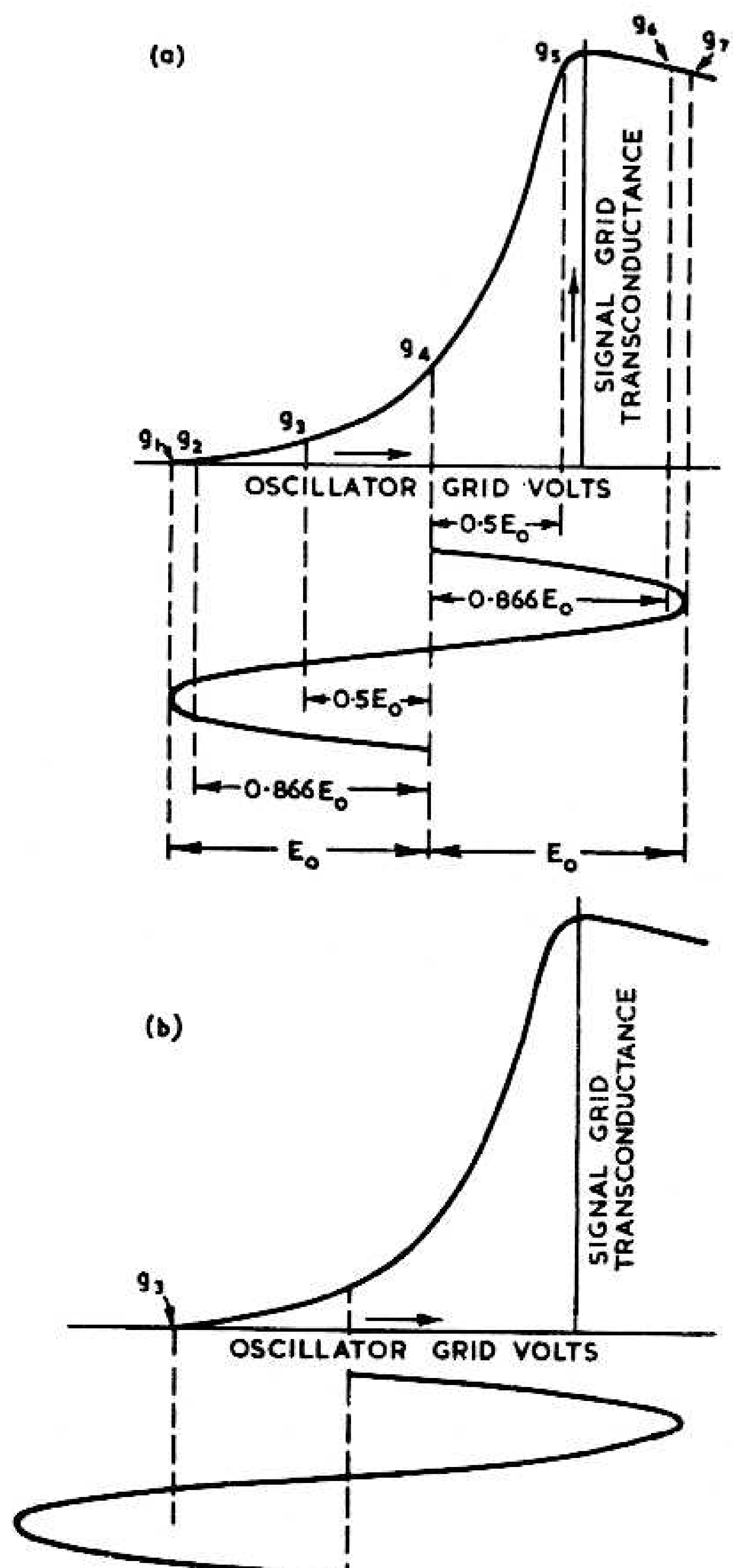


FIG. 25.2

Fig. 25.2a. Points used for 30-degree analysis of conversion transconductance. Fig. 25.2b. Oscillator amplitude and bias adjusted for high conversion transconductance at oscillator fundamental, i.e., $g_1 = g_2 = g_3 = 0$.

In practical cases using grid-controlled tubes of the usual kind, the maximum fundamental conversion transconductance which a given tube will give can quickly be determined within 10 per cent or so by simply taking 28 per cent of the maximum signal-grid-to-plate transconductance which can be attained. For conversion at second harmonic, optimum oscillator excitation gives a conversion transconductance of half this value, while for third-harmonic conversion the value is divided by three.

Although the same characteristic of all modulator or mixer tubes is used to determine the conversion transconductance, the shape of this characteristic varies between different types of mixers. This variation will be more clearly brought out in the later sections of the paper.

B. Conversion transconductance of converter tubes

In converter tubes with oscillator sections of the usual kind, the oscillator voltage is usually present on more than one electrode. Furthermore, the phase of the oscillator-control-grid voltage is opposite to that of the oscillator-anode alternating voltage, so that the two would be expected partially to demodulate each other. The transconductance curve which should be used in this case is the one in which the oscillator electrode voltages are simultaneously varied in opposite directions.

Fortunately, with most of the commonly used converter tubes such as the pentagrid, octode, triode-hexode, etc., the effect of small variations of oscillator-anode voltage on the electrode currents is so small that usually it may be neglected. Thus, the conversion transconductance of these converter tubes may be found exactly as if the tube were a modulator or mixer, only.

With the circuit of Fig. 25.3 (Refs. 19-21) a Hartley oscillator arrangement is used and oscillator-frequency voltage is present on the cathode. The effect of such a voltage is also to demodulate the electron stream through the action of the alternating cathode potential on the screen-to-cathode and signal-grid-to-cathode voltages. When a relatively high-transconductance signal grid is present, as in the figure, this demodulation is considerably greater than in the normal cathode-at-ground circuit.

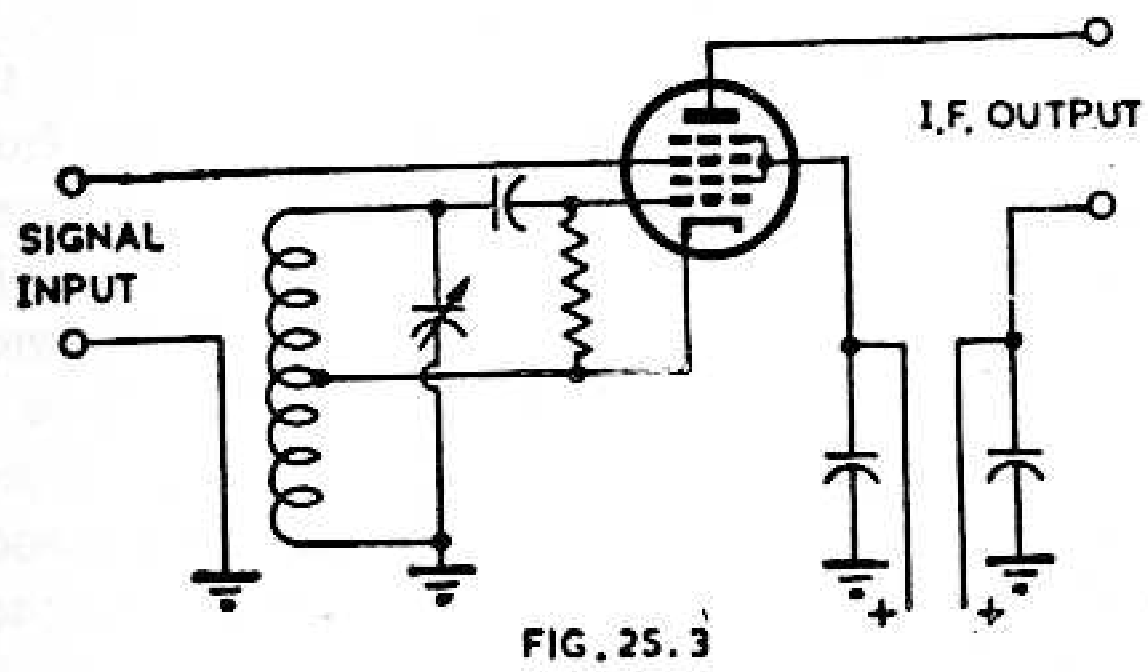


Fig. 25.3. *Converter circuit with oscillator voltage on both grid No.* 1 *and cathode.*

In order to determine the conversion transconductance of a tube to be used in this circuit, a signal-grid transconductance curve is needed. Such a curve, however, must be taken with cathode and oscillator-grid potential varied simultaneously and in their correct ratio as determined by the ratio of cathode turns to total turns of the coil which is to be used. However, because the conversion transconductance is approximately proportional to the peak value of signal-grid transconductance, it is often sufficiently accurate to disregard the alternating-current variation of cathode potential and simply shift the signal-grid in the negative direction by the peak value of the alternating cathode voltage. If the resulting signal-grid-transconductance-versus-oscillator-grid-voltage curve is used for an analysis of conversion transconductance, the data obtained will not be far different from the actual values obtained in the circuit of Fig. 25.3 where normal (unshifted) signal-grid bias values are used.

C. Fluctuation noise

The fluctuation noise of a converter stage is frequently of considerable importance in determining the over-all noise. The magnitude of the fluctuation noise in the output of a converter or mixer tube may be found either by direct measurement using a known substitution noise source such as a saturated diode or by making use of the noise of the same tube used as an amplifier and finding the average mean-squared noise over an oscillator cycle. (Refs. 22, 23). Since these methods give values which are substantially in accord, and since the noise of many of the usual tube types under amplifier conditions is readily derived from theory (Ref. 24), the latter procedure is convenient. Thus, if $\overline{i_{pn}^2}$ is the mean-squared noise current in the output of the converter or mixer tube considered as an amplifier (i.e. steady direct voltages applied) the mean-squared intermediate-frequency noise is

$$\overline{i_{i-f}^2} = \frac{1}{2\pi}\int_0^{2\pi} \overline{i_{pn}^2}\,d(\omega t)$$

or the average of $\overline{i_{pn}^2}$ over an oscillator cycle. The values of $\overline{i_{pn}^2}$ obtained from theory require a knowledge of the currents and transconductance of the tube and are usually proportional to these quantities. Thus, the converter-stage output noise, which is the average of $\overline{i_{pn}^2}$ over the oscillator cycle, is usually proportional to the average electrode currents and average transconductance when the oscillator is applied. Specific examples will be given in following sections of this paper treating typical modes of converter operation.

Tube noise is conveniently treated by use of an equivalent grid-noise-resistance concept whereby the tube noise is referred to the signal grid. The equivalent noise resistance of a converter or mixer tube is

$$R_{eq} = \frac{\overline{i_{i-f}^2}}{(4kT_R\Delta f)g_{cn}^2}$$

where $k = 1.37 \times 10^{-23}$, T_R is room temperature in degrees Kelvin, and Δf is the effective over-all bandwidth for noise purposes. Since Δf is invariably associated with $\overline{i_{i-f}^2}$, the bandwidth cancels in the determination of R_{eq} which is one of the advantages of the equivalent-resistance concept. For $T_R = 20$ degrees centigrade,

$$R_{eq} = 0.625 \times 10^{20} \frac{1}{g_{cn}^2} \frac{\overline{i_{i-f}^2}}{\Delta f}.$$

A summary of values of R_{eq} for common types of converter will be found in a preceding paper. (Ref 23).

The equivalent noise resistance R_{eq} alone does not tell the entire story as regards signal-to-noise ratio, particularly at high frequencies. For example, if the converter stage is the first stage of a receiver, and bandwidth is not a consideration, the signal energy which must be supplied by the antenna to drive it will be inversely proportional to the converter-stage input resistance. On the other hand, the noise energy of the converter or mixer tube is proportional to its equivalent noise resistance. The signal-to-noise ratio therefore, will vary with the ratio of input resistance to equivalent noise resistance, and this quantity should be as high as possible. When bandwidth is important, the input resistance should be replaced by the reciprocal of the input capacitance if it is desired to compare various converter systems for signal-to-noise ratio.

(iii) The oscillator section of converter tubes

The oscillator section of converters is often required to maintain oscillation over frequency ranges greater than three to one for circuits using capacitance tuning. Although this requirement is easily met at the lower broadcast frequencies, the effect of lower circuit impedances, transit-time phenomena in the tube, and high lead reactances combine to make the short-wave band a difficult oscillator problem. Ability to oscillate has, in the past, been measured by the oscillator transconductance at normal oscillator-anode voltage and zero bias on the oscillator grid. Recent data have shown that, in the case of pentagrid and some octode converters, an additional factor which must be considered is the phase shift of oscillator transconductance (i.e. transadmittance) due to transit-time effects*. (Ref. 26).

The ability of a converter to operate satisfactorily at high frequencies depends largely on the undesirable oscillator frequency variations produced when electrode voltages are altered. The frequency changes are mainly caused by the dependence on electrode voltages of oscillator-electrode capacitances, oscillator transconductance, and transit-time effects. There are many other causes of somewhat lesser importance. Because of the complex nature of the problem no satisfactory quantitative analysis is possible. In the case of the pentagrid and the earlier forms of octode converters there are indications that the larger part of the observed frequency shift is due to a transit-time effect. It is found that the phase of the oscillator trans-admittance and, therefore, the magnitude of the susceptive part of this transadmittance varies markedly with screen and signal-grid-bias voltages. Since the susceptive part of the transadmittance contributes to the total susceptance, the oscillation frequency is directly affected by any changes.

(iv) The detailed operation of the modulator or mixer section of the converter stage

This section will be devoted to a consideration of the modulator or mixer portion of the converter stage. This portion may be either a separate mixer tube or the modulator portion of a converter tube. Since with most of the widely used converter tubes in the more conventional circuits the alternating oscillator-anode voltage has a negligible effect on the operation of the modulator portion, only the effect of oscillator control grid need be considered. Thus the analysis of the operation of most converter tubes is substantially the same as the analysis of the same tubes used as a mixer or modulator only, just as in the treatment of conversion transconductance.

*M. J. O. Strutt (see Ref. 26) has published data on this phase shift in octodes. It was measured to be as high as 60 degrees at 33 megacycles.

There are three methods of operation of mixer or modulator tubes. The oscillator voltage may be put on the same grid as the signal voltage, it may be put on the inner grid (the signal applied to an outer grid), or it may be impressed on an outer grid (with the signal on the inner grid). Each of these modes of operation has characteristics which depend on the mode rather than on the tube used in it. Tubes which may be used in any one mode differ from one another mainly in the degree in which they affect these characteristics. The treatment to follow, therefore, will not necessarily deal with specific tube types : instead, the phenomenon encountered will be illustrated by the use of data taken on one or more typical tubes for each of the modes of operation.

A. Tubes with oscillator and signal voltages applied to same grid

Typical tubes used for this type of operation are triodes and pentodes. The oscillator voltage may be introduced in series with the signal voltage, coupled to the signal input circuit inductively, capacitively, and/or conductively, or it may be coupled into the cathode circuit. In all but the last case, by operating below the grid-current point, the oscillator circuit is not loaded directly by the mixer tube. When cathode injection is used, however, an effective load equal to the mean cathode conductance (slightly greater than the mean transconductance) is imposed on the oscillator circuit. The cathode injection circuit has the advantage that oscillator-frequency voltage between the signal input circuit and ground is minimized, thus reducing radiation when the converter stage is also the first stage of the receiver.

A typical transconductance-versus-bias curve for a variable-μ radio-frequency pentode is shown in Fig. 25.4. The use of the Fourier analysis for conversion transconductance at oscillator fundamental indicates that a value of approximately a quarter of the peak transconductance can be attained. Because of the tailing off of the lower end of the curve, highest conversion transconductance requires a large oscillator swing. Very nearly the maximum value is obtained, however, at an operating bias shown by the dotted line, with an oscillator peak amplitude approximately equal to the bias. With lower oscillator amplitudes, and the same fixed bias, the fundamental conversion transconductance drops in approximate proportion to the oscillator amplitude

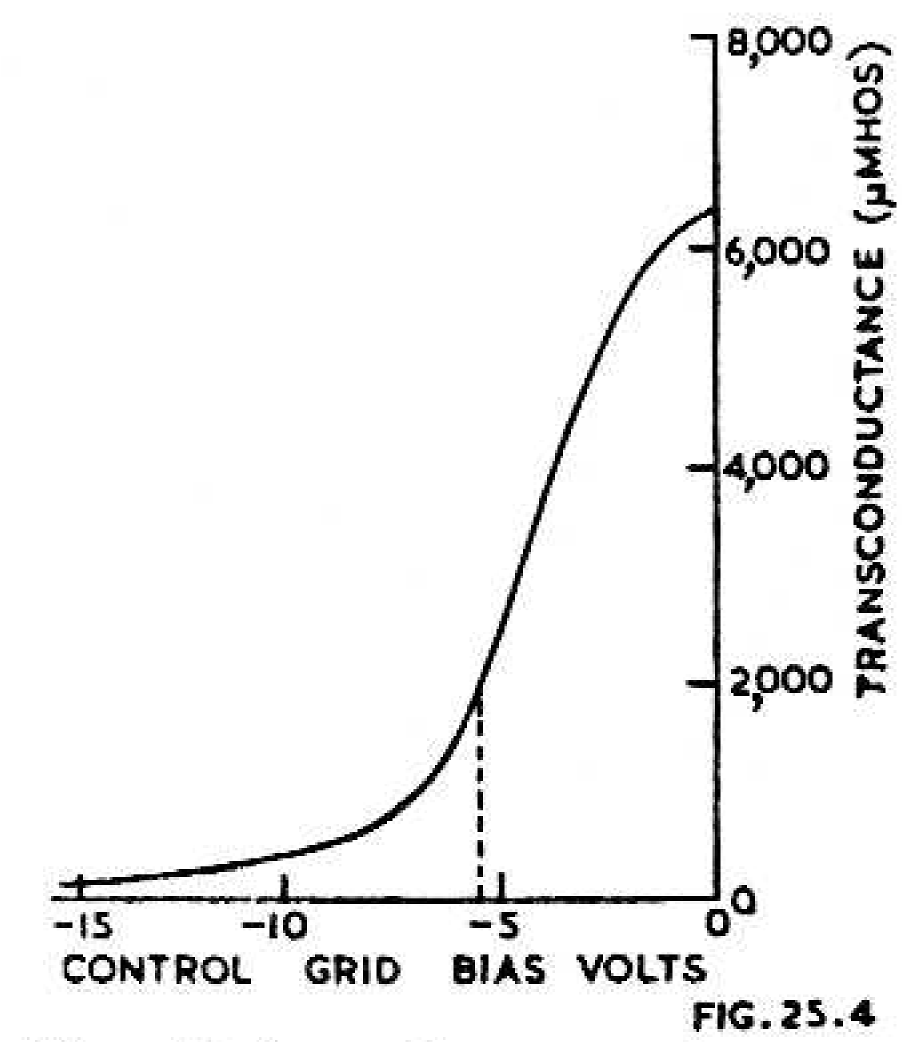


Fig. 25.4. Transconductance characteristic of a typical variable-μ, radio-frequency pentode.

Strictly speaking, when the cathode injection type of operation is used the effect of the oscillator voltage which is impressed between screen and cathode, and plate and cathode should be considered. Practically, however, there is little difference over the simpler circuit in which the oscillator voltage is impressed on the signal grid only. It is for this reason that the cathode-injection circuit is placed in the same category as those in which the oscillator voltage is actually impressed on the same electrode as the signal.

In a practical circuit the effective oscillator voltage is, of course, the oscillator voltage actually existent between grid and cathode of the tube. When the oscillator voltage is impressed in series with the signal circuit or on the cathode, this effective voltage is different from the applied oscillator voltage by the drop across the signal circuit. In the usual case, with the oscillator frequency higher than the signal frequency, the signal circuit appears capacitive at oscillator frequency. This capacitance and the grid-to-cathode capacitance, being in series, form a capacitance divider and reduce the effective oscillator voltage. The reduction would not be a serious matter if it remained a constant quantity ; but in receivers which must be tuned over an appreciable frequency range this is not the case. The result is a variation in conversion gain over the band. A number of neutralizing circuits have been described in the patent literature which are designed to reduce the oscillator-frequency voltage across the signal circuit and thus minimize the variations. (Refs. 27, 28).

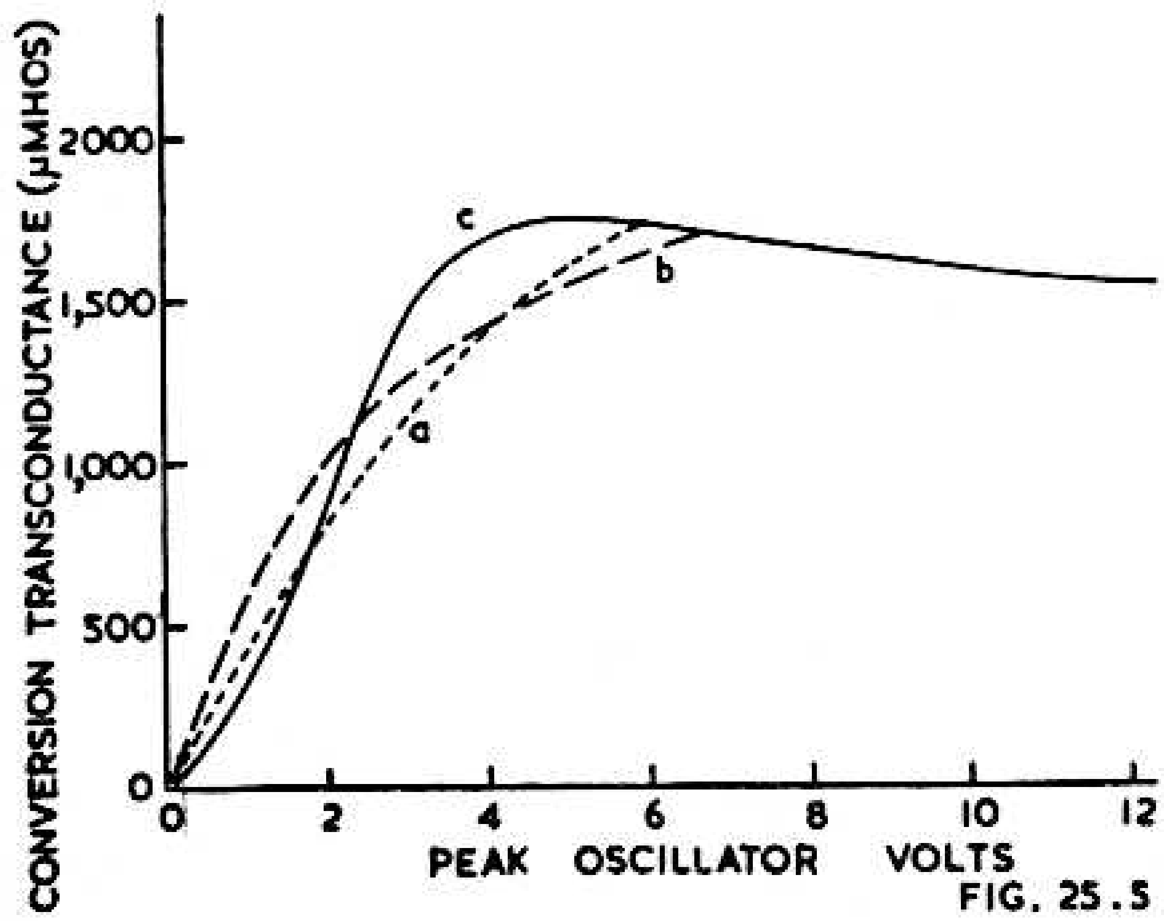
Coupling of the oscillator voltage into, or across, the signal circuit is also accompanied by changes in effective oscillator voltage when the tuning is varied. These changes are not so great with pure inductive coupling as with pure capacitance coupling. In many practical cases, both couplings are present.

A method of reducing the variation of conversion gain with effective oscillator voltage in tubes in which oscillator voltage and signal are placed on the same grid, employs automatic bias. Automatic bias may be obtained either by a cathode self-bias resistor (by-passed to radio frequency) or by a high-resistance grid leak, or both. An illustration of the improvement which may be obtained in this way is shown in Fig. 25.5. Three curves of conversion transconductance, at oscillator fundamental, against effective peak oscillator volts are shown for the typical variable-μ pentode of Fig. 25.4 used as a mixer. For the curve a, a fixed bias was used at approximately an optimum point. The curve is stopped at the grid-current point because operation beyond this point is not practicable in a receiver. Curve b shows the same tube operated with a cathode self-bias resistor. This curve is also stopped at the grid-current point. Curve c shows operation with a high-resistance grid leak. It is evident that, above an oscillator voltage of about 3, curve b is somewhat flatter, and c is considerably flatter than the fixed-bias curve a. The high-resistance grid leak used for c may be made a part of the automatic-volume-control filter but care must be taken that its value is considerably higher than the resistance in the automatic-volume-control circuit which is common to other tubes in the receiver. If this is not done, all the tubes will be biased down with large oscillator swings. When a high-resistance leak is used, the automatic-volume-control action does not begin in the mixer tube until the automatic-volume-control bias has exceeded the peak oscillator voltage. Because of the high resistance of the leak, the signal circuit is not loaded appreciably by the mixer tube. In a practical case, precautions must be taken that a pentode in the converter stage is not operated at excessive currents when accidental failure of the oscillator reduces the bias. A series dropping resistor in the screen-grid supply will prevent such overload. When a series screen resistor is used, the curve of conversion transconductance versus oscillator voltage is even flatter than the best of the curves shown in Fig. 25.5. Series screen operation, therefore, is highly desirable (Ref. 23).

One of the effects of feedback through interelectrode capacitance in vacuum tubes is a severe loading of the input circuit when an inductance is present in the cathode circuit. Thus, in mixers using cathode injection, the signal circuit is frequently heavily damped since the oscillator circuit is inductive at signal frequency in the usual case. The feedback occurs through the grid-to-cathode capacitance and can be neutralized to some extent by a split cathode coil with a neutralizing capacitance. (Ref. 28). Such neutralization also minimizes the voltage drop of oscillator frequency across the signal circuit.

Loading of the signal circuit by feedback from the plate circuit of modulators or mixers may also be serious when the signal-grid-to-plate capacitance is appreciable. This is especially true when a low capacitance intermediate-frequency circuit, which presents a comparatively high capacitive reactance at signal frequency, is used, as in wide-band intermediate-frequency circuits. The grid-plate capacitance of radio-frequency pentodes is usually small enough so that the effect is negligible in these tubes. In triodes, however, feedback from the intermediate-frequency circuit may be serious and the grid-plate capacitance should be minimized in tube and circuit design. Although neutralization is a possible solution to the plate feedback, a more promising solution is the use of a specially designed intermediate-frequency circuit which offers a low impedance at signal frequency by the equivalent of series tuning and yet causes little or no sacrifice in intermediate-frequency performance.

At high frequencies, the converter stage exhibits phenomena not usually observable at low frequencies. One group of phenomena is caused not by the high operating frequency, per se, but rather by a high ratio of operating frequency to intermediate



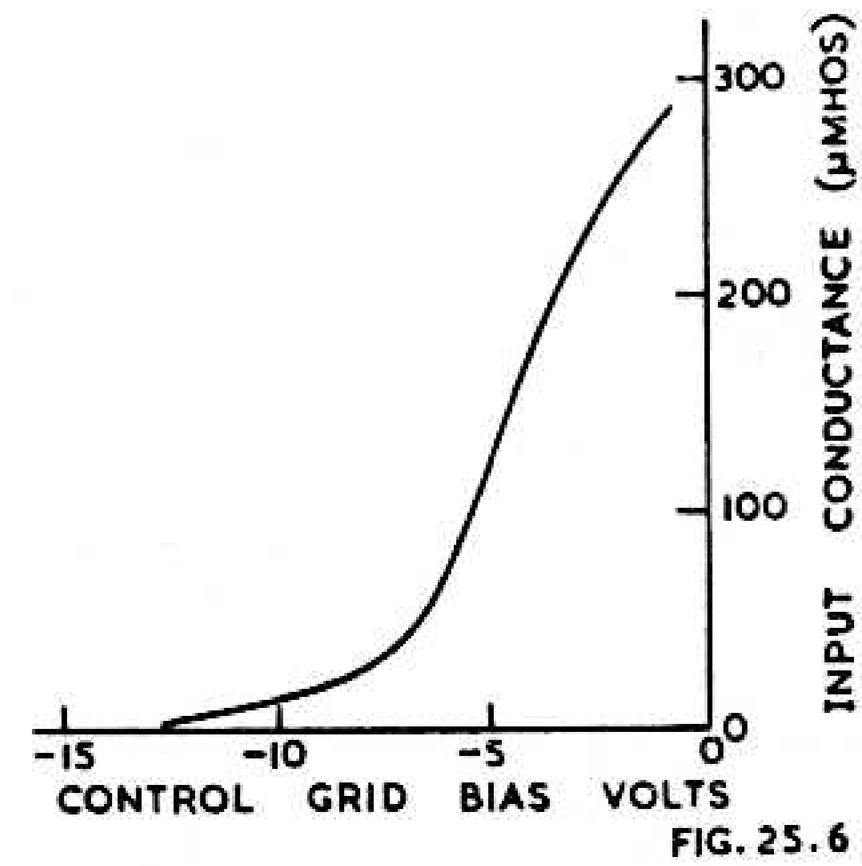


Fig. 25.5. *Conversion transconductance of a typical variable-μ, radio-frequency pentode. Oscillator and signal voltages both applied to grid No.* 1.*a, fixed-bias operation ; b, cathode resistor used to obtain bias ; c, bias obtained by means of a high-resistance grid leak.*
Fig. 25.6. *Input conductance of a typical variable-μ, radio-frequency pentode, at* 60 *megacycles.*

frequency (i.e., a small separation between signal and oscillator frequencies). Among these phenomena may be listed pull-in and inter-locking between oscillator and signal circuits and poor image response. In mixers in which oscillator and signal are impressed on the same grid, the first of these effects is usually pronounced because of the close coupling between the oscillator and signal circuits. It can be reduced by special coupling from the local oscillator at an increase in the complexity of the circuit.

Other phenomena, which are due to the high operating frequency, occur in mixers irrespective of the intermediate-frequency. The most important of these are those caused by transit-time effects in the tube and by finite inductances and mutual inductances in the leads to the tube. When the oscillator and signal are impressed on the grid of a mixer, the effects are not dissimilar to those in the same tube used as an amplifier. So far as the signal is concerned, the operation is similar to that of an amplifier whose plate current and transconductance are periodically varied at another frequency (that of the oscillator). The effects at signal frequency must, therefore, be integrated or averaged over the oscillator cycle. The input conductance at 60 megacycles of the typical radio-frequency pentode used for Figs. 25.4 and Fig. 25.5 as a function of control-grid bias is shown in Fig. 25.6. The integrated or net loading as a function of oscillator amplitude, when the tube is used as a mixer at this frequency, is given in Fig. 25.7, both with fixed-bias operation and with the bias obtained by a grid leak and condenser. The conductance for all other frequencies may be calculated by remembering that the input conductance increases with the square of the frequency. The data given do not hold for cathode injection because of the loading added by feedback, as previously discussed.

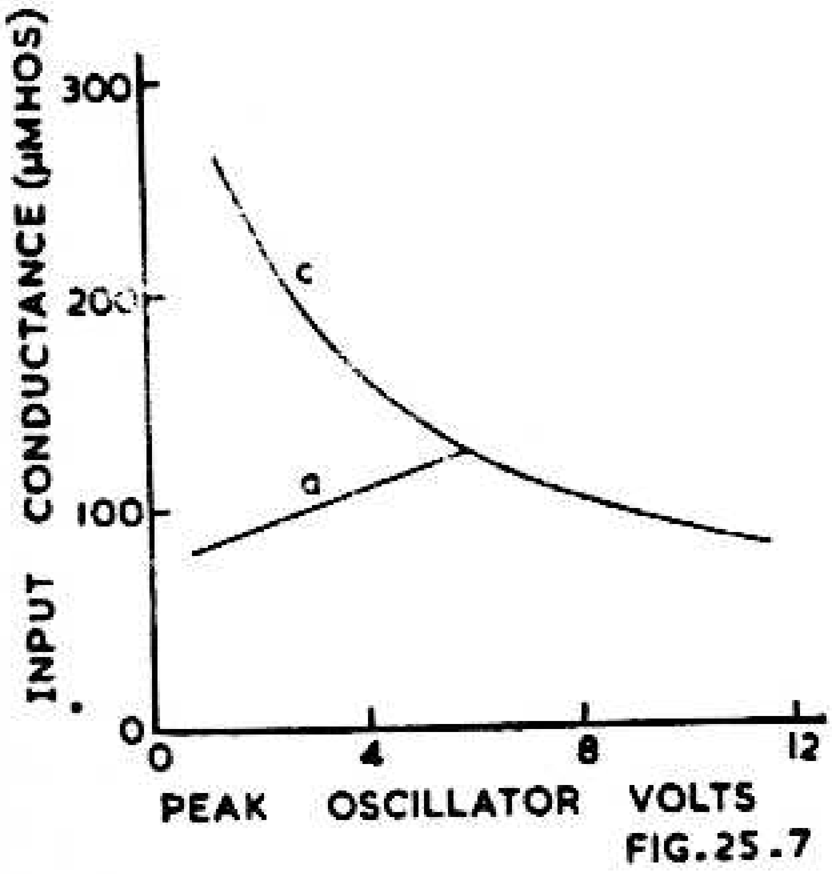


Fig. 25.7. *Input conductance of a typical pentode when used as a mixer at* 60 *megacycles, a, fixed-bias operation ; c, bias obtained by means of a high-resistance grid leak.*

When automatic volume control is used on the modulator tube, an important effect in some circuits is the change in input capacitance and input loading with bias. This is especially true when low-capacitance circuits are in use, as with a wide-band amplifier. With tubes having oscillator and signal voltages on the same grid, because of the integrating action of the oscillator voltage, the changes are not so pronounced as with the same tube used as amplifier. A small, unbypassed cathode resistor may be used with an amplifier tube (Refs. 29, 30) to reduce the variations ; it should give a similar improvement with the modulator.

The question of tube noise (i.e., shot-effect fluctuations) is important in a mixer, or modulator, especially when this tube is the first tube in a receiver. There is little doubt that triode or pentode mixers, in which signal and oscillator voltages are impressed on the control grid, give the highest signal-to-noise ratio of any of the commonly used types of mixers. The reason for this has been made clear by recent studies of tube noise. (Ref. 24). It is now well established that tube noise is the combined result of shot noise in the cathode current which is damped by space charge to a low value and additional fluctuations in the plate current caused by random variations in primary current distribution between the various positive electrodes. Thus, in general, tubes with the smallest current to positive electrodes other than the plate have the lowest noise. It is seen that the tetrode or pentode modulator, with a primary screen current of 25 per cent or less of the total current, is inherently lower in noise than the more complex modulators in which the current to positive electrodes other than the plate usually exceeds 60 per cent of the total current. The triode, of course, has the lowest noise assuming an equivalent tube structure. The conversion transconductance of triode, tetrode, or pentode mixers is usually higher than that of multielectrode tubes using a similar cathode and first-grid structure. That this is so is again largely due to the lower value of wasted current to other electrodes.

The noise of triodes and pentodes used as mixers in the converter stage is conveniently expressed in terms of an equivalent noise resistance R_{eq} as mentioned in Sect. 1(ii)C. The noise as a mixer, of both the triode and the pentode, may be expressed in one formula based on the now well-understood amplifier noise relations. (Ref. 24). The equivalent noise resistance of the triode is obtained simply by equating the screen current to zero. An approximate formula for equivalent noise resistance of oxide-coated-cathode tubes is

$$R_{eq} \text{ (of triode and pentode mixers)} = \frac{2.2\,\overline{g_m} + 20\,\overline{I_{c_2}}}{g_c^{\,2}} \cdot \frac{1}{1+\alpha}$$

where $\overline{g_m}$ is the average control-grid-to-plate transconductance (averaged over an oscillator cycle), $\overline{I_{c2}}$ is the average screen current, g_c is the conversion transconductance, and α is the ratio of the screen current to plate current. Valuable additions to the above relation are given by formulas which enable a simple calculation of noise resistance from amplifier data found in any tube handbook. These additional relations are approximations derived from typical curve shapes and are based on the maximum peak cathode current I_0 and the maximum peak cathode transconductance g_0. The data are given in Table 1. It has been assumed that oscillator excitation is approximately optimum. In this table, E_{c0} is the control-grid voltage needed to cut off the plate current of the tube with the plate and screen voltages applied, and α is the ratio of screen to plate current.

TABLE 1
Mixer Noise of Triodes and Pentodes
(Oscillator and Signal both Applied to Control Grid)

Operation	Approximate Oscillator Peak Volts	Average Transconductance $\overline{g_m}$	Average Cathode Current $\overline{I_k}$	Conversion Transconductance g_c	Equivalent Noise Resistance R_{eq}
At Oscillator Fundamental	$0.7\ E_{c0}$	$\frac{0.47}{1+\alpha} g_0$	$0.35\ I_0$	$\frac{0.28}{1+\alpha} g_0$	$\frac{13}{g_0} + 90 \frac{I_0}{g_0^{\,2}} \alpha$
At Oscillator 2nd Harmonic	$1.5\ E_{c0}$	$\frac{0.25}{1+\alpha} g_0$	$0.20\ I_0$	$\frac{0.13}{1+\alpha} g_0$	$\frac{31}{g_0} + 220 \frac{I_0}{g_0^{\,2}} \alpha$
At Oscillator 3rd Harmonic	$4.3\ E_{c0}$	$\frac{0.15}{1+\alpha} g_0$	$0.11\ I_0$	$\frac{0.09}{1+\alpha} g_0$	$\frac{38}{g_0} + 260 \frac{I_0}{g_0^{\,2}} \alpha$

As an example of the use of the table, suppose it is desired to find the equivalent noise resistance of a particular triode operated as a converter at the oscillator second harmonic. The local oscillator can be permitted to swing the triode mixer grid to zero bias. With a plate voltage of 180 volts and zero bias, the tube data sheet shows a transconductance, $g_0 = 2.6 \times 10^{-3}$ mho. Thus the equivalent noise resistance is $31/g_0$ or 12 000 ohms and the conversion transconductance at second harmonic is 0.13 g_0, or 340 micromhos. Since, with this plate voltage the tube cuts off at about 8 volts, a peak oscillator voltage of around 12 volts will be required.

The above table may also be used to obtain a rough estimate of the input loading of pentode or triode mixers, since the high-frequency input conductance is roughly proportional to the average transconductance $\overline{g_m}$ and to the square of the frequency. Thus, if the loading at any transconductance and frequency is known, the loading as a mixer under the conditions of the table may quickly be computed.

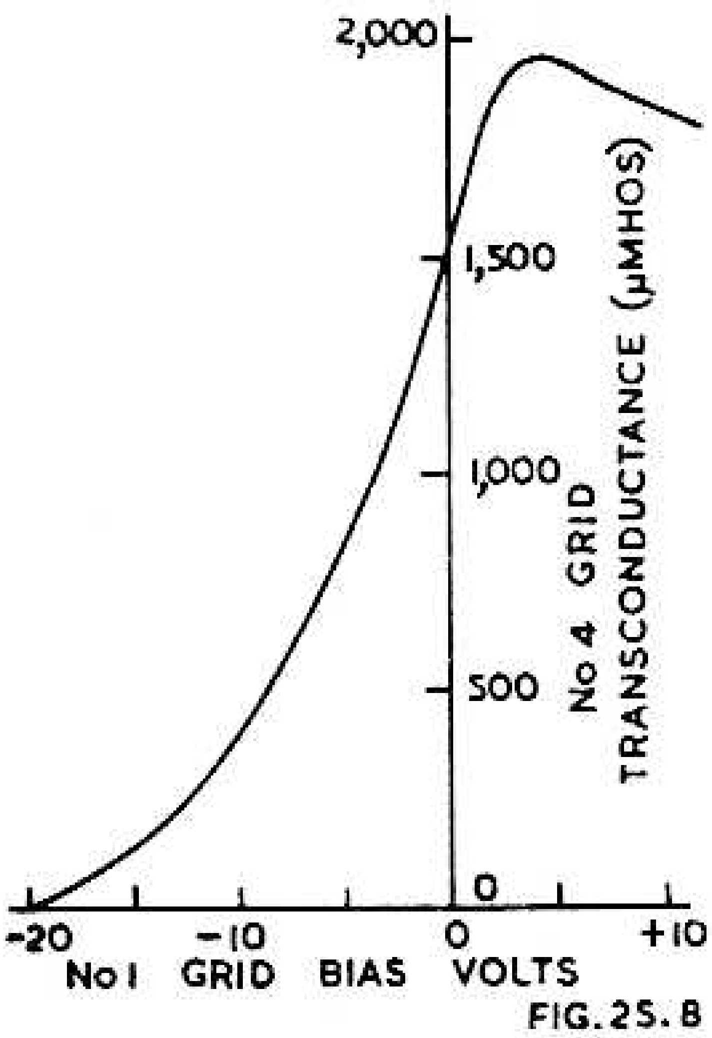


Fig. 25.8. Signal-grid-(grid No. 4) to-plate transconductance versus oscillator-grid (grid No. 1) voltage curve of a typical mixer designed for inner-grid injection. Signal-grid bias = − 3 volts.

B. Tubes with oscillator voltage on an inner grid, signal voltage on an outer grid

When the oscillator voltage is impressed on the grid nearest the cathode of a mixer or converter, the cathode current is varied at oscillator frequency. The signal grid, on the other hand, may be placed later in the electron stream to serve only to change the distribution of the current between the output anode and the other positive electrodes. When the two control grids are separated by a screen grid, the undesirable coupling between oscillator and signal circuits is reduced much below the value which otherwise would be found.

The signal-grid-to-plate transconductance of the inner-grid injection mixer is a function of the total current reaching the signal grid ; this current, and hence the signal-grid transconductance, will vary at oscillator frequency so that mixing becomes possible. The signal-grid transconductance as a function of oscillator-grid potential of a typical modulator of this kind is shown in Fig. 25.8. It will be observed that this characteristic is different in shape from the corresponding curve of Fig. 25.4 for the tube with oscillator and signal voltages on the same grid. The chief point of difference is that a definite peak in transconductance is found. The plate current of the tube shows a saturation at approximately the same bias as that at which the peak in transconductance occurs, indicating the formation of a partial virtual cathode. The signal grid, over the whole of these curves, is biased negatively and so draws no current. The oscillator inner grid (No. 1 grid) however, draws current at positive values of bias. This separation of signal and oscillator grids is advantageous, inasmuch as the signal circuit is not loaded even though the oscillator amplitude is sufficient to draw grid current. In fact, in the usual circuit, the oscillator grid is self-biased with a low-resistance leak and condenser and swings sufficiently far positive to attain the peak signal-grid transconductance.

The conversion transconductance of such a tube has a maximum with an oscillator swing which exceeds the point of maximum signal-grid transconductance in the one direction and which cuts off this transconductance over slightly less than half the cycle, in the other. Curves of conversion transconductance against peak oscillator voltage are shown in Fig. 25.9. Curve *a* is for fixed-bias operation of the oscillator grid, curve *b* is with a high-resistance (i.e. several megohms) grid leak and condenser for bias, and curve *c* is with the recommended value of grid leak (50 000 ohms) for this type of tube. It is seen that best operation is obtained with the lower resistance value of grid leak. With this value, the negative bias produced by rectification in the grid circuit is reduced enough to allow the oscillator grid to swing appreciably positive over part of the cycle. An incidental advantage to the use of the low-resistance leak when the tube is self-oscillating (i.e., a converter) is that undesirable relaxation oscillations are minimized.

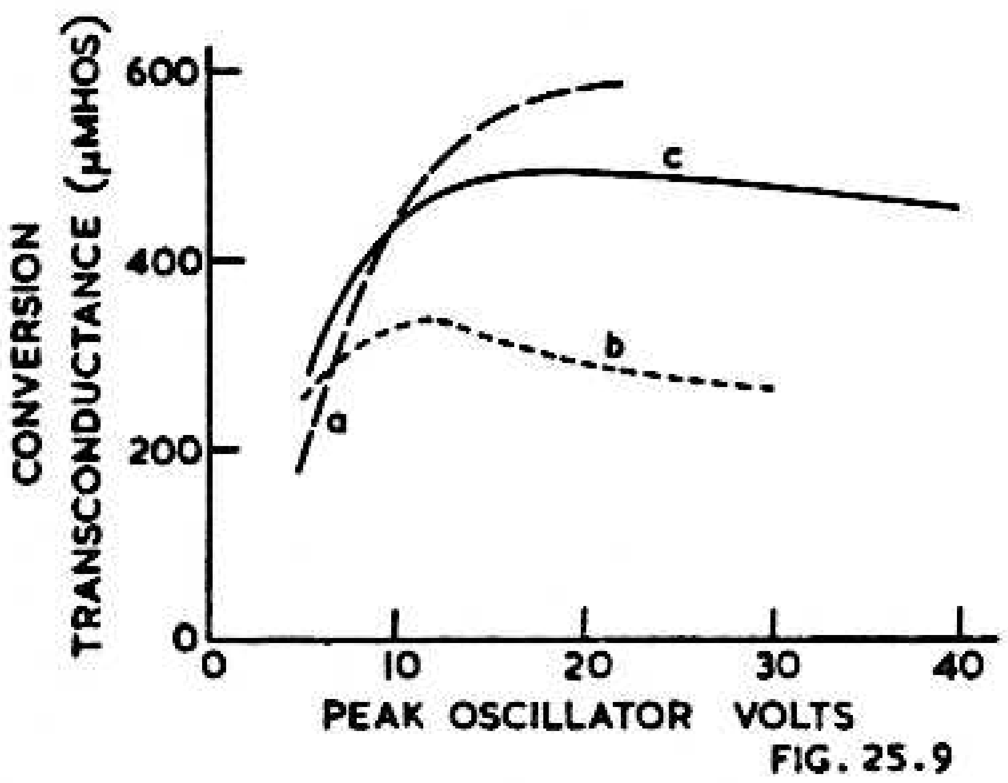


Fig. 25.9. Conversion transconductance of a typical mixer designed for inner-grid injection of oscillator. Signal-grid bias = −3 volts. a, fixed-bias operation of oscillator grid ; b, oscillator-grid bias obtained through high-resistance grid leak ; c, oscillator-grid bias obtained through a 50,000-*ohm grid leak.*

In mixers or converters in which the oscillator voltage is present on both the cathode and the oscillator grid in the same phase (e.g. Fig. 25.3) it is usually necessary to utilize a relatively sharp cut off in the design of the oscillator grid so as to cut off the cathode current when the signal grid is positive (Ref. 19). By this means, the signal grid is prevented from drawing current. At the same time, however, the high currents needed for a high peak value of signal-grid transconductance cannot be obtained without a greater positive swing of the oscillator grid than with a more open oscillator grid structure. Thus, it is clear that it is desirable to have a negative bias on the oscillator electrode which is considerably smaller than the peak oscillator voltage. For this reason, optimum results are obtained on these tubes with very low values of oscillator grid leak (e.g. 10 000 to 20 000 ohms).

The effects of feedback through the interelectrode capacitance are small in well-designed multigrid mixers and converters of the kind covered in this section. The signal-grid-to-plate capacitance is usually small enough to play no part in the operation ; even with a high L-to-C ratio in the intermediate-frequency transformer, the capacitive reactance of the intermediate-frequency circuit at signal frequency is only a very small fraction of the feedback reactance. The other interelectrode capacitance which plays some part in determining circuit performance (excluding, of course, the input and output capacitances) is the capacitance from the oscillator electrode or electrodes to the signal grid. This capacitance is a source of coupling between these two circuits. In well-designed converter or modulator tubes of the type discussed in this section, however, the coupling through the capacitance may be made small compared with another form of internal coupling known as "space-charge coupling" which will be treated later in this discussion.

Coupling between oscillator and signal circuits is of no great consequence except when an appreciable voltage of oscillator frequency is built up across the signal-grid circuit. This is not usually possible unless the signal circuit is nearly in tune with the oscillator as it is when a low ratio of intermediate frequency to signal frequency is used. The effect of oscillator-frequency voltage induced across the signal circuit depends on its phase ; the effect is usually either to increase or to decrease the relative modulation of the plate current at oscillator frequency and so to change the conversion transconductance. This action is a disadvantage, particularly when the amount of induced voltage changes when the tuning is varied, as usually occurs.

In some cases, another effect is a flow of grid current to the signal grid ; this may happen when the oscillator-frequency voltage across the signal-grid circuit exceeds the bias. Grid current caused by this effect can usually be distinguished from grid current due to other causes. By-passing or short-circuiting the signal-grid circuit reduces the oscillator-frequency voltage across the signal-grid circuit to zero. Any remaining grid current must, therefore, be due to other causes.

Current to a negative signal grid of a tube operated with inner-grid oscillator injection is sometimes observed at high frequencies (e.g. over 20 megacycles) even when no impedance is present in the signal-grid circuit. This current is caused by electrons whose effective initial velocity has been increased by their finite transit time in the high-frequency alternating field around the oscillator grid. These electrons are then able to strike a signal grid which is several volts negative. The magnitude of the signal-grid current is not usually as great as with tubes applying the oscillator voltage to an outer grid* although it may prevent the use of an automatic-volume-control voltage on the tube.

An investigation of coupling effects in the pentagrid converter showed that the coupling was much larger than could be explained by interelectrode capacitance. It was furthermore discovered that the apparent coupling induced a voltage on the signal circuit in opposite phase to that induced by a capacitance from oscillator to signal grid. (Ref. 15). The coupling which occurred was due to variations in space charge in front of the signal grid at oscillator frequency. A qualitative explanation for the observed behaviour is that, when the oscillator-grid voltage is increased, the electron charge density adjacent to the signal grid is increased and electrons are repelled from the signal grid. A capacitance between the oscillator grid and the signal grid would have the opposite effect. The coupling, therefore, may be said to be approximately equivalent to a negative capacitance from the oscillator grid to the signal grid. The effect is not reversible because an increase of potential on the signal grid does not increase the electron charge density around the oscillator grid. If anything, it decreases the charge density. The equivalence to a negative capacitance must be restricted to a one-way negative capacitance and, as will be shown later, is restricted also to low-frequency operation.

In general, the use of an equivalent impedance from oscillator grid to signal grid to explain the behaviour of " space-charge coupling " is somewhat artificial. A better point of view is simply that a current is induced in the signal grid which depends on the oscillator-grid voltage. Thus, a transadmittance exists between the two electrodes analogous to the transconductance of an ordinary amplifier tube. Indeed, the effect has been used for amplification in a very similar manner to the use of the transconductance of the conventional tube. (Ref. 32, 33).

It is found that the transadmittance from the oscillator to the signal electrode Y_{mo-s} is of the form

$$Y_{mo-s} = k_1\omega^2 + jk_2\omega.$$

At low frequencies (i.e., $k_1\omega^2 \ll k_2\omega$) the transadmittance is mainly a transusceptance but, as the frequency rises, the transconductance component $k_1\omega^2$ becomes of more and more importance, eventually exceeding the transusceptance in magnitude. The early work on " space-charge coupling " indicated that the effect was opposite to that of a capacitance connected from oscillator to signal grid and could be cancelled by the connection of such a capacitance of the correct value (Refs. 15, 34). The effect of such cancellation could be only partial, however, since only the transusceptance was balanced out by this arrangement. For complete cancellation it is also necessary to connect a conductance, the required value of which increases as the square of the frequency, between the oscillator grid and the signal grid so that the transconductance term is also balanced out. (Ref. 35, 36).

The cancellation of " space-charge coupling " may be viewed in another way. A well-known method of measuring the transadmittance of a vacuum tube is to connect an admittance from control grid to output electrode and to vary this admittance until no alternating-current output is found with a signal applied to the control grid.

*The next part of this section contains a more detailed discussion of signal-grid current in outer-grid oscillator injection tubes.

(Ref. 37). The external admittance is then equal to the transadmittance. In exactly the same way, the transadmittance which results from the space-charge coupling may be measured. As a step further, if an admittance can be found which substantially equals the transadmittance at all frequencies or over the band of frequencies to be used, this admittance may be permanently connected so as to cancel the effects of space-charge coupling. As has been previously stated, the admittance which is required is a capacitance and a conductance whose value varies as the square of the frequency. Such an admittance is given to a first approximation by the series connection of a capacitance C and a resistance R. Up to an angular frequency $\omega = 0.3/CR$ the admittance of this combination is substantially as desired. At higher values of frequency, the conductance and susceptance fail to rise rapidly enough and the cancellation is less complete. Other circuits are a better approximation to the desired admittance. For example, the connection of a small inductance, having the value $L = 1/2CR^2$, gives a good approximation up to an angular frequency $\omega = 0.6/CR$. The latter circuit is, therefore, effective to a frequency twice as high as the simple series arrangement of capacitance and resistance. Inasmuch as in some cases the value of inductance needed is only a fraction of a microhenry, the inductance may conveniently be derived from proper proportioning and configuration of the circuit leads.

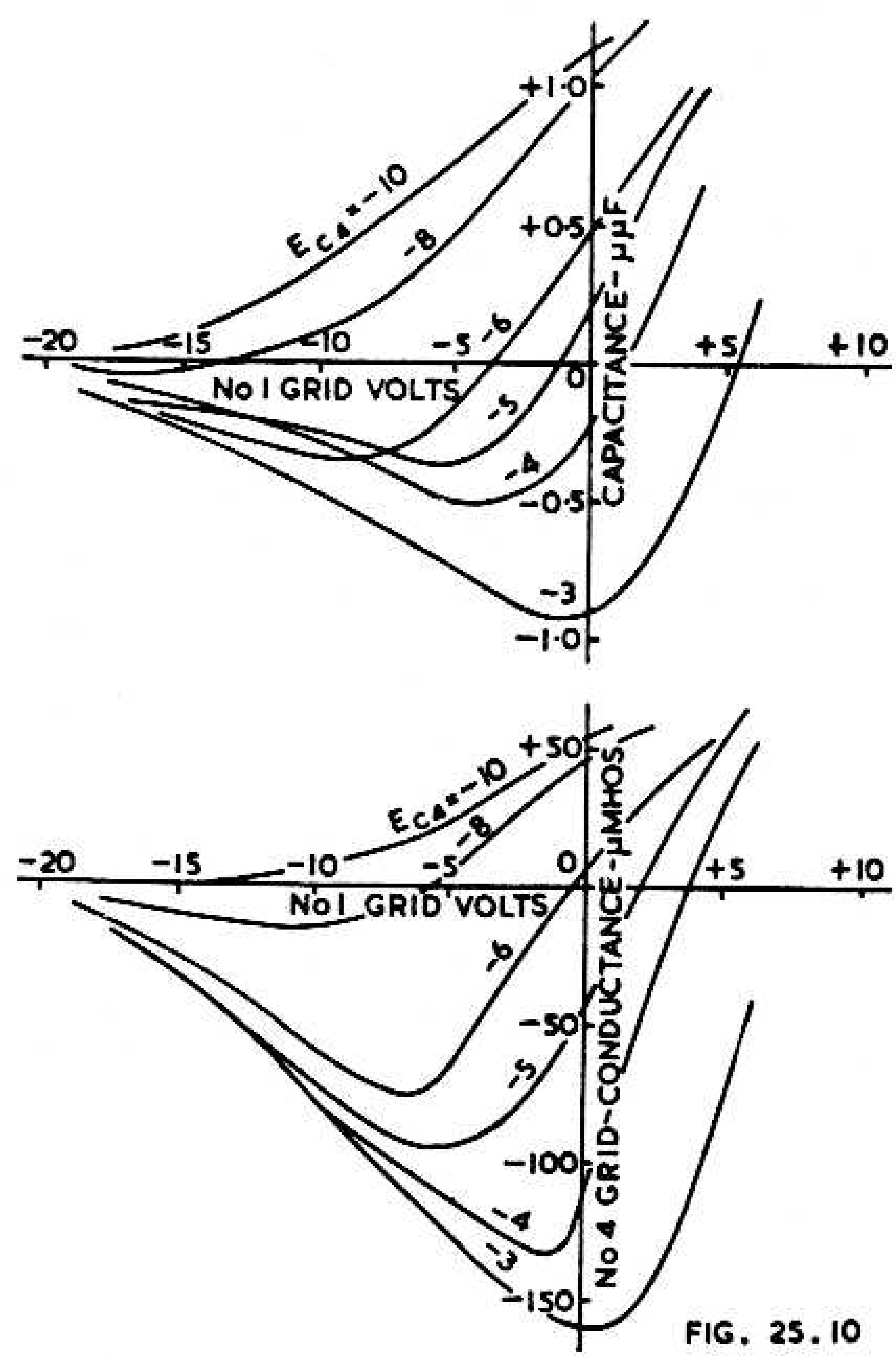


FIG. 25.10

Fig. 25.10. Signal grid (grid No. 4) admittance of a typical mixer designed for inner-grid injection of oscillator at 31.5 megacycles. Curves taken with no oscillator voltage applied. Data represents electronic admittance only (i.e. " cold " values were subtracted from measured values before plotting.)

It is of interest to note the order of magnitude of the transadmittance which is measured in the usual converter and mixer tubes. (Refs. 26, 35, 36, 38). In the formula for Y_{mo-s} given above, k_1 is in the neighborhood of 10^{-21} and k_2 is around 10^{-12}. Cancellation is effected by a capacitance of the order of one or two micromicrofarads and a series resistance of 500 to 1000 ohms.

The correct value of the cancelling admittance may be found experimentally by adjustment so that no oscillator voltage is present across the signal-grid circuit when the latter is tuned to the oscillator frequency. Another method which may be used is to observe either the mixer or converter plate current or the oscillator grid current as the tuning of the signal is varied through the oscillator frequency. With proper adjustment of the cancelling admittance there will be no reaction of the signal-circuit tuning on either of these currents.

There are two disadvantages which accompany the cancellation of space-charge coupling as outlined. In the first place, the signal-grid input admittance is increased by the cancelling admittance. This point will be brought up again after discussing the input admittance. The second disadvantage is that the oscillator frequency shift with voltage changes in converter tubes may be somewhat increased by the use of this cancellating admittance. When separate oscillator and mixer tubes are used, the latter effect may be made less serious.

The next point to be considered is the input admittance of the signal grid. Signal-grid admittance curves of a typical modulator designed for use with the oscillator voltage impressed on the first grid are shown under direct-current conditions (i.e., as a function of oscillator-grid bias for several values of signal-grid bias) in Fig. 25.10. The admittance is separated into conductive and susceptive components, the latter being plotted in terms of equivalent capacitance. The admittance components of the " cold " tube (no electrons present) have been subtracted from the measured value so that the plotted results represent the admittance due to the presence of electrons only. The data shown were taken at 31.5 megacycles with a measuring signal which did not exceed 1.0 volt peak at any time. A modified Boonton Q meter was used to take the data. It should be noted that the presence of a marked conductive component of admittance is to be expected at frequencies as high as those used.

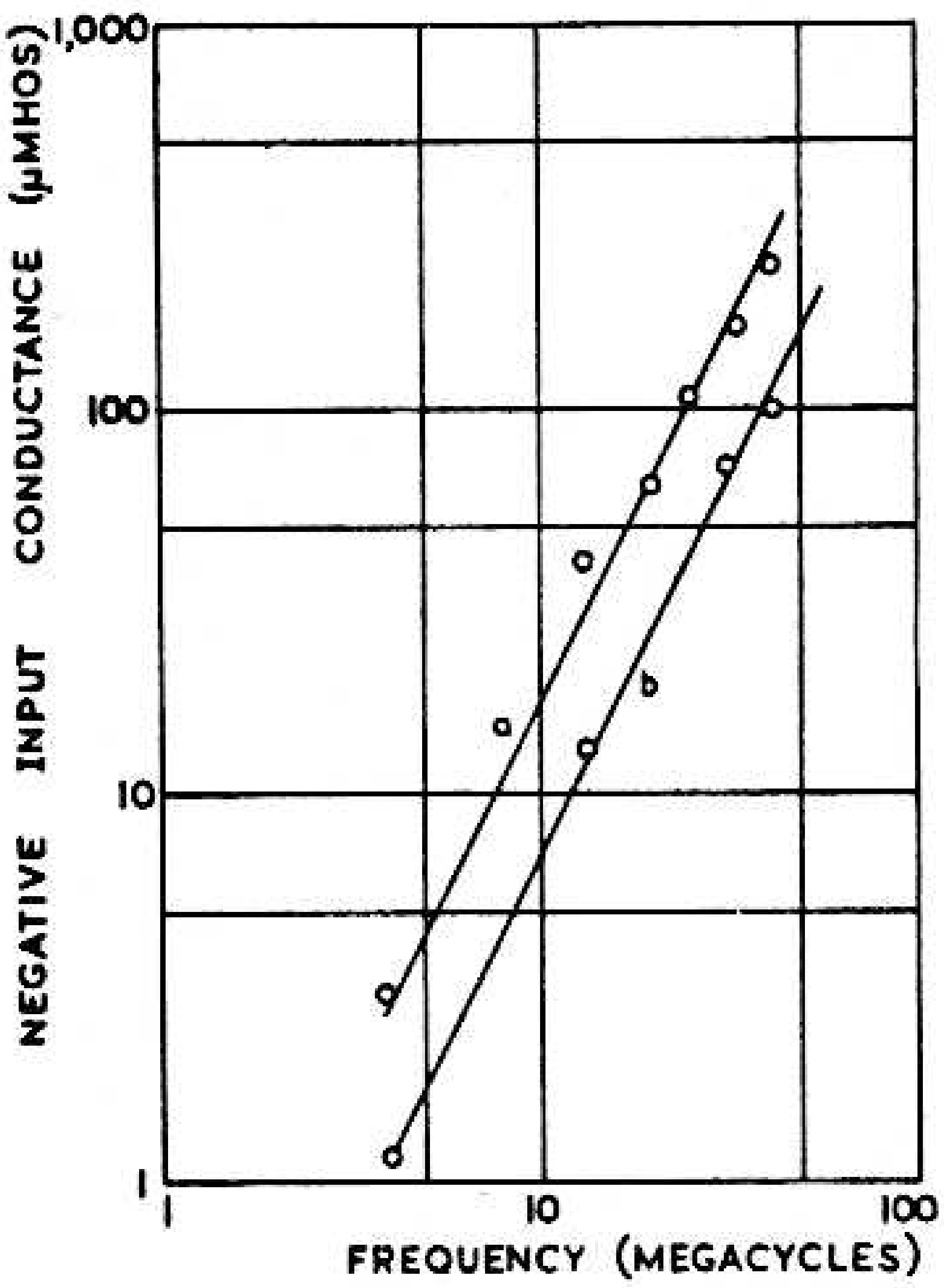


FIG. 25.11

Fig. 25.11. *Signal-grid* (*grid No.* 4) *conductance of a typical mixer designed for inner-grid injection of oscillator. Lines are drawn with slope of* 2. *Curve a taken with* $E_{c_1} = 0$, $E_{c_4} = -3$ *volts. Curve b taken with* $E_{c_1} = -6$, $E_{c_4} = -6$ *volts.*

The most striking feature of the data of Fig. 25.10 is that both susceptive and conductive components are negative over a large portion of the characteristic. The Appendix discusses this feature in somewhat more detail. The measurements show that the susceptive component is analogous to a capacitance. The capacitance curves given are independent of frequency up to the highest frequency used (approximately 50 megacycles). The conductive component, on the other hand, increases as the square of the frequency also up to this frequency. The conductance is, therefore, negative even at very low frequencies although its magnitude is then very small. Thus, the conductance curves of Fig. 25.11 are valid for any frequency by multiplication of the conductance axis by the square of the ratio of the frequency considered, to the frequency used for the data (i.e., 31.5 megacycles). Data taken at various frequencies for two particular values of grid bias voltage E_{c4} are plotted in Fig. 25.11. The square-law relation is shown to check very closely.

Fig. 25.10 should be considered remembering that the oscillator voltage is applied along the axis of abscissas. Considering an applied oscillator voltage, the admittance curves must be integrated over the oscillator cycle to find the admittance to the signal frequency. The operation is just as if the tube were an amplifier whose input admittance is periodically varied over the curve of Fig. 25.10 which corresponds to the signal-grid bias which is used. Curves of the modulator input conductance at 31.5 megacycles for various applied oscillator voltages are shown in Fig. 25.12. The oscillator-grid bias is obtained by means of the recommended value of grid leak for the tube (50 000 ohms). Curves are shown for two values of signal-grid bias voltage E_{c4}. As before, data for other frequencies are obtained by multiplying the conductance by the square of the frequency ratio.

The practical effect of the negative input admittance in a circuit is due to the conductive portion only, inasmuch as the total input capacitance remains positive in general*. An improved image ratio, and somewhat greater gain to the converter signal grid over other types of modulator is to be expected when this type of oscillator injection is used. At high frequencies, when a comparatively low intermediate frequency is used, it is usually desirable to cancel the space-charge coupling of the tube in the manner previously discussed. When this cancellation is made reasonably complete by the use of a condenser and resistor combination connected from the oscillator grid to the signal grid, the losses in this admittance at signal frequency are usually sufficient to wipe out the negative input admittance. The net positive input conductance however is often less than that found with other types of mixer.

The change in signal-grid input capacitance with automatic-volume-control is small in this type of modulator, particularly with the larger values of oscillator swing because of the integrating action of the oscillator voltage.

The fluctuation noise which is found in the output of inner-grid oscillator-injection mixers and converters is not readily evaluated quantitatively. The fluctuation noise is primarily due to current-distribution fluctuations but is complicated by the possibility of a virtual cathode ahead of the signal grid. Data have been taken, however, which indicate some degree of proportionality between the mean-square noise current and the plate current. The signal-to-noise ratio for this type of modulator is, therefore, approximately proportional to the ratio of conversion transconductance to the square root of the plate current. It is considerably less than for the pentode modulator with both signal and oscillator voltages on the control grid.

The noise of the converter or mixer with oscillator on an inner grid may be expressed in terms of an equivalent grid resistance as

$$R_{eq} = \frac{20\bar{I}_b}{g_c^2} F^2$$

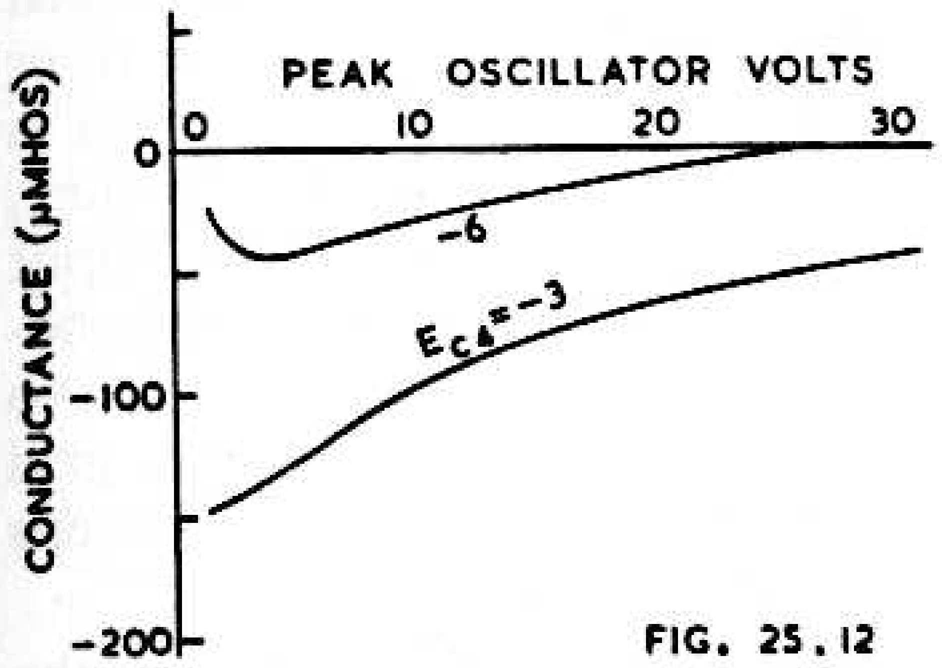
where $\bar{I}_b$ is the operating plate current, g_c is the conversion transconductance, and F^2 is a factor which is about 0.5 for tubes with suppressor grids and at full gain. For tubes without suppressor or for tubes whose gain is reduced by signal-grid bias, F^2 is somewhat larger and approaches unity as a maximum. With this mode of operation there is not so much value in expressions for R_{eq} based on maximum transconductance and maximum plate current because these quantities are neither available nor are they easily measured. For operation at second or third harmonics of the oscillator (assuming optimum oscillator excitation) the plate current $\bar{I}_b$ and the conversion transconductance g_c are roughly ½ or 1/3, respectively, of their values with fundamental operation so that the equivalent noise resistance for second-harmonic and third-harmonic operation is around two and three times, respectively, of its value for fundamental operation.

C. Mixers with oscillator voltage on an outer grid, signal voltage on inner grid

With this type of mixer, the cathode current is modulated by the relatively small signal voltage which is impressed on the control grid adjacent to the cathode. The oscillator voltage, on the other hand, is impressed on a later control grid so that it

*It should not be forgotten that the data given do not include the " cold " susceptance and conductance of the tube. The latter is a relatively small quantity, however.

periodically alters the current distribution between anode and screen grid. The connections of signal and oscillator voltages to this type of modulator are just the reverse, therefore, of the mixer treated in the preceding section. The behaviours of the two types are also quite different although they both include internal separation of signal and oscillator electrodes through a shielding screen grid.



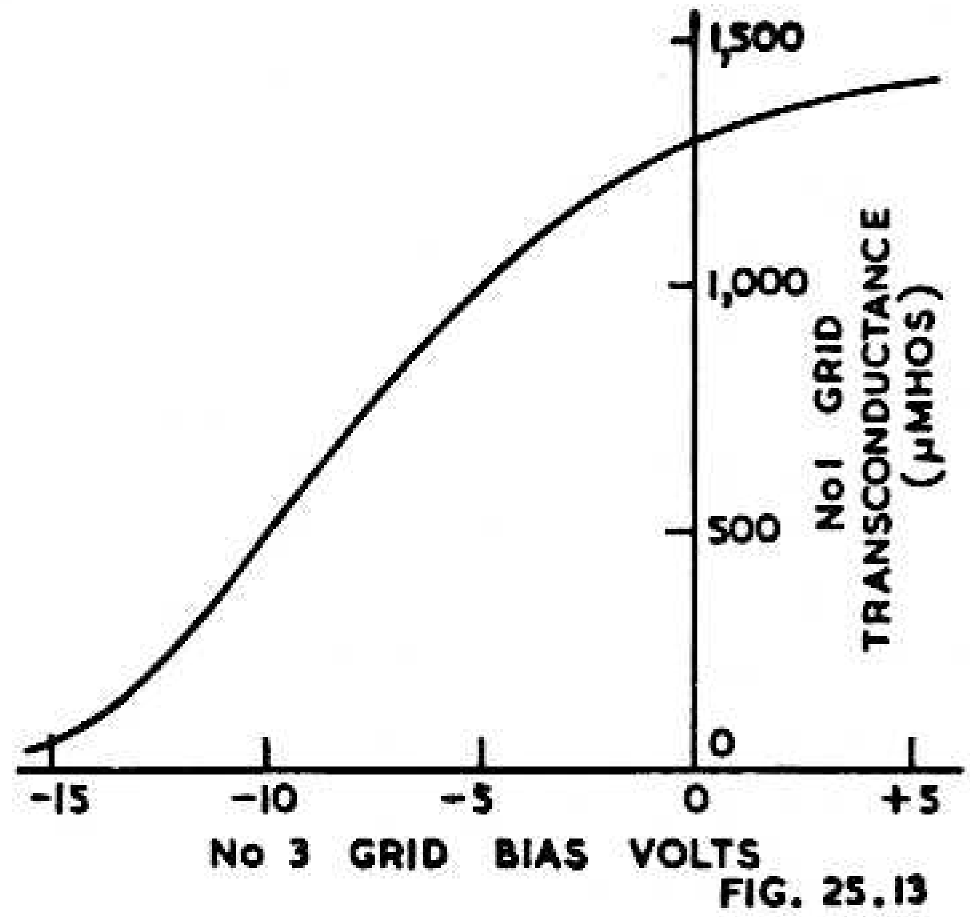


Fig. 25.12. *Signal-grid (grid No.* 4) *conductance of a typical mixer designed for inner-grid injection of oscillator, at* 31.5 *megacycles. Oscillator voltage applied. Oscillator-grid bias obtained through* 50,000-*ohm grid leak. Electronic portion of conductance, only, plotted.*

Fig. 25.13. *Signal-grid (grid No.* 1) *transconductance versus oscillator-grid (grid No.* 3) *voltage of a typical mixer designed for use with outer-grid injection of oscillator. Signal-grid bias* = −3 *volts.*

The signal-grid transconductance curve as a function of oscillator-grid voltage of a typical mixer designed for use with the oscillator on an outer grid is shown in Fig. 25.13. It differs in shape from similar curves for the other two classes of modulator in that an approximate saturation is reached around zero bias on the oscillator grid. The conversion transconductance for such a tube is, therefore, more accurately predicted from normal amplifier transconductance. In fact, in the manufacture of this type of mixer, a test of signal-grid transconductance at somewhere near the saturation point (e.g., zero bias) on the oscillator grid has been found to correlate almost exactly with the conversion transconductance. The cut off point of the curve must remain approximately fixed, of course, since this point affects the oscillator amplitude which is necessary.

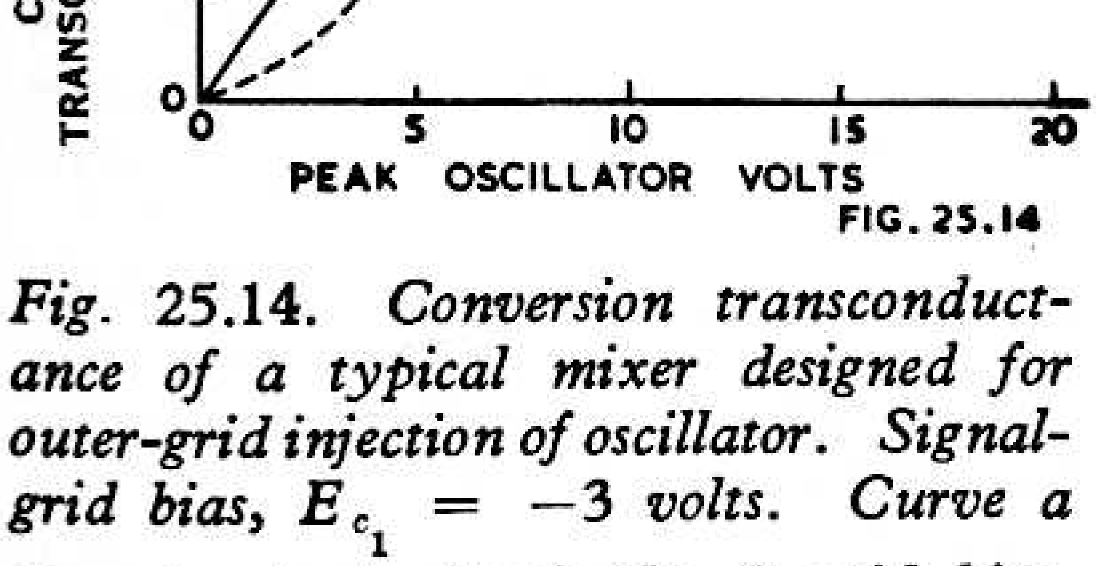


Fig. 25.14. *Conversion transconductance of a typical mixer designed for outer-grid injection of oscillator. Signal-grid bias,* $E_{c_1} = -3$ *volts. Curve a corresponds to fixed No.* 3 *grid bias,* $E_{c_3} = -8$ *volts. Curve b corresponds to bias obtained through a* 50 000-*ohm grid leak.*

The conversion transconductance of the typical outer-grid injection mixer tube which was used for Fig. 25.13 is shown in Fig. 25.14. Curve *a* which is for fixed bias on the oscillator grid is seen to be higher than curve *b* for which bias is obtained by a 50 000-ohm grid leak and condenser. The latter connection is most widely used, however, because of its convenience. A compromise using fixed bias together with a grid leak is most satisfactory of all. (Ref. 40). When this combination is used, the curve of conversion transconductance follows curve *a* of Fig. 25.14 to the intersection with curve *b* and then follows along the flat top of curve *b*.

In a well-designed mixer with the signal voltage on the grid adjacent to the cathode and the oscillator voltage on an outer grid, effects due to feedback through the inter-electrode capacitance may usually be neglected. The only effect which might be of importance in some cases is coupling of the oscillator to the signal circuit through the signal-grid-to-oscillator-grid capacitance. In many tubes a small amount of space-charge coupling between these grids is also present and adds to the capacitance coupling (contrary to the space-charge coupling discussed in Section B which opposes the capacitance coupling in that case). Measurements of the magnitude of the space-charge coupling for this type of modulator show that it is of the order of 1/5 to 1/10 of that present in inner-grid-injection modulators. Coupling between oscillator and signal circuits causes a voltage of oscillator frequency to be built up across the signal input circuit. This oscillator-frequency voltage, depending on its phase, aids or opposes the effect of the normal oscillator-grid alternating voltage. The action is additive when the signal circuit has capacitative reactance to the oscillator frequency, as in the usual case. When the oscillator-frequency voltage across the signal input circuit exceeds the bias, grid current is drawn to the signal grid, an undesirable occurrence. This grid current may be distinguished from signal-grid current due to other causes by short-circuiting the signal-input circuit and noting the change in grid current. With the majority of tubes, another cause of signal-grid current far exceeds this one in importance. This other cause will now be discussed.

The most prominent high-frequency effect which was observed in mixers of the kind under discussion, was a direct current to the negative signal grid even when no impedance was present in this grid circuit. This effect was investigated and found to be due to the finite time of transit of the electrons which pass through the signal grid and are repelled at the oscillator grid, returning to pass near the signal grid again. (Refs. 17, 41, 42). When the oscillator frequency is high, the oscillator-grid potential varies an appreciable amount during the time that such electrons are in the space between screen grid and oscillator grid. These electrons may, therefore, be accelerated in their return path more than they were decelerated in their forward path. Thus, they may arrive at the signal grid with an additional velocity sufficient to allow them to strike a slightly negative electrode. Some electrons may make many such trips before being collected; moreover, in each trip their velocity is increased so that they may receive a total increase in velocity equivalent to several volts. A rough estimate of the grid current to be expected from a given tube is given by the semi-empirical equation

$$I_{c1} = A I_k E_{osc} \omega \tau_{2-3} \epsilon^{BE_{c1}}.$$

Where A and B depend on electrode voltages and configuration, I_{c1} is the signal-grid current, E_{c1} is the signal-grid bias, I_k is the cathode current, E_{osc} is the impressed oscillator voltage on the oscillator grid, ω is the angular frequency of the oscillator, and τ_{2-3} is the electron transit time in the space between screen grid and oscillator grid.

Data on the signal-grid current of a typical mixer at 20 megacycles are shown in Fig. 25.15 where a semi-logarithmic plot is used to indicate the origin of the above equation.

The reduction of signal-grid current by operation at more negative signal-grid bias values is an obvious remedy. When this is done, in order to prevent a reduction in conversion transconductance, the screen voltage must be raised. A better method of reducing the undesired grid current lies in a change of tube design. It will be shown in a later part of this discussion that the constant A and/or the transit time τ_{2-3} of the above formula may be reduced considerably by proper electrode configuration.

Another high-frequency phenomenon which is particularly noticed in outer-grid-injection mixers is the high input conductance due to transit-time effects. The cause for this was first made evident when the change of signal-grid admittance with oscillator-grid potential was observed. Fig. 25.16 gives data on the susceptive and conductive components of the signal-grid admittance of this type of modulator as a function of oscillator-grid bias (no oscillator voltage applied). The data were taken at 31.5 megacycles and, as in the other input admittance curves, show the admittance components due to the presence of electrons only. It is seen that when the No. 3 grid is made sufficiently negative the input admittance is greatly increased. This

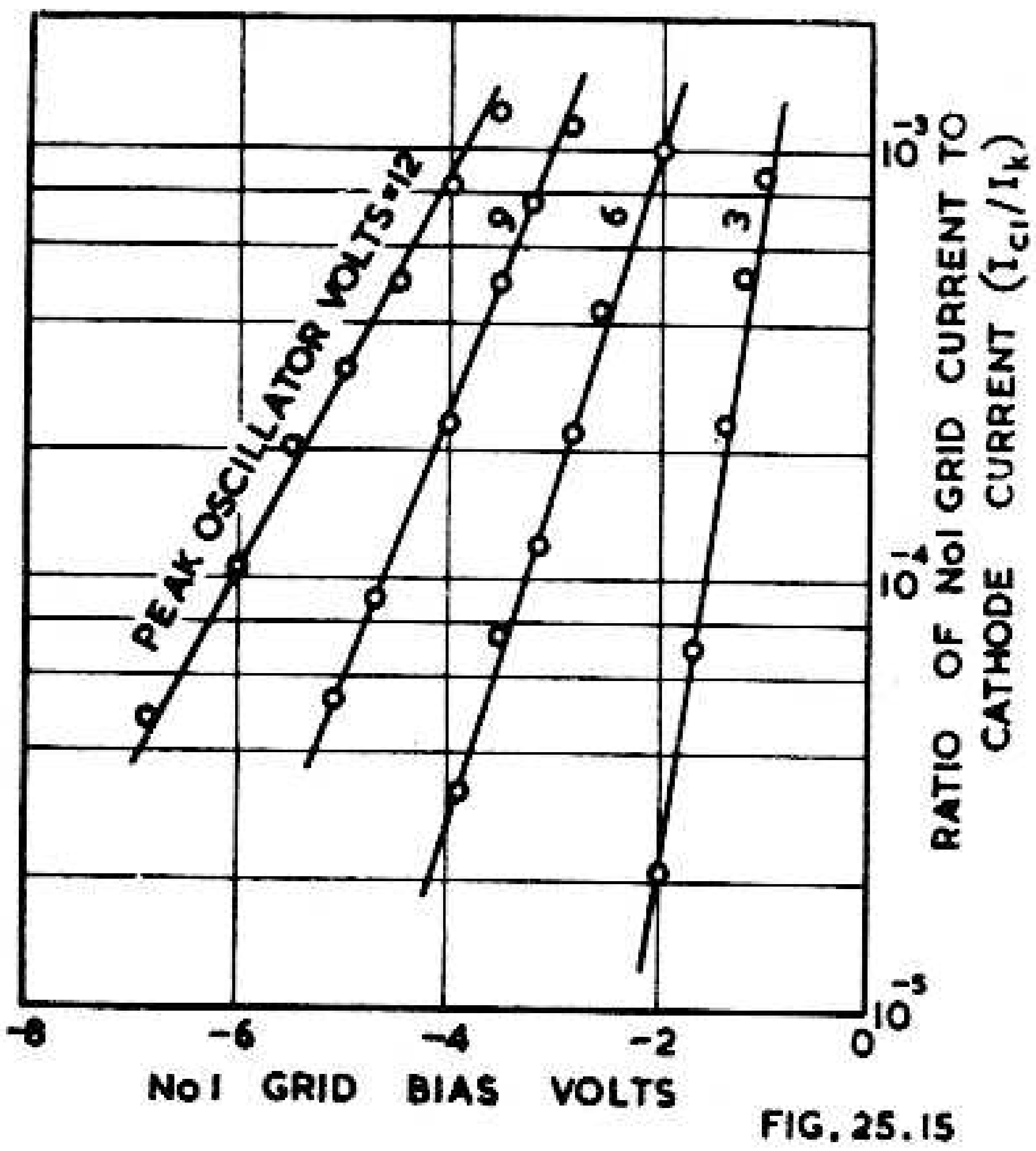


FIG. 25.15

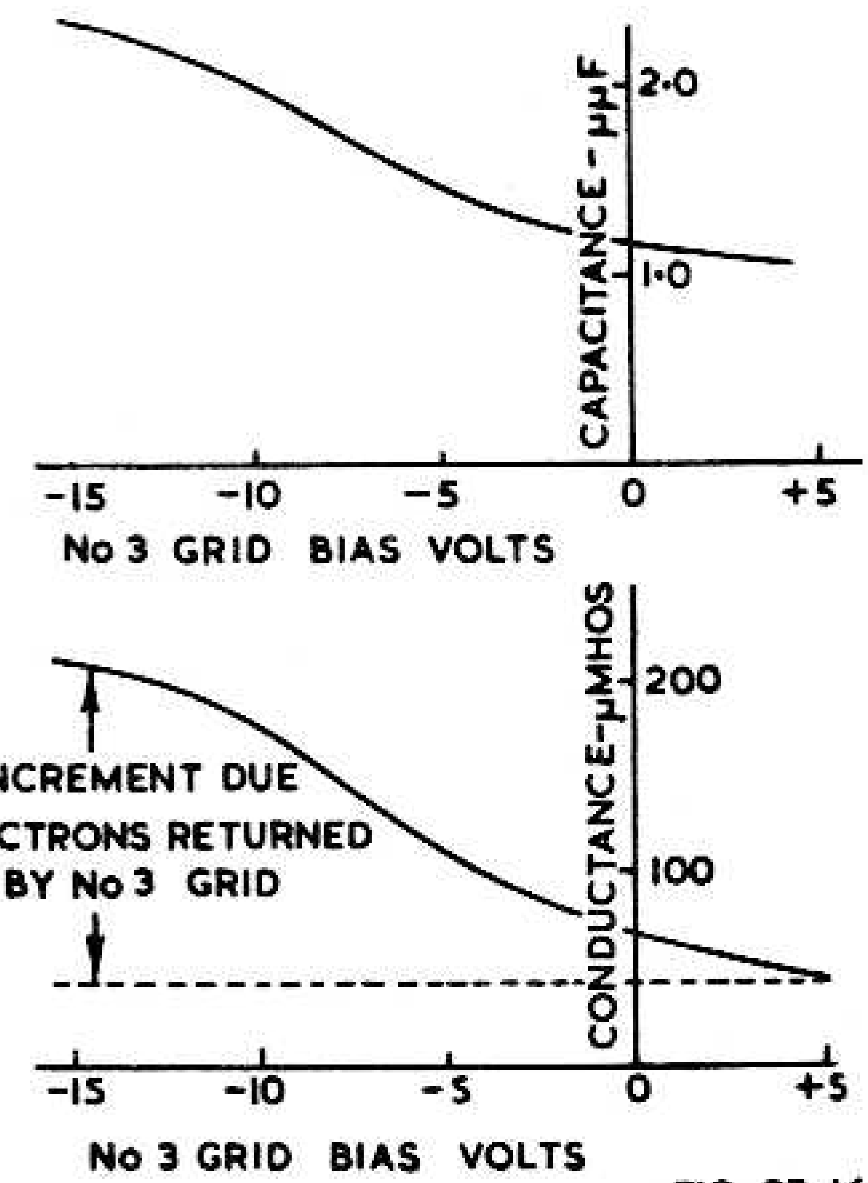


FIG. 25.16

Fig. 25.15. Signal-grid (grid No. 1) current in a typical mixer with a 20-megacycle oscillator voltage applied to grid No. 3. $E_{c_3} = -10$ *volts,* $E_{c_{2\ and\ 4}} = 100$ *volts,* $E_b = 250$ *volts.*

Fig. 25.16. Signal-grid (grid No. 1) admittance of typical mixer designed for outer-grid injection of oscillator. Data taken at 31.5 megacycles with no oscillator voltage applied. $E_{c_1} = -3$ *volts,* $E_{c_{2\ and\ 4}} = 100$ *volts,* $E_b = 250$ *volts.*

behaviour coincides, of course, with plate-current cut off. It seems clear that the electrons which are turned back at the No. 3 grid and which again reach the signal grid are the cause of the increased admittance. Calculations based on this explanation have been published by M. J. O. Strutt (Ref. 43) and show reasonable quantitative agreement with experiment. As in the other cases above, the upper curve of Fig. 25.16 is approximately independent of frequency while the lower one may be converted to any other frequency by multiplying the ordinates by the square of the frequency ratio.

When an oscillator voltage is applied, the No. 3 grid bias is periodically varied at oscillator frequency. The net input admittance is then the average value over the oscillator cycle. Such net values of the conductance component are shown in Fig. 25.17. The frequency for these curves is 31.5 megacycles. Values for other frequencies are obtained by multiplying the ordinates by the square of the frequency ratio. Curve *a* coincides with the fixed bias condition of curve *a* of Fig. 25.14 while curve *b* corresponds to the grid-leak-and-condenser bias as in *b* of Fig. 25.14. The conductance is approximately twice as high when the tube is used as a mixer as when it is used as an amplifier. This is a serious disadvantage, particularly at very high frequencies.

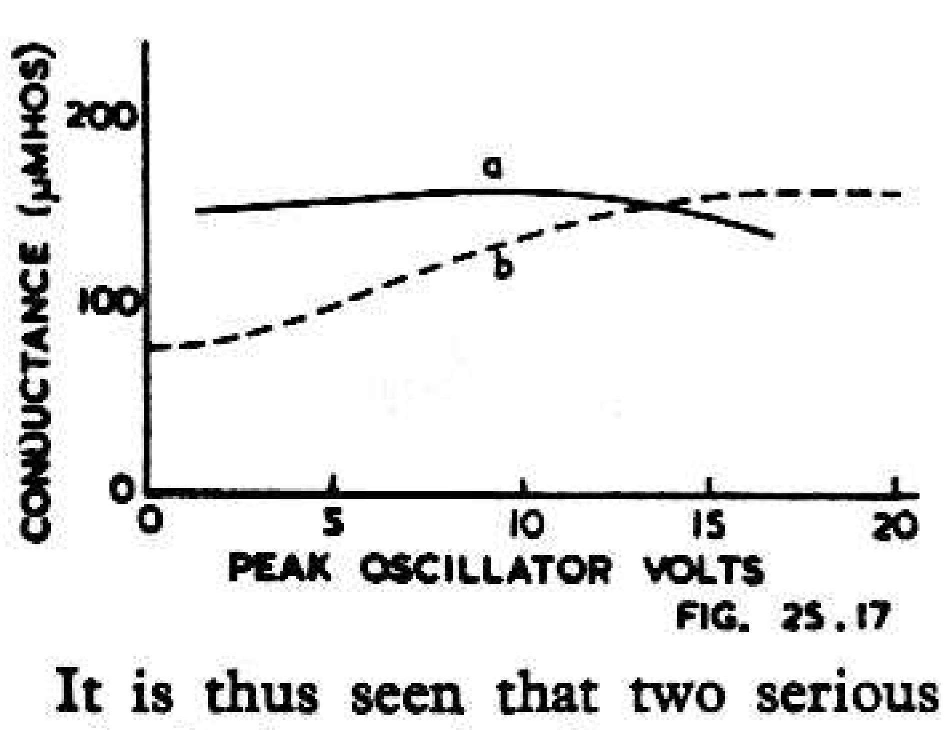


FIG. 25.17

Fig. 25.17. Signal-grid (grid No. 1) conductance of typical mixer designed for outer-grid injection of oscillator. Frequency, 31.5 megacycles, signal-grid bias, $E_{c_1} = -3$ *volts. Curve a corresponds to fixed No. 3 grid bias,* $E_{c_3} = -8$ *volts. Curve b corresponds to bias through a 50 000-ohm grid leak.*

It is thus seen that two serious disadvantages of the outer-grid-injection mixer are both due to the electrons returned by the oscillator grid which pass again to the signal-grid region. It was found possible to prevent this in a practical tube structure by causing the returning electrons to traverse a different path from the one which

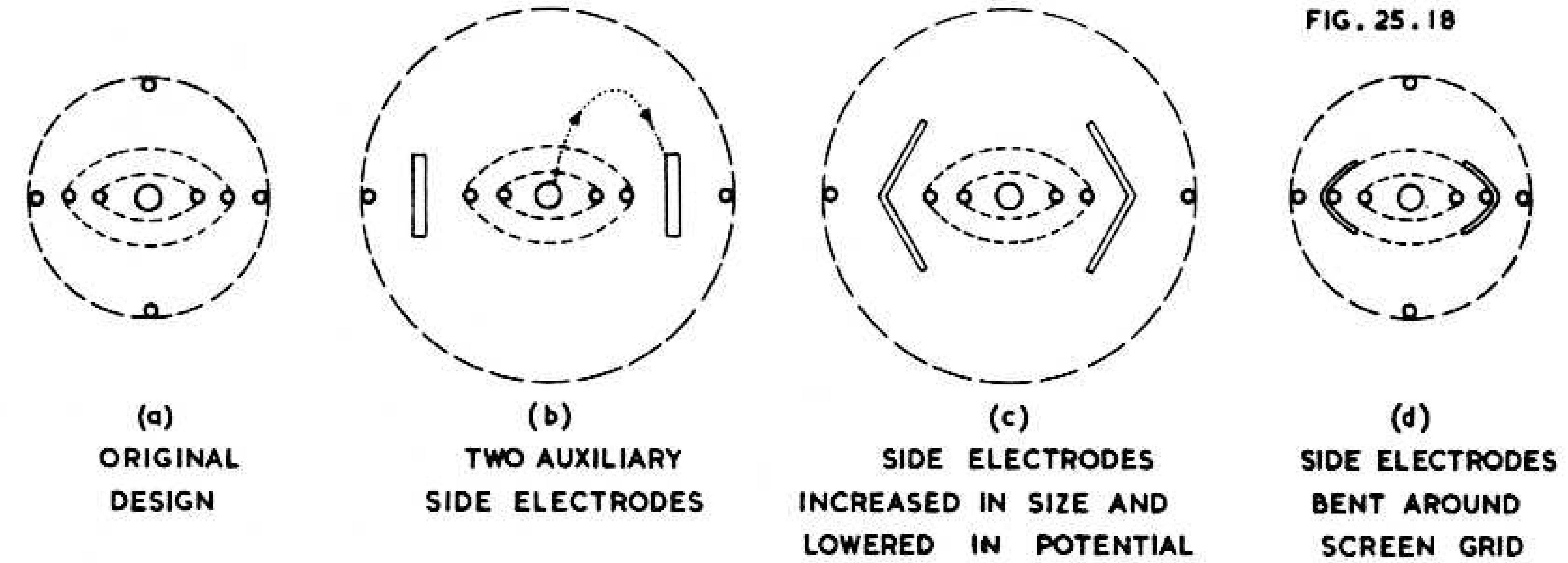


Fig. 25.18. Cross-sectional views of mixer designed for outer-grid-injection of oscillator. The views show only the portions of the tube inside of and including the oscillator injection grid.

they travelled in the forward direction*. (Ref. 45). The progressive steps towards an improvement of this kind are illustrated in Fig. 25.18 where cross-sectional views of the portion inside the oscillator grid of various developmental modulators are shown. The drawing (a) shows the original design, data on which have been given in Figs. 25.15, 25.16 and 25.17. Drawing (b) of Fig. 25.18 shows a tube in which two side electrodes operated at a high positive potential were added. In a tube of this kind many of the electrons returned by the No. 3 grid (oscillator grid) travel paths similar to the dotted one shown ; they are then collected by the auxiliary electrodes and thus do not re-enter the signal grid space. Tubes constructed similarly to (b) showed a considerable improvement in the signal-grid admittance increment due to returned electrons. Construction (c) shows the next step in which the side electrodes are increased in size and operated at somewhat lower potential. Because of the undesirability of an additional electrode and lead in the tube, the construction shown at (d) was tried. In this case the auxiliary electrodes are bent over and connected electrically and mechanically to the screen grid. Curves showing the progressive reduction in the signal-grid conductance increment due to returned electrons are shown in Fig. 25.19. The curves are labelled to correspond with the drawings of Fig. 25.18.

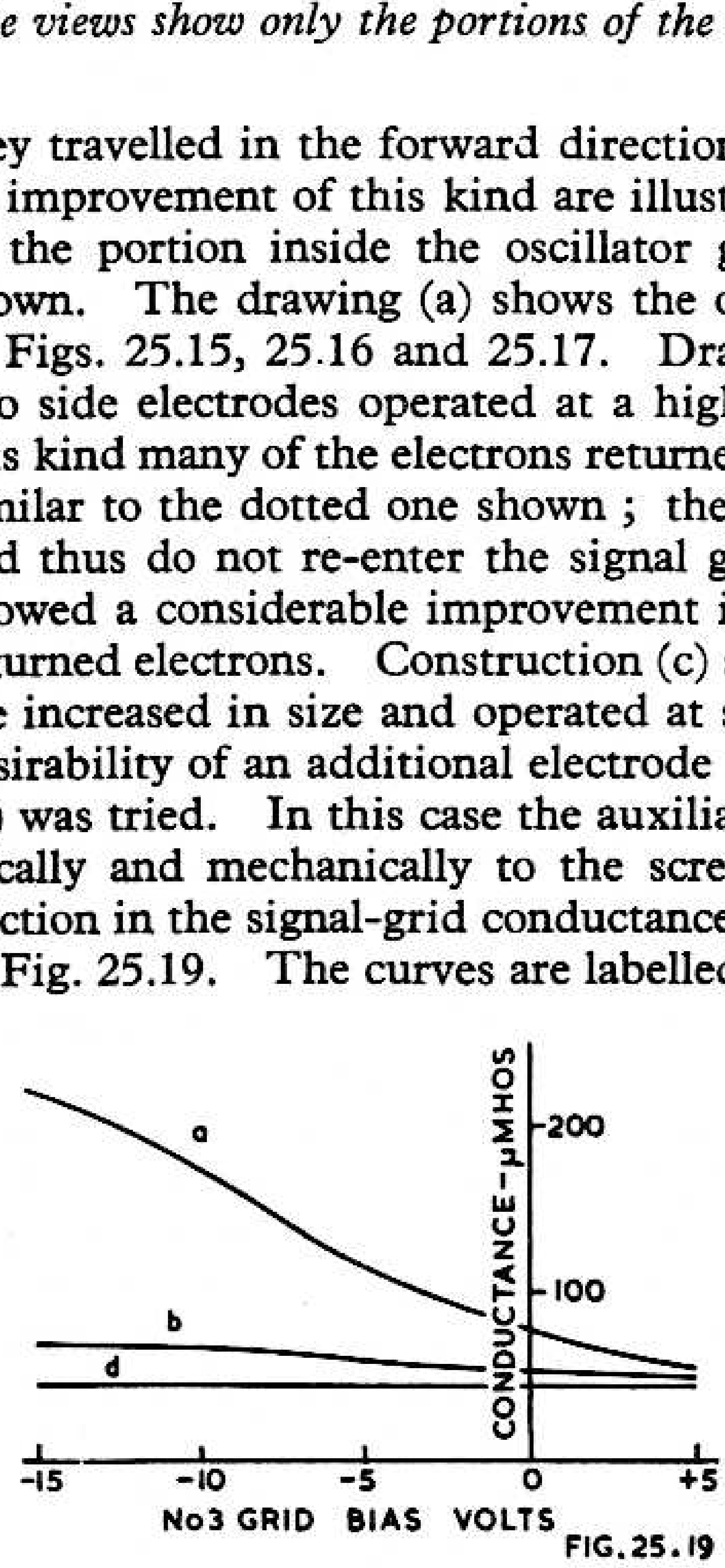


Fig. 25.19. Signal-grid (grid No. 1) conductance of the outer-grid-injection mixers shown in Fig. 25.18. Data taken at 31.5 megacycles with no oscillator voltage applied.

It should be noted that the use of the oscillator-grid support rods in the centre of the electron streams as shown in Fig. 25.18 (d) was found to improve the performance. No change in signal-grid conductance with oscillator-grid potential could be observed with this construction.† The conductance of the tube as a modulator, therefore, was reduced to less than half of that of construction (a). At the same time, a check of signal-grid current with a high-frequency oscillator applied to the No. 3 grid showed that this current was reduced to 1/20 of that of the original construction (a). The change in construction may be looked upon as dividing the constant A in the grid-current formula previously given, by a factor of more than 20.

*The same principles have now been applied to inner-grid-injection mixers and converters. See references 19 and 45.

†It should be mentioned that it is also possible to construct tubes in which the signal-grid conductance decreases somewhat with increasingly negative No. 3 grid bias. This effect is caused by the inductance of the inner screen-grid lead which causes a negative conductance in the input circuit when the inner screen current is high, as at negative No. 3 grid bias values. This negative conductance cancels part of the positive conductance of the signal grid.

FIG. 25.20

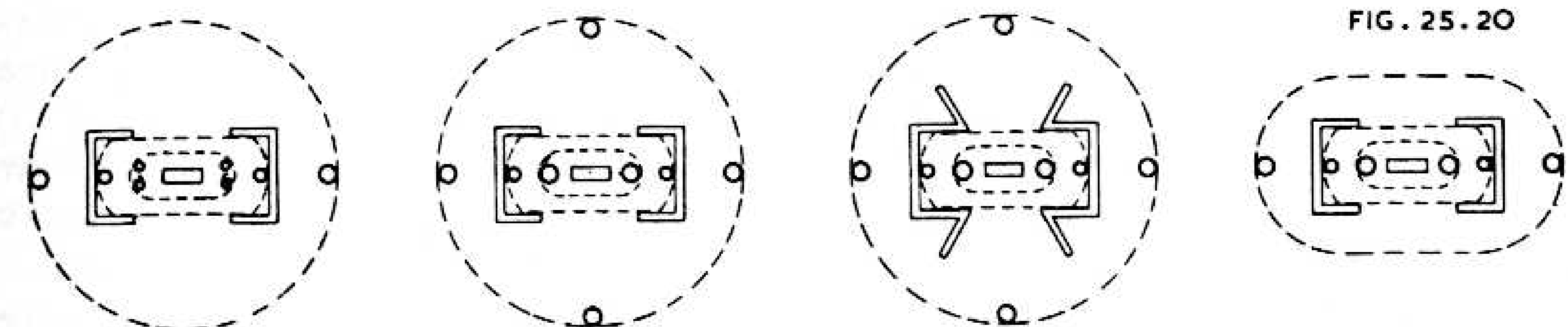


Fig. 25.20. *Cross-sectional views of improved mixers designed for outer-grid injection of oscillator. The views show only the portions of the tube inside of and including the oscillator injection grid.*

Another method of reducing the effect of electrons returned by the oscillator grid is to reduce the effect of electron transit time in the tube. This may be done by reducing the spacings, particularly the screen-grid-to-oscillator-grid spacing. This method of improving modulator performance has two disadvantages compared with the one discussed in connection with Fig. 25.18. The reduction in spacing is accompanied by a more sloping (i.e. less steep) signal-grid transconductance versus oscillator-grid voltage curve. This change in construction requires an increase in applied oscillator voltage to attain the same conversion transconductance. The second disadvantage is that such a method reduces the transit time and hence, the undesirable high-frequency effects only by an amount bearing some relation to the reduction in spacing. Since this reduction is limited in a given size of tube, the method whereby electron paths are changed is much more effective. The method of reducing spacing, on the other hand, is extremely simple to adopt. A combination of both methods may be most desirable from the point of view of best performance with least complexity in the tube structure.

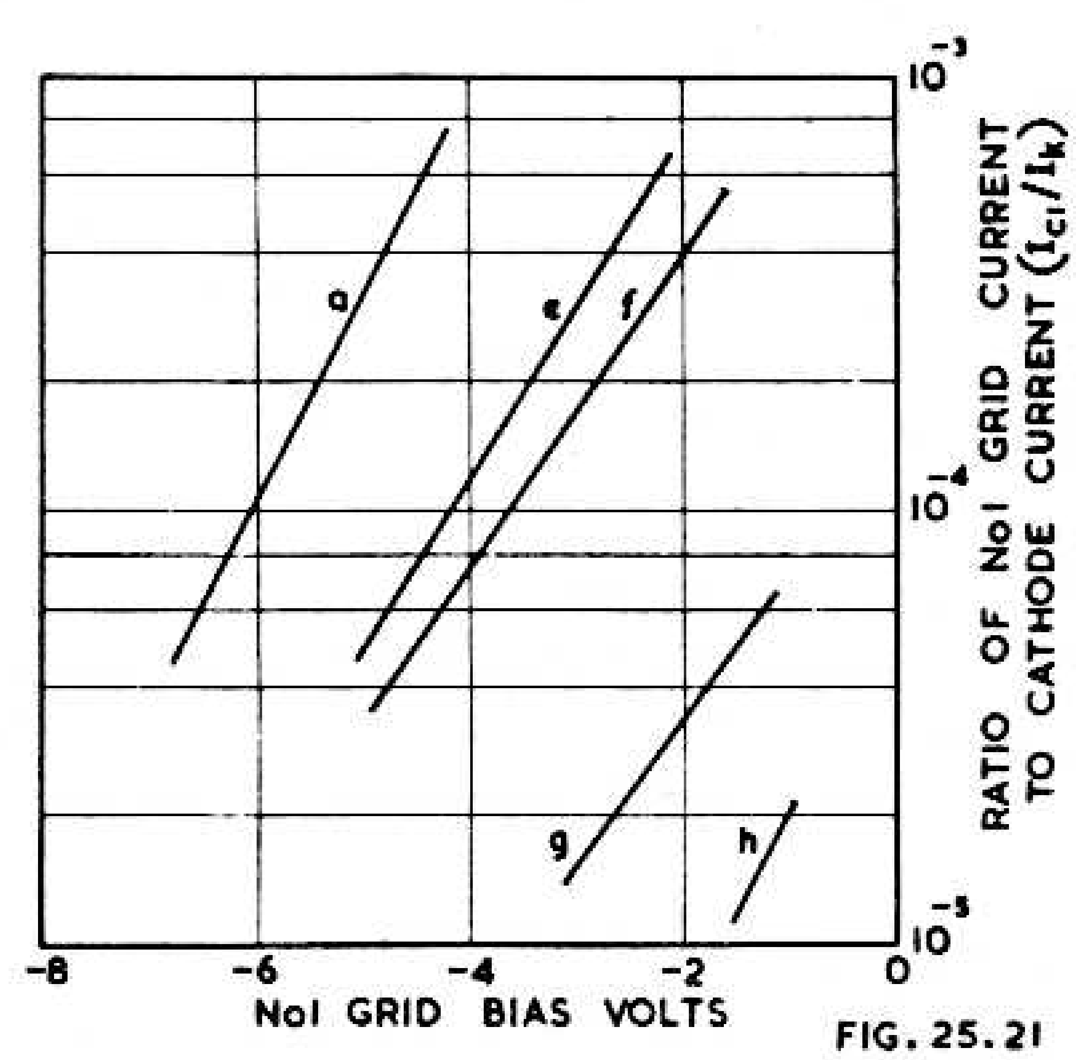


FIG. 25.21

Fig. 25.21. *Signal-grid (grid No.* 1) *current of the outer-grid injection mixers shown in Fig.* 25.20. *Curve a corresponds to the original design (a) of Fig.* 25.18 *and is shown for comparison. Data taken with a* 20-*megacycle oscillator voltage of* 12 *volts peak amplitude applied to the oscillator-grid.* $E_{c_3} = -10$ *volts,* $E_{c_{2\ and\ 4}} = 100$ *volts,* $E_b = 250$ *volts.*

In a mixer which must operate at high frequencies, it is not usually sufficient to eliminate the effects of returned electrons in order to assure adequate performance. For this reason the development of the principles shown in Fig. 25.18 was carried on simultaneously with a general programme of improving the tube. To this end, tubes were made with somewhat reduced spacing and with a rectangular cathode and a beam-forming signal grid (i.e., one with comparatively large supports). A number of developmental constructions are shown in Fig. 25.20. Construction (g), it will be noted, has finlike projections on the screen-grid channel members*. In construction (h) a reduction of spacing between screen and oscillator grids was combined with the channel construction. The relative performance of these constructions, so far as signal-grid current is concerned is shown in Fig. 25.21. The frequency used was 20 megacycles. The curve for the original design (taken from Fig. 25.15)

*This construction was devised by Miss Ruth J. Erichsen who was associated with the writer during part of the development work herein described.

is included and is drawn as *a*. All four of the constructions of Fig. 25.20 were satisfactory as regards signal-grid conductance; in every case the change in conductance as the oscillator grid was made negative was a negligible factor. Construction (h) required approximately 20 per cent more oscillator voltage than (e), (f), or (g) because of the reduction in slope of the transconductance versus No. 3 grid voltage curve which accompanied the reduced spacing between the screen and the No. 3 grid.

Outer-grid-injection mixers have the same or slightly greater signal-grid capacitance changes with automatic volume control as are found in amplifier tubes. In this respect they are inferior to inner-grid-injection converters or mixers. The use of a small unbypassed cathode resistance (Ref. 29, 30) is a help, however.

In closing this section, the subject of fluctuation noise will be considered. Experimental evidence indicates that the major portion of the noise in mixers with oscillator voltage on an outer grid is due to current-distribution fluctuations. (Ref. 24). The oscillator voltage changes the current distribution from plate to screen so that the mixer noise is given by the average of the distribution fluctuations over the oscillator cycle. In terms of the equivalent noise resistance the average has been found to be (Ref. 23),

$$R_{eq} = \frac{20\left[\overline{I_b} - \dfrac{\overline{I_b^2}}{I_a}\right]}{g_c^2}$$

where $\overline{I_b}$ is the average (i.e. the operating) plate current and $\overline{I_b^2}$ is the average of the square of the plate current over an oscillator cycle. I_a is the cathode current of the mixer section and is substantially constant over the oscillator cycle. This relation is not very useful in the form given. It is usually sufficiently accurate for most purposes to use an expression identical with that which applies to tubes with inner-grid oscillator injection, namely,

$$R_{eq} = \frac{20\overline{I_b}}{g_c^2} F^2,$$

where F^2 is about 0.5 for tubes with suppressor grids and somewhat higher for others. By assuming a typical tube characteristic, the noise resistance may be expressed in terms of the cathode current I_a of the mixer section and the maximum signal-grid-to-plate transconductance g_{max} as

$$R_{eq} = 120 \frac{I_a}{(g_{max})^2}$$

for operation at oscillator fundamental. For operation at second or third harmonic of the oscillator, the noise resistance will be approximately doubled, or tripled, respectively.

(v) Conclusion

It has been shown that the principle of frequency conversion in all types of tubes and with all methods of operation may be considered as the same (i.e. as a small-percentage amplitude modulation). The differences in other characteristics between various tubes and methods of operation are so marked, however, that each application must be considered as a separate problem. The type of tube and method of operation must be intelligently chosen to meet the most important needs of the application. In making such a choice, it is frequently of assistance to prepare a table comparing types of tubes and methods of operation on the basis of performance data. An attempt has been made to draw such a comparison in a qualitative way for general cases and for a few of the important characteristics. Table 2 is the result. It must be understood, of course, that the appraisals are largely a matter of opinion based on experience and the present state of knowledge. Furthermore, in particular circuits and with particular tubes, the relative standings may sometimes be quite different. A study of the fundamentals brought out in the previous sections of this paper should help in evaluating such exceptions.

TABLE 2

Approximate Comparative Appraisals of Methods of Frequency Conversion

Desirable Characteristic	Oscillator and Signal Voltages on No. 1 Grid		Oscillator Voltage on No. 3 Grid, Signal on No. 1 Grid		Oscillator Voltage on No. 1 Grid Signal on No. 3 Grid
	Triode	Pentode	Pentode	Hexode or Heptode	Hexode or Heptode
High conversion transconductance	Good	Good	Fair	Fair	Fair
High plate resistance	Poor	Good	Poor	Good	Good
High signal-to-noise ratio	Good	Good	Poor	Poor	Poor
Low oscillator-signal circuit interaction and radiation	Poor	Poor	Good	Good	Fair
Low input conductance at high frequencies	Poor[1]	Fair	Poor	Poor[2]	Good
Low signal-grid current at high frequencies	Good	Good	Poor	Poor[2]	Fair
Low cost of complete converter system	Good	Fair	Fair	Poor	Good

1. Due to feedback ; may be increased to Fair by proper circuit design.
2. May be increased to Fair by special constructions as described in text.

(vi) Appendix

Discussion of negative admittance of current-limited grids

In Figs. 25.10, 11 and 12 it was seen that the electronic signal-grid (i.e. input) admittance components (i.e. the admittance due to the presence of electrons) of a mixer designed for No. 1 grid injection of the oscillator are negative over a considerable portion of the normal operating range. Figs. 25.10 and 11, however, were taken with static voltages applied and so indicate that the phenomenon is not caused by an alternating oscillator voltage but is associated with the characteristics of the tube itself.

The input admittance of negative grids in vacuum tubes is the sum of three factors : (1) the "cold" admittance, or the admittance of the tube with the electron current cut off ; (2) the admittance due to feedback from other electrodes through tube and external capacitance, etc. ; and (3) the admittance due to the presence of the electrons in the tube. The first two factors have been well known for many years although certain aspects of the second have only recently received attention. (Refs. 29, 30 and 48). The third factor, however is not so well understood although the excellent work done during the last ten years has paved the way for a complete understanding of the subject. (Ref. 49). The present discussion is concerned only with this last point, namely the admittance of negative grids due to the presence of electrons in the tube.

Early work on transit-time effects in diodes and negative-grid triodes had indicated that, at very high frequencies, the conductance became negative in certain discrete bands (i.e., at large transit angles). It was not, at first, appreciated that conditions were possible with negative-grid triodes in which the input conductance could become negative even at low frequencies (i.e., at small transit angles). Data taken on the input (No. 4 grid) conductance of pentagrid converters by W. R. Ferris of this laboratory during 1934 showed that these tubes had a negative input conductance which varied as the square of the frequency and which remained negative at low frequencies. The conductance appeared, therefore, to behave in the same way as the positive input conductance of ordinary negative-grid tubes, except for a reversal in sign. The data on the pentagrid were taken with an external oscillator voltage applied to the No. 1 grid. The work of Bakker and de Vries (Ref. 50) disclosed the possibility of a negative input conductance at small transit angles in a triode operated under current-limited conditions. They gave an experimental confirmation for a triode operated at reduced filament temperature. Data taken by the writer during 1936 on a pentagrid converter showed that the negative conductance was present in this tube even when direct voltages, only, were applied and that it was accompanied by a reduction in capacitance. A fairly complete theory of the effect was developed in unpublished work by Bernard Salzberg, formerly of this laboratory, who extended the theory of Bakker and de Vries to the more general case of multigrid tubes with negative control grids in a current-limited region. Other experimental work was done on the effect during 1936 by J. M. Miller and during the first half of 1937 by the writer. In the meantime, the papers of H. Rothe (Ref. 51), I. Runge (Refs. 52, 53), and L. C. Peterson (Ref. 54) showed that independent experimental and theoretical work had been done on the negative-admittance effect in other laboratories.

In a rough way, the negative admittance found under current-limited conditions may be explained as follows : The electron current in a tube is equal to the product of the charge density and the electron velocity. If this current is held constant, a rise in effective potential of the control electrode raises the velocity and so lowers the charge density. A reduction in charge density with increase in potential, however, results in a reduction in capacitance, provided no electrons are caught by the grid. Thus, the susceptive component of the part of the admittance due to the current through the grid, is negative. Because of the time lag due to the finite time of transit of the electrons, there is an additional component of admittance lagging the negative susceptance by 90 degrees, i.e., a negative conductance. The value of the negative conductance will be proportional to both the transit angle and to the value of the susceptance. Since both of these quantities are proportional to frequency, the negative conductance is proportional to the square of the frequency.

The general shape of the curves of Fig. 25.10 may be explained as follows : At a No. 1 grid bias of about -20 volts, the cathode current is cut off and the electronic admittance is zero. At slightly less negative values of No. 1 grid bias, the electron current is too small to build up an appreciable space charge ahead of the signal grid (No. 4 grid). The latter grid, although it exhibits some control of the plate current does not control the major portion of the current reaching it and is thus in a substantially current-limited region. Its susceptance and conductance are, therefore, negative. Higher currents increase the negative admittance until, at some value of No. 1 grid bias, the electron current is increased to the point at which a virtual cathode is formed in front of some parts of the signal grid. At these parts, the current which reaches the grid is no longer independent of this grid potential, and as a result, a positive susceptance and conductance begin to counteract the negative admittance of other portions of the grid. The admittance curves reach a minimum and for still higher currents approach and attain a positive value. The current necessary to attain the minimum admittance point is less when the signal-grid bias is made more negative so that the minima for increasingly negative No. 4 grid-bias values occur at increasingly negative No. 1 grid-bias values.

It may be noted that the signal-grid-to-plate transconductance is at a maximum in the region just to the right of the admittance minima of Fig. 25.10 (compare Fig. 25.8). The admittance of such a tube used as an amplifier remains negative, therefore, at the maximum amplification point.

SECTION 2 : CONVERTER APPLICATIONS

By E. Watkinson, A.S.T.C., A.M.I.E. (Aust.), S.M.I.R.E. (Aust.).

(i) Broadcast frequencies (ii) Short waves (iii) Types of converters.

(i) Broadcast frequencies

A. Spurious responses

If a sinusoidal signal is applied to the signal grid of a converter and a sinusoidal oscillator voltage to the oscillator grid, harmonics of each signal appear in the mixer plate circuit, together with sum and difference frequencies between each of the applied voltages and their harmonics. There is a component of plate current at each of these spurious response frequencies and each could be selected by a suitably tuned circuit.

Alternatively when the plate circuit is fixed tuned to the intermediate frequency, undesired combinations of signal and oscillator harmonics can produce components of plate current at or near the intermediate frequency. These components heterodyne the desired difference frequency between signal and oscillator voltages.

The most important spurious response is at the image frequency of the desired signal, i.e. removed from the oscillator frequency by an amount equal to the intermediate frequency but on the side of the oscillator frequency remote from the desired signal. Such a signal mixes with the oscillator to produce the intermediate frequency in the same way as the desired signal, and interference from it is not dependent on the characteristics of the converter. With other responses, produced by a combination of signal and oscillator harmonics, the amount of interference is dependent on the operating conditions of the converter. Thus it is desirable to reduce the oscillator amplitude as far as possible without affecting sensitivity or low-voltage operation, with the object of decreasing the magnitude of the higher order components in the transconductance-time curve (B in Fig. 25.1) of the mixer.

Spurious responses are also possible at the harmonics of the intermediate frequency but these are due to feedback from the output of the i-f amplifier rather than to the converter.

For a discussion and chart of the various possible combinations of the signal, harmonics of the signal, the oscillator voltage and harmonics of the oscillator voltage see Ref. 78.

B. The signal-frequency circuit

1. Signal-grid loading

The reason for positive and negative loading effects produced respectively by outer-grid and inner-grid oscillator injection converters is explained in Sect. 1 of this chapter.

Converters with high capacitance between signal-grid and plate may also load the input circuit by feedback from the plate circuit, particularly at the low-frequency end of the broadcast band where the signal frequency approaches the intermediate frequency.

The input loading has resistive and reactive components, and the value of the resistive component (Ref. 78) is

$$R_g = \frac{(G_p + G_0)^2 + B_0{}^2}{g_m B_{gp} B_0}$$

where $G_p = 1/r_p$ = anode slope conductance,

G_0 = conductance of the external anode-load admittance at the signal frequency,

B_0 = susceptance of the external anode-load admittance at the signal frequency,

g_m = mutual conductance of the signal grid with oscillator operating,

and B_{gp} = susceptance of the grid-plate capacitance at the signal frequency.

Since the resonant frequency of the plate circuit is so far from the resonant frequency of the grid circuit when tuned to any point in the broadcast band, the expression can be simplified to

$$R_g = \frac{B_0}{g_m B_{gp}} \,.$$

The grid-circuit loading is thus directly proportional to the grid-plate capacitance of the converter, and it is only in types in which this is high that input loading becomes appreciable. For example the 6A8-G has a plate to signal grid capacitance of 0.26 $\mu\mu$F and with an i-f transformer tuned by 85 $\mu\mu$F to 455 Kc/s the resistive component of the loading due to feedback across the valve is about 0.25 megohm at 600 Kc/s rising to 0.45 megohm at 1000 Kc/s and 0.55 megohm at 1400 Kc/s.

The resistive component of the loading is also proportional to the reactance of the capacitor tuning the plate circuit of the converter. This capacitor should be given as large a value as possible if grid loading is the main consideration.

The reactive component of the input loading, which is always capacitive, appears in parallel with the tuning capacitor and can usually be ignored.

Some converters have much lower plate-grid capacitance, e.g. the 6J8-G has a maximum of 0.01 $\mu\mu$F, and in such cases the signal-grid circuit loading due to feedback from the plate circuit is negligible.

External coupling between oscillator and signal frequency circuits can cause another effect which may be mistaken for input circuit loading. This effect occurs in triode-hexode and similar types of converters when a tuned-plate oscillator circuit is used, the loss of gain being most noticeably at the high-frequency end of the broadcast band. In such a case, conversion sensitivity over the broadcast band is reasonably flat, but aerial (or r-f grid) sensitivity shows poor gain at the high-frequency end even although the coil, checked separately, gives constant gain over the band.

The coupling (capacitive or inductive) between the oscillator plate and the signal-grid causes an oscillator voltage to be applied to the signal grid which opposes the effects of the correctly injected oscillator voltage—see Sect. 1(iv)A of this chapter. In severe cases the coupling may even be sufficient to give an oscillator voltage on the signal grid in excess of the bias, leading to severe damping of the input circuit. Tuned-plate oscillator circuits are most likely to produce this effect because the required oscillator grid voltage is a constant whether plate tuning or grid tuning is used and with a tuned plate circuit this voltage must be developed across the smaller untuned grid winding of the oscillator coil. At the same time the oscillator plate voltage, instead of being appreciably smaller than the oscillator grid voltage as in a tuned-grid circuit, is appreciably greater.

A mechanical re-arrangement of the layout to isolate the oscillator-plate circuit from the signal-grid circuit will usually cure the trouble. In a difficult case it may be possible to arrange the wiring so that an opposite coupling between oscillator-grid and signal-grid will produce the required result.

2. **Signal-circuit regeneration**

It is possible to use controlled regeneration in a converter stage to give improved gain, image ratio and signal-to-noise ratio. Various possibilities are given in Ref. 65.

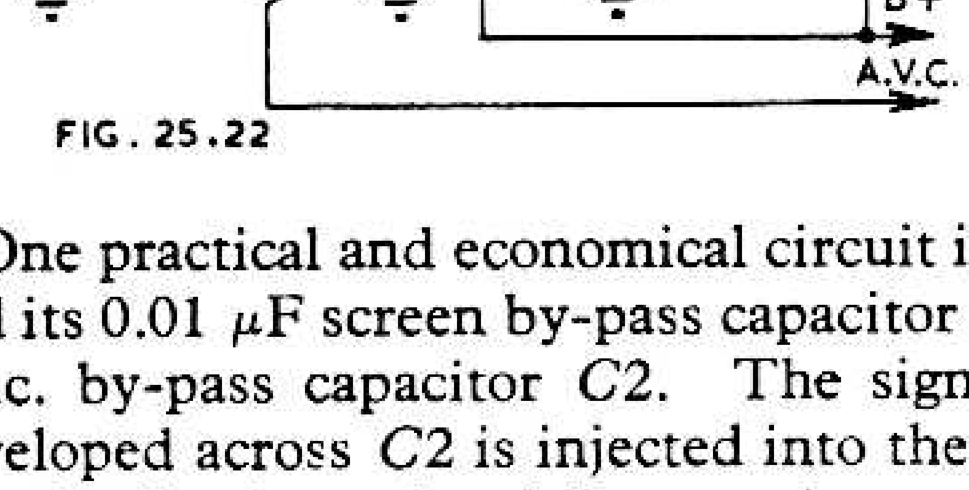


FIG. 25.22

Fig. 25.22. Screen regeneration in converter stage.

One practical and economical circuit is shown in Fig. 25.22. V1 is a triode-hexode and its 0.01 μF screen by-pass capacitor $C1$ is returned to ground through the 0.05 μF a.v.c. by-pass capacitor $C2$. The signal-frequency voltage from the screen circuit developed across $C2$ is injected into the grid circuit in the correct phase to give positive feedback at signal frequencies.

The regeneration obtained with this circuit is proportional to the reactance of $C2$ and is therefore inversely proportional to frequency. An improvement of about 4 db in sensitivity and image ratio and 2 db in signal-to-noise ratio can be obtained at the low-frequency end of the broadcast band with sufficient stability margin for production purposes. The regeneration is negligible at 1400 Kc/s.

In a small 3/4 valve receiver in which the aerial coil trimmer is connected from aerial to grid for maximum gain, sensitivity is greater at 1400 Kc/s than at 600 Kc/s, thus the circuit of Fig. 25.22 is useful in minimizing the sensitivity difference between the ends of the band.

Regeneration always has the disadvantage that variations in gain e.g. with different valves, are emphasized, but this need not be serious so long as moderate amounts of regeneration are used—sufficient for instance to give an improvement of the order outlined above. The circuit is applicable to other outer-grid injection converters, although component values may need modification.

C. Operating conditions

Recommended conditions for all types of converters are published by valve manufacturers and these are satisfactory for normal applications. However variations are often required, perhaps to use common voltage supplies for electrodes in the converter and other valves, or perhaps to obtain say maximum signal-to-noise ratio even at the expense of sensitivity.

The usual design procedure in such cases is to supply screen, oscillator plate and bias voltages by means of variable resistor boxes and to control the oscillator amplitude by means of resistors shunted across the untuned primary. In some cases, to be mentioned later, the effect of varying the primary impedance may need investigation.

The valve operating conditions are set approximately to those recommended and then each voltage in turn is adjusted for maximum performance. This may need to be done more than once as variations in, say, bias voltage will affect the required screen voltage.

In a typical case, the requirement from a converter might be maximum sensitivity and as the bias voltage is reduced the gain might rise to a maximum and then fall. This is due to an increase in conversion transconductance as the bias is reduced, followed by a decrease in plate resistance which more than offsets the increasing transconductance. However, with the bias voltage set for maximum sensitivity the screen dissipation may be excessive. A decrease in screen voltage followed by a further decrease in bias may give a similar sensitivity with satisfactory screen dissipation. It is important that the bias should not be reduced to a point at which some valves may draw grid current. The contact potential of indirectly-heated types normally does not exceed -1.0 volt and the applied bias in such cases can be safely reduced to between -1.5 to -2.0 volts provided that electrode ratings are not exceeded. In the case of battery types, the contact potential is normally zero or positive, so that it is not necessary to apply a negative bias voltage to avoid grid damping.

Increased oscillator amplitude is likely to improve sensitivity to a broad maximum followed by a slight fall. However, as spurious responses increase rapidly with oscillator drive, the amplitude should not be any greater than necessary. It is advisable to carry out checks at least at the limits of the tuning range to be covered.

To obtain maximum signal-to-noise ratio a similar procedure would be adopted but the final operating conditions would probably differ, e.g. in the case of a triode-hexode the screen voltage might be reduced. On the other hand, if the converter noise is of the same order as the signal-grid circuit noise, as is possible on the broadcast band, the effect of variations in converter conditions on the signal-to-noise ratio of the receiver would not be great.

When operating voltages have been determined, resistor boxes may still be useful for obtaining a suitable screen voltage-divider circuit if a.v.c. design is at all critical. If a minimum amount of control is required on the converter, for example to avoid a rapid increase of noise as a.v.c. is applied, a single series resistor from B+ will allow the screen voltage to rise in a triode-hexode type of converter and so decrease the amount of control for a given a.v.c. voltage. Such a circuit will also minimize cross-modulation. However in converters without a suppressor grid the screen voltage must not approach the plate voltage because secondary emission from the plate will cause a large reduction in plate resistance and lead to damping of the first i-f transformer, so that some bleed from screen to ground may be needed. The best compromise can readily be determined with resistor boxes.

If maximum a.v.c. control is required from a triode-hexode converter, for example when it is followed by a reflexed amplifier, and play-through must be reduced to a minimum, a large screen bleed may be needed.

Maximum a.v.c. control can be obtained in pentagrid converters by reducing the oscillator plate voltage to a minimum consistent with satisfactory oscillator performance. For example with 20 volts bias on the signal grid, the conversion transconductance of a 6A8-G is reduced to less than one half of the original figure when the oscillator plate voltage is changed from 200 to 100 volts, although at minimum signal-grid bias the oscillator plate voltage has little effect on the conversion transconductance.

Final operating conditions should not be decided upon until the tests outlined above have been repeated with a number of valves, some of which should preferably have characteristics near the upper and lower acceptance limits for the type.

(ii) Short waves

A. Alignment

Short wave alignment of the converter stage is complicated by two factors, firstly by " pulling " of the oscillator frequency due to adjustments to the signal-grid circuit, and secondly by the fact that in many cases the range of the oscillator circuit trimmer is sufficient for it to be adjusted either to the correct response or to the image response.

1. **Pulling** : Pulling is due to coupling, either in the converter valve or externally, between the signal and oscillator circuits, and unless suitable precautions are taken it can result in faulty alignment at the high-frequency trimming point on the short-wave band. The effect is that as the signal-frequency trimmer is adjusted towards its correct setting, thus increasing the output, the oscillator is simultaneously detuned, which decreases the output when the signal is detuned from the peak of the i-f response. Thus the output rises to a maximum and then falls again, as is normal, but the maximum, instead of being the correct setting, is the point at which the detuning of the oscillator causes the gain to fall more rapidly than the increase due to the approach towards resonance of the signal-grid trimmer.

The true maximum can be found by alternately peaking the signal circuit and retuning the oscillator circuit to resonance (by means of the main tuning control on the receiver) or by carrying out the two operations simultaneously—i.e. by slowly moving the signal-grid trimmer and continually tuning through the signal, noting the amount of output on each occasion and continuing until a maximum is reached. This procedure is known as " rocking " the tuning control.

If there is a noise source available with constant output over a band of frequencies at the alignment point (e.g. valve noise in a sensitive receiver, or a noise diode) it is possible to trim the signal-frequency circuit without rocking as the wide-band signal source avoids the effects of detuning. The signal input should not be large enough to operate the a.v.c.

2. **Images** : Some of the difficulty experienced in short wave alignment at the end of the band is due to confusion between the correct response and the image. This results from the fact that the correct signal appears to differ depending on whether the signal generator is tuned or the receiver is tuned. For example, if a receiver with a 450 Kc/s intermediate frequency and with its oscillator on the high-frequency side of the signal frequency is correctly tuned to an 18 Mc/s signal the receiver oscillator is set to 18.45 Mc/s. The generator can then be tuned to a frequency 450 Kc/s **higher** than the oscillator i.e. 18.9 Mc/s and the image response will be found, or alternatively with the generator still set to 18 Mc/s the receiver can be tuned to the **lower** frequency of 17.1 Mc/s when the local oscillator will be at 17.55 Mc/s, 450 Kc/s from the signal, and an image will be heard again.

Whenever there is any doubt as to which is the correct response, the frequencies should be worked out—conditions will be reversed from the example above if the oscillator is at a lower frequency than the signal—but a simple rule is that if the receiver is being tuned and the oscillator is on the **high** frequency side of the signal, then the **higher** frequency signal is the correct one.

A similar effect to an image response can also occur during signal-frequency trimmer adjustment. With the oscillator on the high-frequency side of the signal it may be found that as the signal-frequency trimming capacitance is increased, the output, with rocking, rises to a maximum, falls and rises to a second maximum, which is the correct one. On the other hand if the receiver is left aligned at the first maximum it is found that the adjustment is approximately correct for the image frequency.

This effect has no direct connection with the image response, but is due to the decrease in sensitivity which can occur when the signal-grid circuit is tuned to the oscillator frequency—owing to grid damping in a severe case, or out-of-phase oscillator injection in other cases. Thus as the signal-circuit tuning initially approaches the correct setting the gain rises, decreases again when the separation from the oscillator frequency is small, and then rises to the correct peak. In a sensitive receiver the illusion of correct alignment on the wrong peak is heightened by an increase in the level of the background noise due to the increased receiver sensitivity near the image frequency.

B. Operating conditions

Converters can be set up for operation on the short-wave band in a manner similar to that described for the broadcast band. However additional steps are also worthwhile. Particularly when harmonic mixing—see C(4) below—is being used, it is advisable to investigate the effects of different values of oscillator-grid resistor. To avoid varying two quantities at the same time the oscillator voltage should be kept as constant as possible while the value of the grid leak is altered, and readings should be taken at least at each end of the short-wave band. After the optimum grid leak value has been chosen, previously selected operating potentials should be rechecked.

When the converter is a type with inner-grid oscillator voltage injection, improved sensitivity may be obtained with the oscillator operating at a lower frequency than the signal instead of under the more usual higher frequency condition.

In Sect. 1 of this chapter it was explained that the space charge in front of the signal grid develops a voltage at oscillator frequency on the signal grid and that this voltage, depending upon its phase, increases or decreases the oscillator-frequency modulation of the electron stream. In a particular case with the oscillator at a higher frequency than the signal, the tuned circuit connected to the signal grid would present capacitive reactance to the oscillator voltage. A positive increment of oscillator-grid voltage increases the current flow and thus increases the space charge in front of the signal grid, i.e. the space charge becomes more negative. Since there is capacitive reactance between the space charge and signal grid and between signal grid and ground, a negative increment of space charge potential results in a (smaller) negative increment of signal-grid potential. Thus a positive change in potential on the oscillator grid results in a negative change in potential on the signal grid, so that the effective oscillator modulation is reduced with a consequent reduction in sensitivity.

However when the oscillator is operated at a lower frequency than that of the signal, the signal-grid circuit presents an inductive reactance to the oscillator voltage and the phase of the oscillator voltage on the signal grid is reversed. Thus the effective modulation of the electron stream is increased and sensitivity is improved.

In practice, if neutralizing is not used, an improvement in sensitivity of two or three times is sometimes obtainable by changing from high-frequency to low-frequency oscillator operation, with an increase of the image ratio in the same proportion. The improvement in sensitivity and image ratio decreases towards the low-frequency end of the short-wave band as the separation between signal and oscillator circuits (i.e the intermediate frequency) becomes a larger fraction of the oscillator frequency

For correct tracking with the oscillator on the low-frequency side of the signal, the padder is moved from the oscillator to the signal-frequency circuit(s) but this does not introduce serious complications. However, a large band coverage is more difficult with the oscillator on the low side because the oscillator circuit covers a greater frequency ratio than the signal circuit, whereas on the high side it covers a smaller ratio.

It is possible for external coupling between oscillator and signal circuits to exceed that which occurs due to internal coupling in the converter itself and it is advisable to check that the external coupling is negligible before investigating the effects of low-side and high-side oscillator operation. A valve voltmeter can be used across the circuit and the variation of the indication as the signal circuit is tuned through the oscillator frequency, with the circuit connected to and disconnected from the signal grid, shows the relative amount of coupling due to internal and external sources.

A further desirable step in the investigation of short-wave performance is the determination of the best sensitivity that can be obtained by adjustment of the magnitude and phase of the oscillator voltage on the signal grid. This can be done by using a very small capacitor, variable in order to adjust the amplitude of the oscillator voltage, and connected between either signal-grid and oscillator grid or signal-grid and oscillator plate to vary the phase of the oscillator voltage. In the absence of appreciable coupling within the receiver, some increase in sensitivity is usually obtained with a particular value of neutralizing capacitor, because maximum sensitivity is obtained not when the oscillator voltage on the signal grid is a minimum but when it is the correctly-phased maximum that can be present without signal-grid current flowing.

Thus, when using a plate tuned triode-hexode converter, it may be found experimentally that sensitivity is improved at 18 Mc/s with a 0.5 $\mu\mu$F capacitor connected between oscillator grid and signal grid. In such a case it is usually possible to obtain a capacitance of approximately the correct value without using a separate component by a suitable arrangement of the wave-change switch wiring.

However, when neutralizing is used, a measurement should be made of the effect of the neutralizing capacitance on the frequency stability of the receiver at 18 Mc/s. In general, sensitivity improvements due to oscillator voltage on the signal-grid must be offset against decreased frequency stability. The frequency shift in pentagrid converters of the 6SA7 type increases so rapidly with increased capacitance between signal-grid and oscillator grids that a neutralizing capacitance cannot be used in a normal short-wave receiver if a.v.c. is applied to the control grid.

Unless special precautions are taken, grid current variations over the short-wave band are greater than desired. The increase at the high-frequency end of the band occurs because, although the coupling between primary and secondary of the oscillator coil which is required to maintain oscillation is reduced, the actual coupling remains unchanged. A small resistor (say 25 ohm $\frac{1}{4}$ watt) wired between the tuned circuit and the oscillator grid is the usual method of obtaining reasonably uniform grid current over the band. The effective parallel circuit damping of the resistor in series with the capacitance of the oscillator grid is inversely proportional to the square of the frequency so that a large reduction in grid current can be obtained at the high-frequency end of the band without noticeably affecting the grid current at the low-frequency end.

Such a resistor is sometimes also required to prevent squegging of the local oscillator when the amplitude of oscillation is high and the time constant of the oscillator grid circuit is much greater than the time required for one cycle of the oscillator frequency. For further information on squegging see Chapter 35 Sect. 3(vi)C and Ref. 73.

C. Frequency stability

All the problems of frequency stability experienced with a separate oscillator (see Chapter 24) occur when the oscillator section of a converter is used as the local oscillator in a superheterodyne receiver and in addition there are further causes of instability due to interaction between the oscillator and mixer sections of the valve.

1. Frequency variations due to the local oscillator

Local oscillators are always self-biased triodes so that the only potential applied to the circuit from external sources is the plate-supply voltage. Two common types of frequency instability are due to this link with the rest of the receiver. The first is " flutter " due to fluctuations in the B supply voltage as a result of the output-valve current varying at audio frequencies. These fluctuations vary the oscillator frequency so that the signal is continually detuned, giving a-f signals which make the " flutter " self-sustaining. The mechanism of " flutter " is more fully described in Chapter 35 Sect. 3(vi)B and methods of overcoming it are given.

The second detuning effect for which the B supply is responsible is due to the application of a.v.c. to the controlled stages of a receiver. This reduces the B current, causes the B voltage to rise, and thus produces detuning which will vary, for instance, with the instantaneous level of a fading station.

Although in a particular case the detuning may not be sufficient to produce the symptoms described, it may still give unpleasant tuning at the high-frequency end of the short-wave band, a typical effect being that the receiver does not tune smoothly but jumps from one side of a signal to the other.

When testing for these effects, modulated and unmodulated signals of magnitudes varying from maximum to minimum should be used at the highest tuning frequency of the receiver and the volume control setting should also be varied. If equipment for measuring frequency shift is not available a useful test is to tune the receiver as accurately as possible to a large signal (perhaps 0.1 volt) with the volume control suitably retarded and then reduce the input to the smallest usable signal and turn up the volume control. If the signal is still tuned, frequency shift may be considered satisfactory. The test should be repeated with different signal inputs in case positive and negative frequency shifts should cancel over the range of inputs first selected

If flutter occurs on a large unmodulated signal there are circumstances under which it can be ignored. When the modulation is switched on it may be found that the output valve is severely overloaded, and that when the volume control is turned down to give no more than full a-f output the flutter may not occur. In such a case the flutter would not be noticed in normal use of the receiver and if a cure cannot be effected cheaply it may not be warranted.

In practice it is found that frequency stability can often be improved by the simple expedient of increasing the oscillator grid current. In one particular triode-hexode the 18 Mc/s frequency shift for a given change in control grid bias was reduced to one tenth by increasing the oscillator grid current from 200 to 600 μA. Frequency shift due to other causes, and with other types of converters can be improved in the same way, although in varying degrees. The improvement may of course have to be offset against a decrease in sensitivity and an increase in noise.

A third type of oscillator frequency instability due to the B supply is experienced in F-M receivers. If there is a hum voltage superimposed on the B supply to the oscillator plate this will produce frequency modulation of the local oscillator, and the F-M detector will convert the resultant hum modulation of the signal carrier into a-f hum.

2. Frequency variations due to the mixer

The space charge adjacent to oscillator electrodes in a converter valve can be varied by altering the potentials on mixer electrodes. This, of course, gives rise to oscillator-frequency variations. The mixer screen voltage is particularly important because one of the two grids comprising the screen is adjacent to an oscillator electrode, so that it is desirable to keep this voltage as constant as possible. This frequently necessitates a separate decoupling resistor and capacitor for the converter screen because with a common supply to converter and i-f amplifier screens the application of a.v.c to the i-f amplifier (even if the converter stage is not controlled) alters the mixer screen voltage and thus causes frequency shift. On rare occasions this shift may be used to offset another in the opposite direction due to a different effect.

When frequency shift is troublesome it is desirable to isolate each of the causes and determine its direction, as cancellation is sometimes possible even if a complete cure is not.

The effects upon oscillator frequency of varying a.v.c. voltages applied to a signal-grid electrode are more complex. With either outer-grid or inner-grid oscillator injection a space charge adjacent to the oscillator grid can be varied by altering the direct potential applied to the signal grid. That this is not always the main source of frequency drift can be seen in some receivers with this type of instability by measuring the frequency shift for a given change in converter a.v.c. voltage with the signal generator connected firstly to the aerial terminal and then to the converter grid. It may be found that the frequency shift in the former case is many times that with the signal generator connected to the converter-grid, and it may even be in the opposite direction.

This type of instability is due to coupling between oscillator and control-grid circuits. In Sect. 1(iv)B of this chapter it is explained that at high frequencies it is usual for oscillator voltage to appear at the signal-grid of the converter. Coupling between signal- and oscillator-grid circuits results in some of this voltage being returned to the oscillator circuit and, in a manner similar to the operation of a reactance-tube modulator, this results in a modification of the effective capacitance at the oscillator grid. A constant variation of this capacitance would be of no significance, but any alteration of converter operating conditions which resulted in a variation of oscillator voltage on the signal grid, e.g. the application of a.v.c. voltages, would alter the effective reflected capacitance at the oscillator grid and thus cause oscillator frequency shift. (Ref. 87).

With a given set of conditions the simplest method of curing this type of frequency shift is usually to minimize the coupling between oscillator and signal grid circuits. In general, coupling is capacitive, partly within the valve and partly external to it. In most modern valves the internal capacitance is not troublesome and to effect a cure it may be necessary to separate short-wave aerial and oscillator coils from each other, to separate the wiring from the two grids or from the two sections of the gang condenser, and to use wave-change switch contacts on the opposite side of the wafer. In the particular case of the 6BE6, it is desirable to reduce the external capacitance between the two grids of the valve to 1 $\mu\mu$F or less.

Another possible cure for this type of frequency shift is the use of harmonic mixing, in which the frequency difference between signal and oscillator circuits is so great that no significant amount of oscillator frequency voltage appears on the signal grid.

3. Tuned-plate operation

An analysis of tuned-grid and tuned-plate oscillators (Chapter 24 Sect. 2) shows the differences between the two types of operation. However, it is mainly because of the structure of some types of converters that tuned-plate operation is preferable. When the oscillator section of a converter has a common electron stream with the mixer section, the oscillator-plate electrode is designed to modulate the electron stream as little as possible since modulation by the oscillator plate cancels that from the oscillator grid. It is also desirable in equipment design to keep the oscillator plate alternating voltage as small as possible. With a tuned-grid oscillator, the plate voltage is smaller than the grid voltage and the reverse is the case with a tuned-plate oscillator. Accordingly, tuned-grid oscillators are always used for converters having a common electron stream for the oscillator and mixer sections.

Triode-hexode and triode-heptode converters have a separate triode for use as a local oscillator, and oscillator injection into the mixer section is carried out by a special injector grid which is in most cases internally connected to the oscillator grid. The capacitance and coupling effects of the injector grid are thus introduced into the oscillator circuit, and if a tuned-grid oscillator is used they appear directly across the tuned circuit.

However with a tuned plate oscillator these effects are coupled to the tuned circuit from the oscillator primary winding and their effect is thus minimized. An improvement in oscillator frequency stability, with respect to the effect of a.v.c. application to the signal grid, of five times has been measured with a typical triode-hexode on converting from a tuned-grid to a tuned-plate oscillator, while frequency shift for a given change in B supply voltage was halved under the same conditions.

4. Harmonic mixing

Many of the difficulties experienced in high-frequency applications of converters are due to the small percentage of frequency difference between the oscillator and signal-frequency circuits. It was shown in Sect. 1 of this chapter that a sinusoidal oscillator voltage applied to the injector grid of the mixer section can also give mixing at harmonics of the oscillator frequency. By taking advantage of this inherent converter characteristic it is possible to separate widely the resonant frequencies of the oscillator and signal frequency tuned circuits and thus eliminate many difficulties in high-frequency converter applications.

There are disadvantages to harmonic mixing but in some cases they are outweighed by the advantages to be obtained. Firstly there is some loss in sensitivity over a part, at least, of the tuning range and secondly there are spurious responses due to incoming signals mixing with the oscillator fundamental.

With careful design the loss in sensitivity is small. In the normal 6 to 18 Mc/s short-wave range it will probably not be greater and may be less than 2 db from 10 to 18 Mc/s. The maximum loss is usually at 6 Mc/s, and should not exceed 3 db, about 2 db or less being normal. Even when the converter is the first valve in the receiver no serious decrease in signal-to-noise ratio of the complete receiver should be experienced.

Because o. a number of secondary effects which occur with harmonic mixing, this performance is better than would be expected from an investigation of maximum second harmonic conversion transconductance alone, although it presupposes that the best conditions for harmonic mixing are used. It will be found that higher grid current is necessary, particularly at the low-frequency end of the band, and a lower value of grid leak than normal is often useful in obtaining maximum sensitivity. In general, it is desirable to investigate the converter conditions carefully, as described previously, and for satisfactory performance inner-grid mixers should operate with the second harmonic of the oscillator on the low-frequency side of the signal.

The reason for this is as follows. The modulation of the electron stream by grid 1 gives a transconductance-time curve similar to Fig. 25.1B, and the space charge in front of the signal grid fluctuates at the same rate. This fluctuation, although of the same frequency as the local oscillator, has large harmonic components and since the space charge is coupled capacitively to the signal-grid circuit and this is tuned approximately to the second harmonic frequency, most of the voltage induced into the signal-grid circuit is at the frequency of the second harmonic of the oscillator. This second harmonic voltage remodulates the electron stream with a phase which depends on whether the signal frequency circuit has capacitive or inductive reactance at the second harmonic frequency so that the re-modulation aids or opposes, depending on its phase, the original second-harmonic modulation by grid 1. For inner-grid injection converters, the oscillator second-harmonic frequency should be lower than the signal frequency for aiding re-modulation.

It will be found that with harmonic mixing the same padder can be used as that required for the same frequency coverage with fundamental mixing.

Although the spurious responses which arise due to fundamental mixing are a disadvantage of harmonic operation, a tuned r-f stage ahead of the converter will reduce them to negligible proportions, not greater than -60 db in a typical case. Moreover, even when fundamental mixing is used there are responses, although smaller ones, due to signals mixing with the oscillator harmonics. Fundamental spurious responses 30 db below the signal could be taken as representative of receivers, without a r-f stage, using second-harmonic mixing so that these responses cause appreciably less interference than the normal image.

The benefits to be obtained from harmonic mixing are evident when operating, for example, the 1R5 converter on the 6 to 18 Mc/s short-wave band. Because the 1R5 has no separate oscillator-plate electrode, an oscillator circuit is often used in which the screen is connected to the oscillator primary winding to operate as the oscillator plate. A neutralizing capacitor is then needed between signal grid and oscillator grid and the determining of a suitable capacitance and its tolerances, the minimizing of coupling between signal and oscillator circuits and the stabilizing of the oscillator with reference to supply-voltage variations involve a considerable amount of work, as does the duplicating of the results in production receivers.

However, with harmonic mixing neutralizing is not required. This is because a neutralizing capacitor is used with a 1R5 to balance-out the grid 3 oscillator voltage due to internal capacitive coupling from grids 2 and 4 and not due to space charge coupling. Since the oscillator frequency voltage on grids 2 and 4 is developed across a tuned circuit the harmonic content is low, and thus with harmonic mixing the second harmonic voltage developed on grid 3 is low. Moreover, no other voltages of any magnitude appear on grid 3 because, at frequencies other than those in the vicinity of the oscillator second harmonic, the impedance between grid 3 and ground is small.

Thus a major difficulty experienced with fundamental oscillator frequency voltage on the control grid, is eliminated by harmonic operation. On the other hand the small amount of second-harmonic voltage that does appear on the grid 3 cannot be neutralized because no source consisting only of second harmonic voltages is available.

Another feature of harmonic operation of the 1R5, and similar types of converters, is that sensitivity is increased by negative input impedance at high frequencies due to a negative transconductance between the signal grid and screen and to large capacitance between these electrodes. The oscillator primary winding acts as a capacitive load at signal frequencies and feedback through the inter-electrode capacitance produces negative resistance across the signal-grid circuit.

In a normal case, the amount of regeneration is small, but if there is an unusually large impedance in the screen circuit, e.g. if a very large oscillator primary winding is used, the regeneration may become excessive.

The regeneration is not usually noticed when the oscillator fundamental is used for mixing because, as the screen circuit impedance is increased, the feedback of oscillator frequency voltage from screen to signal grid becomes excessive before noticeable signal-frequency regeneration occurs.

Another aspect of harmonic mixing, common to all types of converters, is that because the signal and oscillator circuits are tuned to widely differing frequencies, coupling between the two circuits can be ignored and variations of coupling within the valve due to changing electrode potentials have little effect on the frequency of the local oscillator. The improvement in stability during alignment is very noticeable and in most cases the need for rocking the tuning control while aligning the short-wave aerial trimmer disappears.

Frequency stability due to effects such as varying oscillator-input capacitance is not improved by harmonic oscillator operation because in general the oscillator tuning capacitance is unchanged at a given signal-frequency setting so that for a given capacitance change the frequency change is proportional to frequency. Although the oscillator operating on half the normal frequency has half the normal shift, the second harmonic will be no more stable than an oscillator fundamental on the same frequency.

(iii) Types of converters

A. Outer-grid oscillator injection

Converters using outer-grid oscillator injection always have a separate oscillator section and are characterized by relatively good oscillator-frequency stability, relative freedom from interaction between oscillator and signal circuits on high frequencies and positive input loading of the signal-grid circuit, which increases in proportion to the square of the frequency.

In the absence of undesired coupling between signal and oscillator-grid circuits, improved sensitivity will be obtained on short waves with these types by operating the oscillator at a higher frequency than the incoming signal.

(1) **Type 6J8-G** : The electrode arrangement of the triode-heptode type 6J8-G is shown in Fig. 25.23. This converter is very stable in operation which makes it useful in dual-wave receivers in spite of its low conversion transconductance. It has a high plate resistance (4 megohms) under recommended operation conditions so that the use of a high impedance first i-f transformer is more effective in increasing conversion gain than in the case of other converter types, for example the 6A8-G.

It is desirable not to apply a.v.c. to the 6J8-G on short waves if this is possible, but if a.v.c. is used a tuned-plate oscillator circuit will greatly improve the frequency stability.

(2) **Types 6AE8, X79 and X61M** : The electrode arrangements of these triode-hexodes is similar to that of Fig. 25.23 but there is no suppressor grid adjacent to the plate. Because of this it is necessary in circuit design to avoid conditions which may lead to secondary emission from the hexode plate to the screen.

Frequency stability with respect to B voltage variations is good, although it is desirable to use plate tuning of the oscillator if a.v.c. is applied to the hexode section on short waves. The conversion transconductances are high (approximately 750 micromhos with 2 volts bias and a screen voltage of 85) and few special precautions are required to obtain non-critical operation and good performance on the broadcast and short-wave bands.

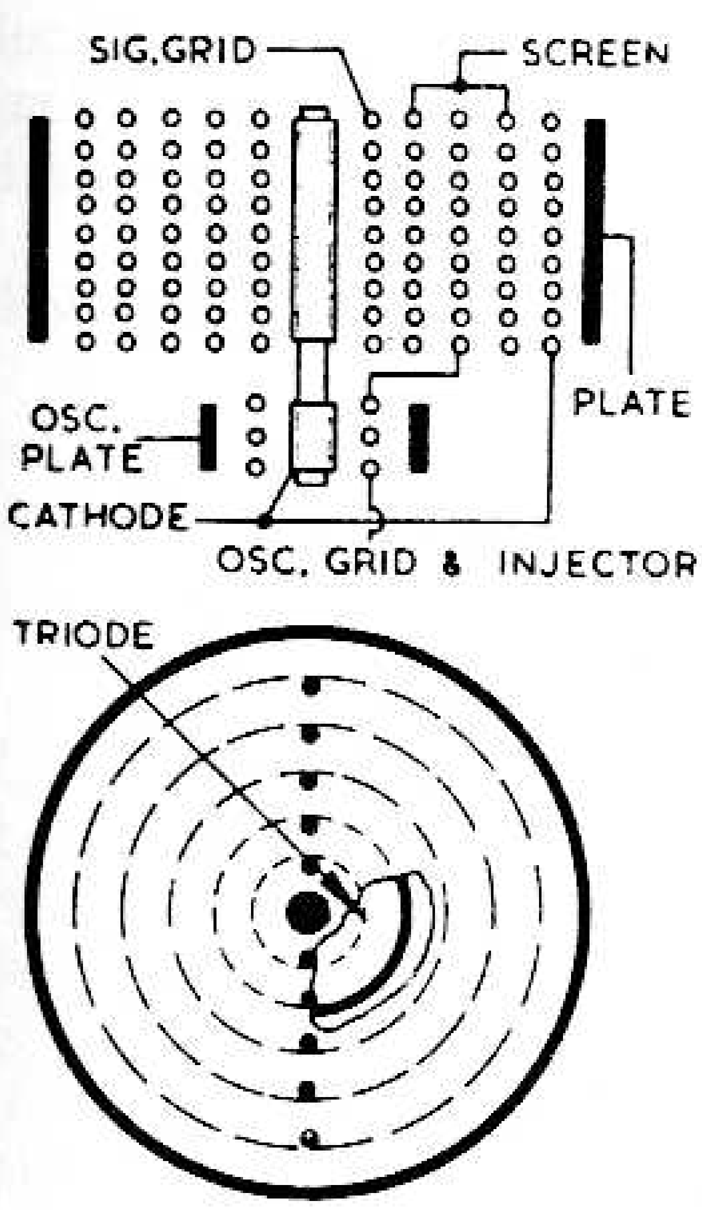


FIG. 25.23

Fig. 25.23. *Electrode arrangement of the triode-heptode type* 6J8-G.

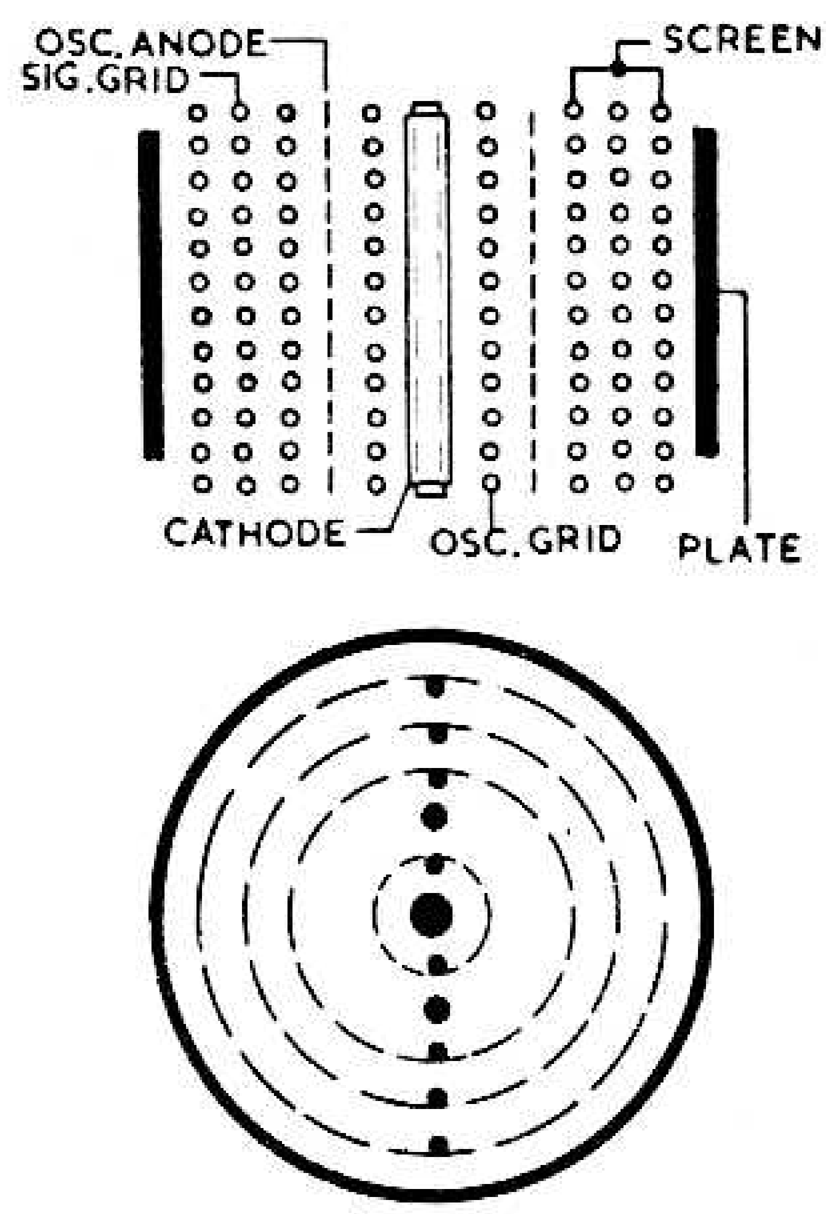


FIG 25.25

Fig. 25.25. *Electrode arrangement of the pentagrid type* 6*A*8(*G*).

The operating characteristics of most outer-grid injection triode-hexodes are similar to those of the above types.

B. Inner-grid oscillator injection

Converters using inner-grid oscillator injection, with the exception of the 6K8(G), have combined oscillator and mixer sections using a common cathode stream. In general the oscillator frequency stability is poorer than that of outer-grid injection types and coupling between control and signal grids is greater, although modern valves are much improved in these respects. Under maximum gain conditions the input loading of the signal grid is negative but becomes less negative and ultimately positive as the signal grid bias is increased.

To obtain maximum short-wave sensitivity without neutralizing, the oscillator should operate on the low-frequency side of the signal.

(1) **Type 6A8-G** : The structure of the pentagrid (or heptode) type 6A8-G is shown in Fig. 25.25. It will be noticed that the oscillator plate consists only of two side rods without a normal grid winding, and that there is no suppressor grid, which leads to a low plate resistance (0.36 megohm under typical operating conditions). Nevertheless the conversion transconductance of 550 micromhos is sufficient to give reasonable conversion gain.

When used with high impedance i-f transformers the 6A8-G may introduce some grid loading at the low-frequency end of the broadcast band due to feedback (Miller

Effect) from the plate circuit to the control grid through the relatively high (0.26 $\mu\mu$F) plate-to-grid capacitance. This effect is normally only just noticeable but may be aggravated by locating the 6A8-G against the back of the signal-frequency section of the tuning condenser so that appreciable capacitance is present between the plate of the valve and the stator assembly of the gang. The effect shows up as apparently poor aerial coil gain so that the cause may not be suspected. It is not of course peculiar to the 6A8-G but the already high plate-to-grid capacitance, the size of the valve and the lack of internal or external shielding make the effect more likely to occur with this type. The effect may be minimized by external shielding of the valve.

Part of the oscillator plate (grid 2) current is due to electrons which have passed from the cathode through grids 1 and 3 and then been repelled by grid 4 back through grid 3 to grid 2 again. The oscillator characteristics and thus the oscillator frequency stability are therefore very dependent on the potentials applied to grids 2, 3 and 4 and, for short wave operation, a.v.c. should not be applied to the valve. In addition it may be necessary to decouple the oscillator-plate voltage supply with an electrolytic capacitor and even to obtain it directly from the rectifier output—suitably decoupled to eliminate hum—in order to avoid feeding a-f variations from the output valve back to the oscillator plate and thus causing flutter.

To obtain consistent performance it is desirable to keep the oscillator-plate voltage higher than the screen voltage.

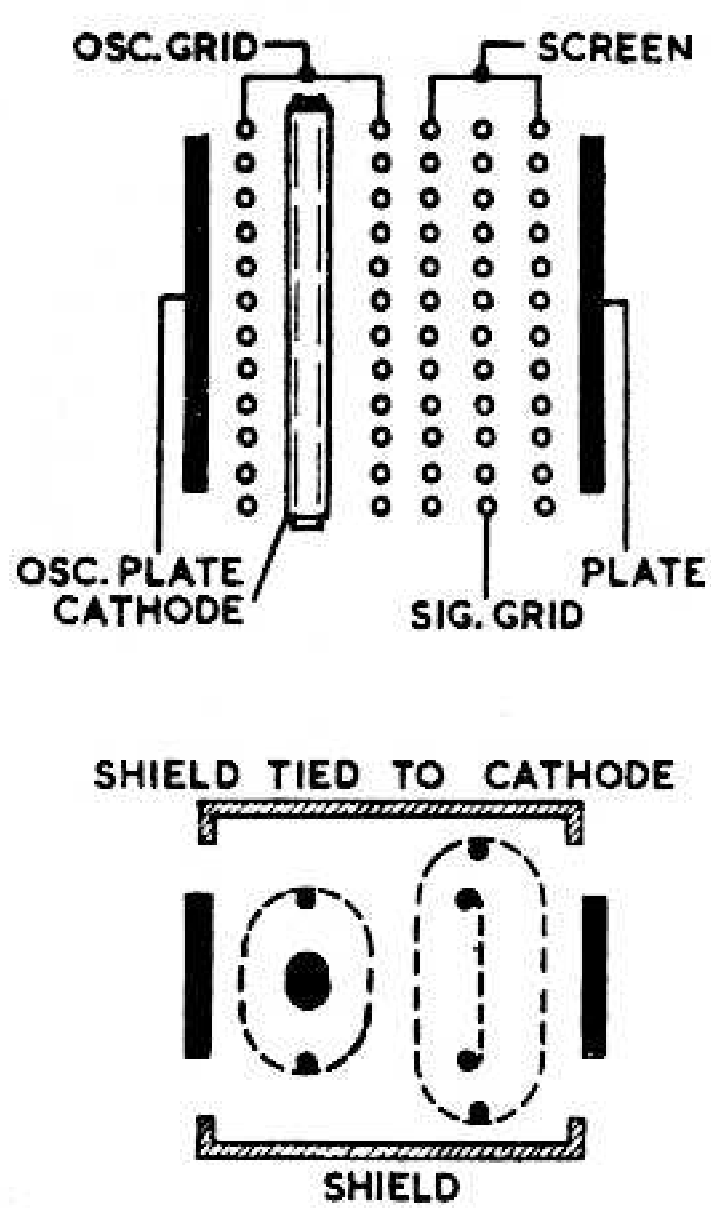


FIG. 25.26

Fig. 25.26. *Electrode arrangement of the triode-hexode type* 6K8(G).

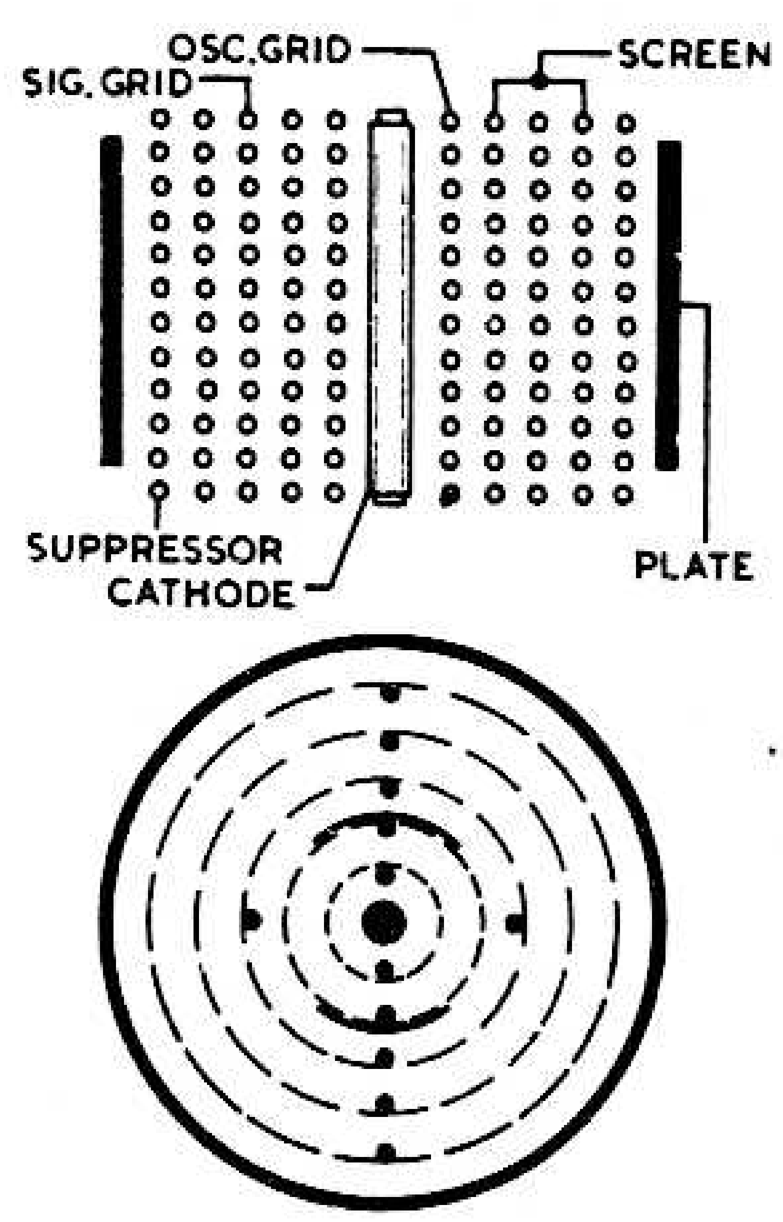


FIG. 25.27

Fig. 25.27. *Electrode arrangement of the pentagrid type* 6*SA*7(*GT*).

(2) **Type 6K8-G** : The structure of the inner-grid injection triode-hexode type 6K8-G is shown in Fig 25.26. It was designed to provide more stable short-wave performance than the 6A8-G and has a higher oscillator transconductance and operates with lower oscillator excitation. Its lower conversion transconductance (350 micromhos) gives a lower conversion gain than that obtainable from the 6A8-G although the plate resistance (0.6 megohm) is higher.

A.V.C. can be applied to the 6K8-G on short-waves but the frequency stability with respect to changes in other electrode voltages is only fair.

(3) **Types 6BE6 and 6SA7(GT)** : The main point of interest in the structure of the pentagrid types 6BE6 and 6SA7(GT), Figs. 25.27, and 25.29 is that there is no electrode which functions only as an oscillator plate.

By the omission of this electrode, it is possible to obtain a relatively high oscillator transconductance, without greatly increasing the total cathode current.

The oscillator circuits employed with these types have certain unconventional features and Fig. 25.28 may be taken as typical. The lack of a separate oscillator plate results in certain disadvantages in short-wave performance but because a voltage supply is required for one less electrode than usual the cost of components required is a minimum. The main difference between the 6BE6 and the 6SA7(GT) is the considerably higher oscillator transconductance of the 6BE6, due to the use of a formed No. 1 grid in this type.

When the circuit of Fig. 25.28 is used, the oscillator provides peak plate current at the instant when the oscillating voltage (E_k) on the cathode (with respect to ground)

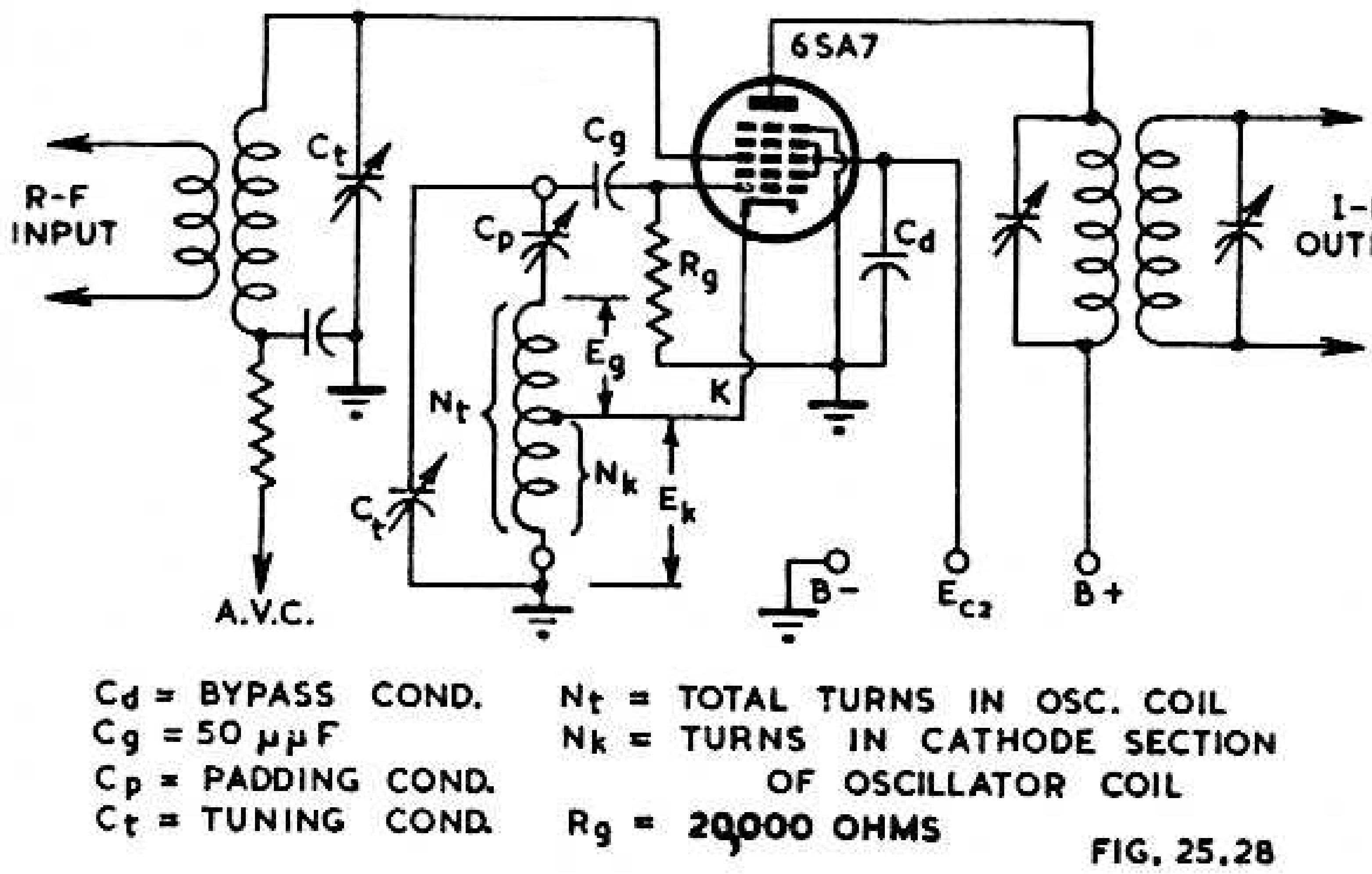


Fig. 25.28. *Typical self-excited converter circuit for type* 6*SA*7(*GT*).

and the oscillating voltage (E_g) on the No. 1 grid are at their peak positive values. For maximum conversion transconductance this peak value of plate current should be as large as possible. The effect on plate current of the positive voltage on the cathode is approximately the same as would be produced by an equal voltage of negative sign applied to the signal grid. Hence the amplitude of oscillator voltage on the cathode limits the peak plate current. This amplitude should therefore be small.

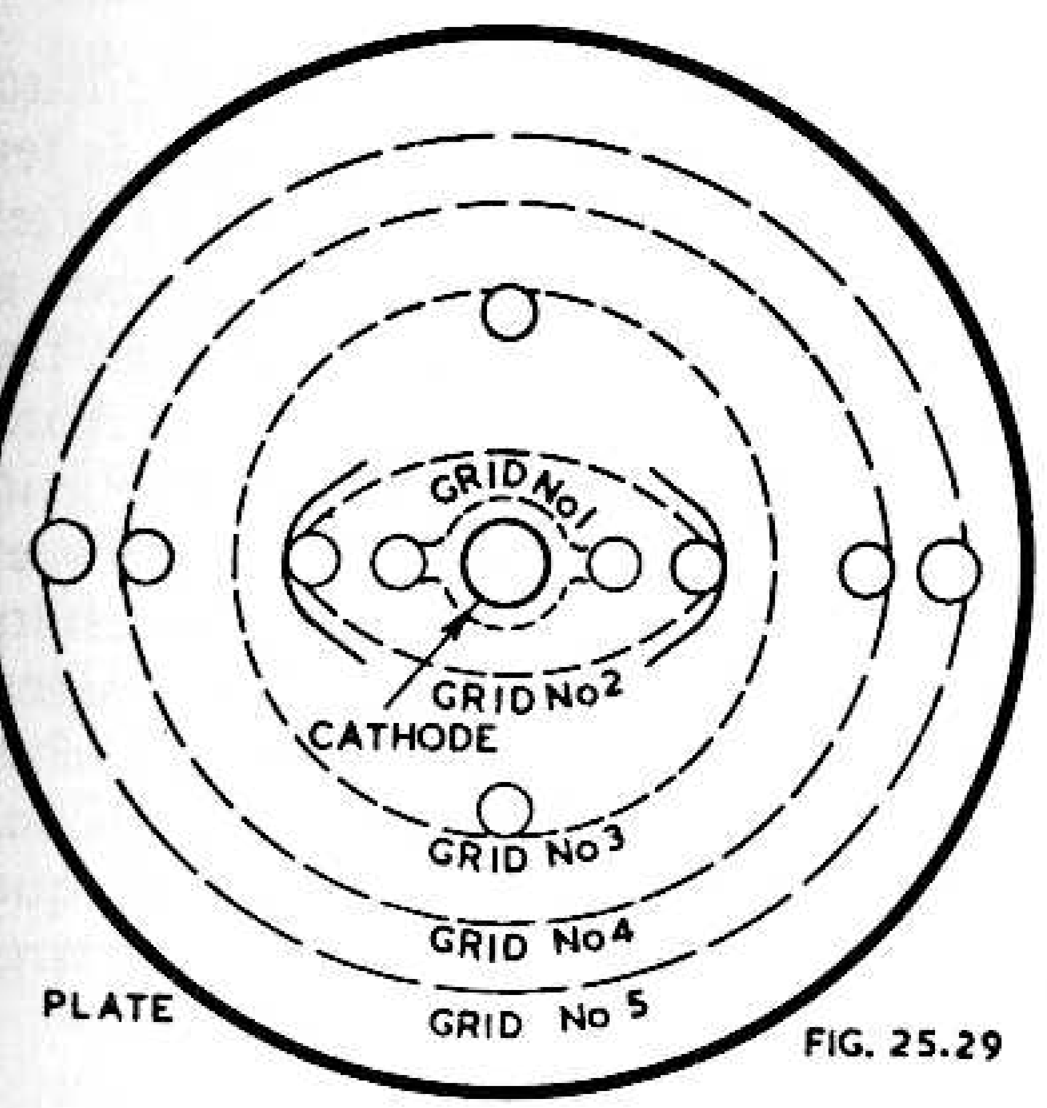


Fig. 25.29. *Electrode arrangement of pentagrid type* 6*BE*6.

During the negative portion of an oscillation cycle the cathode may swing more negative than the signal grid. If this occurs, positive signal-grid current will flow unless the oscillator grid is sufficiently negative to cut-off the cathode current. This signal-grid current flowing through the signal-grid circuit resistance will develop a negative bias on the signal grid and may also cause a negative bias to be applied to the i-f and r-f stages through the a.v.c. system. As a result, sensitivity will be decreased. In order to prevent signal-grid current, the d.c. bias developed by the oscillator grid should be not less than its cut-off value.

Because the peak plate current depends on how far positive the oscillator grid swings with respect to the cathode, it is desirable that this positive swing be as large as possible. It follows that the oscillator grid-leak resistance should be low, but not so low as to cause excessive damping of the tank circuit. It has been found, for operation in frequency bands lower than approximately 6 Mc/s, that all these requirements are generally best satisfied when the oscillator circuit is adjusted to provide, with recommended values of plate and screen voltage, a value of E_k of approximately 2 volts peak, and a d.c. oscillator-grid current of 0.5 mA through a grid-leak resistance of 20 000 ohms. This will give a peak positive voltage of the oscillator grid with respect to cathode of about 4 volts

On the normal short-wave band of 6 to 18 Mc/s, minimum grid current occurs at the low-frequency end of the band and the design procedure consists in adjusting the oscillator circuit so that sufficient grid current (200 μA minimum) is obtained at 6 Mc/s without developing excessive cathode voltage (approximately 2.5 volts r.m.s. maximum) at 18 Mc/s. The oscillator-grid bias is then somewhat less than cut-off at 6 Mc/s, but the signal-grid current should not be so high as to cause trouble. Oscillation at the high-frequency end of the band however, may be greater than optimum unless a grid stopper is used, but over-excitation will improve frequency stability.

If, for manufacturing reasons, the use of a tapped coil is not desirable, a normal primary winding can be used for the cathode connection, although it will be necessary to reverse the connections from those for a plate-tickler oscillator. This coil arrangement allows one side of the padder to be grounded.

Another method of connection which is satisfactory for the broadcast band is to use a plate-tickler oscillator circuit with the screen electrode as the oscillator plate and the cathode grounded. This connection can also be used on the short-wave band but the interaction between signal and oscillator circuits causes severe pulling and other associated troubles. A neutralizing capacitor between signal and oscillator grids will minimize these effects—neutralizing is not recommended with a cathode-coupled oscillator—and an alternative system is the use of harmonic mixing.

Whenever the screen is connected to an oscillator primary winding the oscillator voltage on this electrode must be kept to a minimum, i.e. primaries must have as few turns as possible with maximum coupling to secondaries. In addition, it is often desirable to use a small carbon resistor in series with the oscillator grid electrode to prevent the oscillator voltage on the screen from rising excessively at the high-frequency end of the band. The reasons are, firstly, that modulation by the oscillator voltage on the screen is out of phase with the modulation due to the oscillator grid, which decreases conversion sensitivity and, secondly, that as the signal grid is between the two electrodes forming the screen a large oscillator voltage on the screen results, through capacitive coupling, in a comparatively large oscillator voltage on the signal grid, particularly on short waves. This voltage can be neutralized, but normal neutralization is effective for only one frequency and set of operating conditions and when a large amount of neutralization is used and a balance between two large equal and opposite voltages is obtained, any small change in conditions results in a large oscillator voltage reappearing on the signal grid.

The arrangement of shields on grid 2 and of the siderods of grid 3 make the oscillator section of these types much more stable with respect to electrode voltage variations than the 6A8-G, and it can be used satisfactorily with a.v.c. applied on short waves.

Nevertheless in any short-wave application care must be taken to reduce to a minimum any coupling between signal and oscillator grids if frequency stability is to be satisfactory. In the case of the 6BE6, the external capacitive coupling between oscillator and signal grids should be limited to 1 $\mu\mu$F for satisfactory short wave performance.

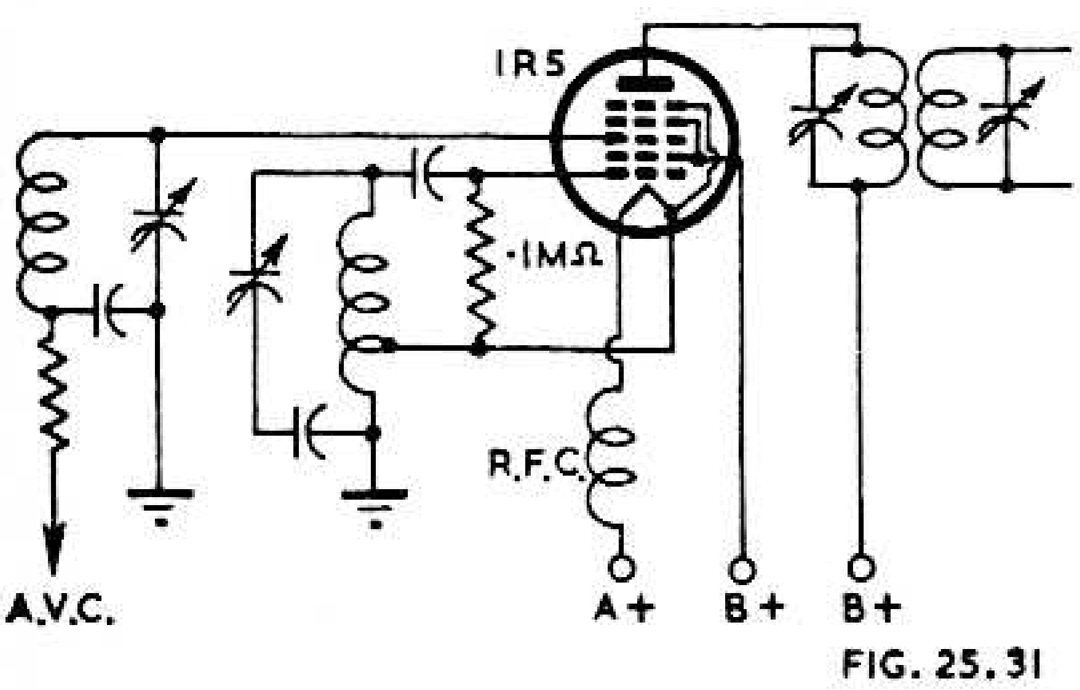


Fig. 25.31. *Self-excited* 1*R*5 *converter circuit.*

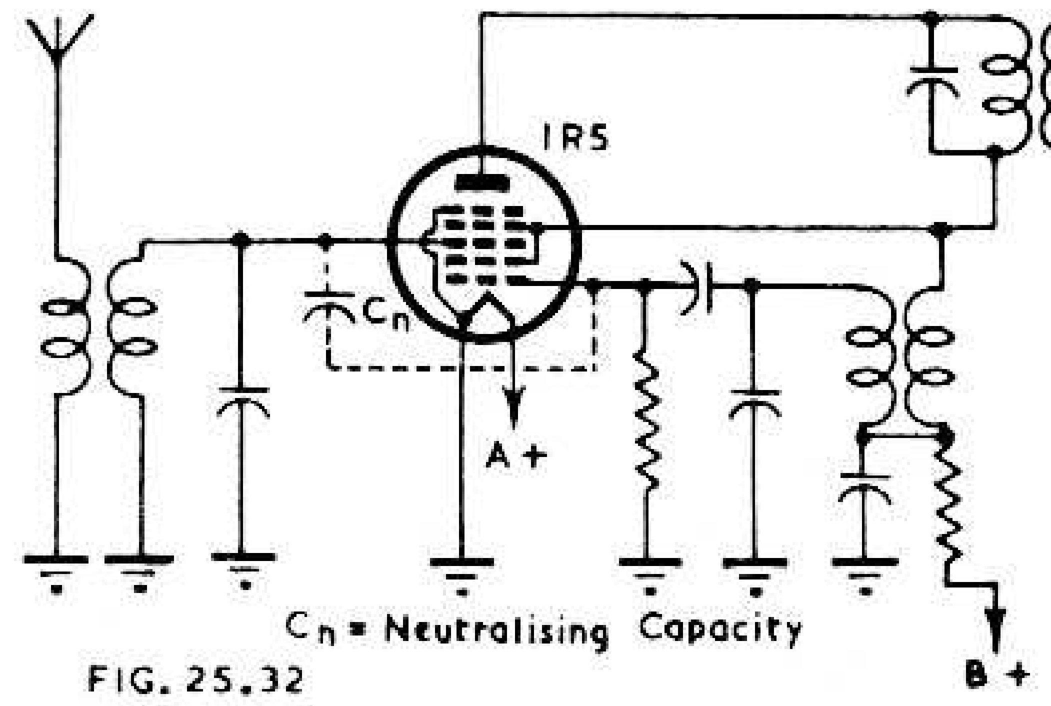


Fig. 25.32. *Circuit for short-wave operation of* 1*R*5 *converter.*

(4) **Type 1R5** : The electrode structure of the 1R5 is similar to that of the 6BE6 and 6SA7(GT) without the shields on grid 2, and with a filament instead of the heater and cathode. Circuits for use with the 1R5 are complicated by the lack of a separate cathode electrode but this may be overcome by using a choke as shown in Fig. 25.31 Alternatively the choke may be omitted and a filament lead returned to A+ through a separate winding of the same number of turns as, and wound over the tapped section of, the oscillator coil.

However, neither of these circuits is satisfactory on short waves if the full range—6 to 18 Mc/s—is to be covered, and even on the broadcast band there may be difficulty in obtaining sufficient grid current. The circuit of Fig. 25.32, with or without the padder feedback shown, is more suitable.

Neutralizing is desirable on short waves if conversion is carried out with the oscillator fundamental. The neutralizing capacitor should be connected across the short-wave coils only, because a different value is normally required on the broadcast band, and oscillator voltage on grid 3 is usually less with no neutralizing than with the capacitance (from 2½ to 5 $\mu\mu$F) required on short waves.

The purpose of the neutralizing capacitor in a circuit such as Fig. 25.32 is not so much the balancing out of oscillator voltage due to the space charge in front of grid 3, but the neutralizing of a voltage due to capacitive coupling between grid 3 and grids 2 and 4. Therefore, as previously mentioned, it is essential to reduce the oscillator voltage on the screen to the lowest practicable value.

If there is a tendency for the 1R5 to squeg at the high-frequency end of the short-wave band it is often possible to save the small carbon resistor in the oscillator grid lead, which is the usual cure, by reducing the value of oscillator grid leak to say 30 000 ohms or slightly lower. This frequently increases sensitivity, other conditions being unchanged, and it is worth trying even if there are no signs of squegging. In most receivers the oscillator grid circuit damping caused by the low value of oscillator grid leak or by the grid stopper is desirable to reduce the oscillator voltage on the screen of the converter at the high-frequency end of the band with a consequent improvement in sensitivity.

A very satisfactory way to operate the 1R5 is to use harmonic mixing as described previously. The increased freedom in layout, the increased stability in all respects—which simplifies production alignment—and the removal of the need for neutralizing more than compensate for the slight disadvantages which in any case are of little importance if the receiver has a r-f stage.

SECTION 3 : SUPERHETERODYNE TRACKING

BY B. SANDEL, A.S.T.C.

(i) General (ii) (A) Formulae and charts for superheterodyne oscillator design (B) Worked examples (iii) (A) Padded signal circuits (B) Worked example.

(i) General

The problem of tracking (whether it be in a straight or a superheterodyne receiver) is to set, simultaneously, to some desired resonant frequency, each of a series of tuned circuits which are mechanically coupled together and operated from a single control

In a superheterodyne receiver the problem is one of maintaining a constant frequency difference (equal to the intermediate frequency) between the signal circuits (such as the aerial and r-f stages) and the oscillator circuit. It is a relatively simple matter to make the difference between the signal and oscillator frequency equal to the i-f at two points in the tuning range. This condition is called two point tracking, and is applied whenever the error in frequency difference between the circuits does not become a large percentage of the total pass band. Where the tuning error is likely to become excessive, it can be reduced by the addition of another component, in the form of a capacitor or inductor, into the oscillator or signal circuits. In this case it is possible to proportion the circuit components so that zero frequency error exists between the intermediate frequency and the difference between the signal and oscillator circuits at three frequencies in the tuning range, instead of two ; furthermore the error at frequencies between the tracking points is appreciably reduced.

The most usual application of three point tracking has been in receivers covering the standard long, medium and short wave bands, and where the tuning element is a variable capacitor. For this reason attention will be confined to this system of tuning and a suitable design method will be detailed. Those interested in tracking permeability tuned circuits are referred to the articles of Refs. 107 and 111. These give a method of three point tracking using identical variable inductors in the signal and oscillator circuits.

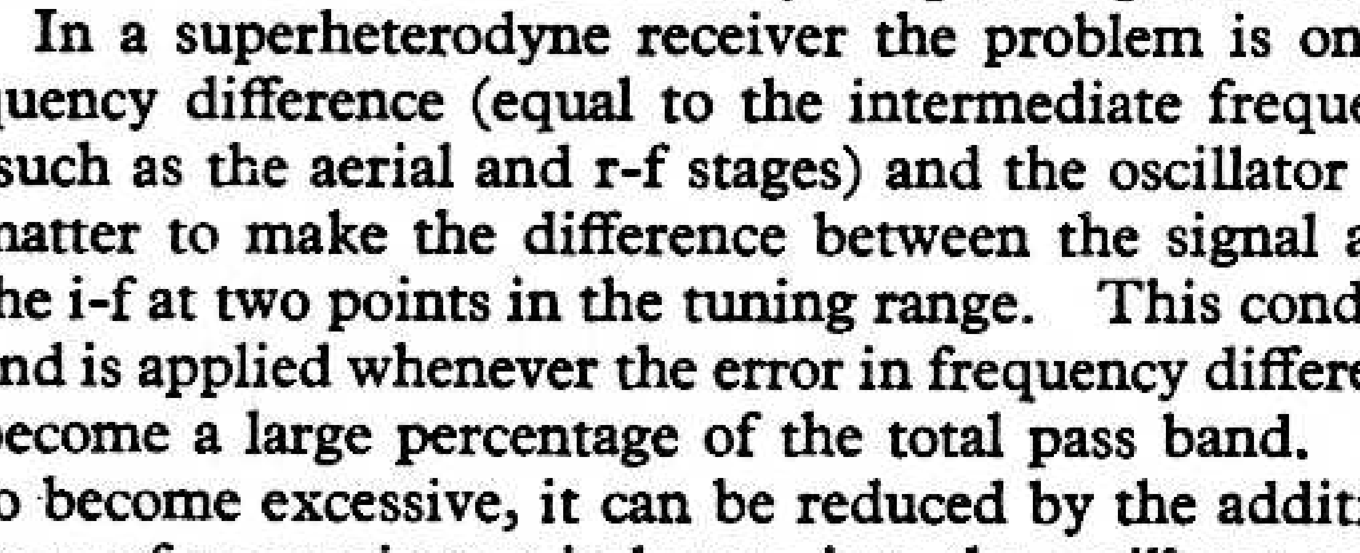


FIG. 25·33

Fig. 25.33. Circuits used for tracking analysis.

It is necessary to point out that the theoretical solutions, so far published, of the three point tracking problem are all idealized in so far that they ignore in part, or completely, the effects of primary windings on the signal and oscillator coils. The best approach under these conditions appears to be to select a method of determining values for circuit components which

(a) Does not involve an excessive number of arithmetical operations and will allow the use of a slide rule (or four figure logs) for all calculations, except in special cases.

(b) Gives values for the components that fall within a few per cent of those actually required in the circuits.

(c) Allows the change required in other circuit components to be rapidly estimated, when the value of one component in the circuit is changed by an amount which falls within a previously determined range of values.

(d) Lends itself to graphical methods of estimating the component values.

After the component values have been calculated it is usually advisable to build a pilot model receiver in which the values of all the elements (padder, coils and trimmers) can be varied over a small range (say $\pm$ 10%). In this way it is a fairly simple matter to secure tracking at the three points required, and at the same time to determine what tolerances are permissible in the component values before mistracking becomes excessive. Fixed padders and coils can then be used for further models of the same receiver ; although production difficulties can be reduced by using " slug " inductance variation in the oscillator coils as well as the usual variable parallel capacitance trimmers in the oscillator and signal circuits (many receiver manufacturers also use " slug " tuning of the signal circuits).

A few points which may be of interest are :

(1) The tracking error on the broadcast bands need not exceed a few Kc/s e.g. the error between tracking points need not exceed about 3 Kc/s on the 540-1600 Kc/s medium wave band using tracking frequencies of 600, 1000 and 1400 Kc/s. This error is negligible, in most cases, since the oscillator tuning takes charge and the lack of alignment affects the aerial and r-f circuits which are relatively unselective.

A typical curve of tracking errors is shown in Fig. 25.34.

(2) High impedance primaries, on the aerial and r-f coils, are practically always used in modern receivers covering the medium waveband. The coefficient of coupling in aerial and r-f coils, and the location of the primary resonant frequencies have very important effects on tracking. Suitable values for these factors are discussed in Chapter 23, Sects. 2 and 3.

(3) When a series capacitance (padder) is used in the oscillator circuit only, the operating frequency of the oscillator must be higher than the signal frequency if three point tracking is to be secured.

Three point tracking is obtainable if both the signal and oscillator circuits use series padders, whether the oscillator frequency is higher or lower than the signal frequency provided, of course, that the component values are correctly proportioned. If only the signal circuits are padded the oscillator frequency must be lower than the signal frequency to obtain three point tracking.

A typical example of padding of the signal and oscillator circuits occurs in band-spread receivers, where the short wave band is covered in a number of steps (e.g. 6 bands may be used to tune from 6-18 Mc/s), and it is required to have approximately straight line frequency tuning with a standard variable capacitor (i.e. one whose capacitance versus rotation approximates to a straight-line frequency law). The tuning law need not be linear, however, and the shape of the calibration curve is largely under the control of the receiver designer.

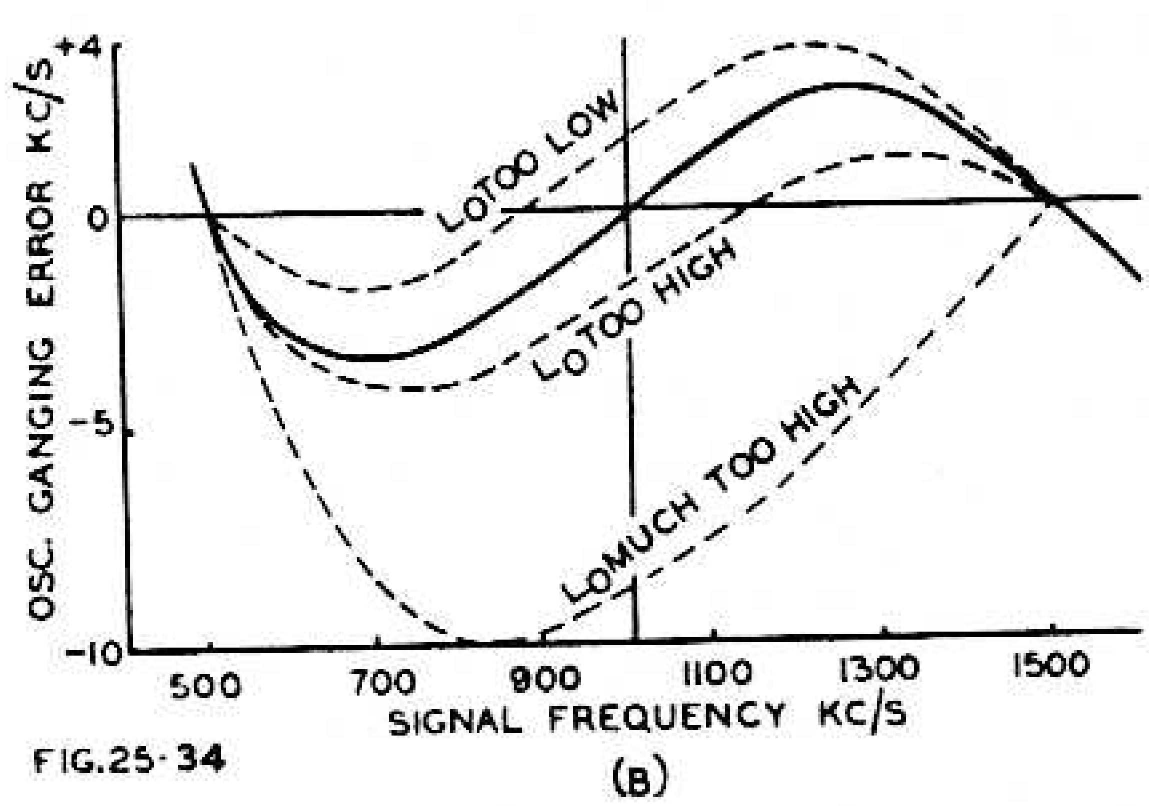


Fig. 25.34. *Typical tracking curves.*

(4) Care is necessary in the placement of components so as to minimize untracked stray capacitances. For example, the padder may be placed at the earthy end of the oscillator coil to assist in this regard.

(5) The selection of the best tracking points to give minimum error over a band of frequencies has received considerable attention from a number of designers. It has been generally accepted that the tracking error is reduced by setting the two outer tracking points somewhat in from the band limits, and locating the third tracking point at the geometrical-mean frequency (or some frequency near this value) of the outer tracking frequencies. However, it has been shown by Green (Ref. 105) that the **maximum** tracking error is reduced by bringing the high frequency point in from the band limit, setting the low frequency tracking point at or very close to the band limit, and making the third frequency the geometrical-mean of the two outer tracking frequencies. It has also been shown that the tracking errors are independent of the manner in which the total trimming capacitance is distributed across the oscillator coil and the variable capacitor ; although it should be noted that the manner in which these capacitances are distributed can have an adverse effect on the L/C ratio, particularly at low frequencies.

(6) Although the details set out in (5) are of interest they result in additional complication in the initial receiver design. A more suitable arrangement, for preliminary calculations, is to take the band limits as coinciding with the two outer tracking fre-

quencies, and make the third point either the arithmetical-mean or geometrical-mean of the outer frequencies. This leads to fairly simple design calculations, and gives values for the components which are reasonably close to the optimum. The small differences in component values required to set to the optimum tracking frequencies (or whatever the designer chooses) can then be made experimentally. It should be observed that it has been usual to align receivers to tracking points on various bands which have been established by long practice as giving sufficiently satisfactory results e.g. on the medium wave band 600, 1000 and 1400 Kc/s are in common use, but these are a compromise, due partly to a lack of more exact knowledge, and also because of the frequency allocations of the main broadcasting stations. If a station were to be located at, say, 540 Kc/s then the loss in receiver sensitivity, due to mistracking, would be serious if 600 Kc/s were to be taken as the low frequency tracking point.

(7) The feedback winding on the oscillator coil should be as small as possible, consistent with correct oscillator operation, if good tracking is desired. For minimum error the oscillator feedback winding should have its natural resonant frequency well above the highest oscillator tuning frequency ; this means that stray capacitances across the winding should be kept as low as possible. The natural resonant frequency is important because the coefficient of coupling between the two oscillator coil windings is usually large, and the amount of reactance reflected into the tuned circuit not only varies with frequency, but also sets a limit to the maximum possible tuning range. The limitation in tuning range is a particularly serious factor in receivers covering the short wave range in one band.

An additional and important reason for making the feedback winding as small as possible, is that for good oscillator stability the highest possible coefficient of coupling should be used, consistent with the smallest possible value of mutual inductance ($M = k\sqrt{L_p L_s}$).

(8) When tracking a superheterodyne receiver at three points in the tuning range (we will assume that the oscillator frequency is higher than the signal frequency) it often happens that the centre tracking (crossover) frequency does not fall at the frequency required. The question then arises as to how the oscillator series padder value should be altered.

First the signal circuit trimmer capacitance (or inductance) is altered so as to give maximum output. If the capacitance was increased then the oscillator series padder should be increased. For a decrease in signal trimmer capacitance (or inductance) the oscillator series padder should be decreased.

Complete re-alignment and re-checking is necessary after the padder value has been altered and it is important to remember that the oscillator inductance and trimmer capacitance values will also require alteration. This is a simple process when the oscillator coil is " slug " tuned, and the trimmer capacitance is adjustable.

Fig. 25.34 is also helpful in this regard, and the tracking error can be considered in terms of L_0 if so desired. It should be clear that if the value of L_0 is increased the padder value should be decreased (and vice versa) to retain the desired oscillator frequency coverage.

A good alignment procedure is to set first the signal and oscillator circuits at the band limits. Tracking is then obtained at the required points by setting the signal generator to the required tracking frequency and rocking the receiver dial while altering the oscillator trimmer capacitance (at the high frequency tracking point) and the oscillator inductance (at the low frequency tracking point) until maximum signal output is obtained. Tracking at the centre frequency can be checked either by alteration of the signal circuit trimmers, which have not been altered after the initial adjustment at the band limits, or, alternatively, by leaving the signal circuit trimmers untouched and rocking the receiver dial while adjusting the oscillator inductance (or capacitance) for maximum output. It is preferable to track the oscillator to the signal circuits since these give the required tuning law. A complete re-check is always necessary after the initial alignment procedure is completed Of course correct tracking can be obtained by a number of methods, but the procedure suggested above has proved very satisfactory for receiver development.

(ii) (A) Formulae and charts for superheterodyne oscillator design

The equations quoted below are due to Payne-Scott and Green (Ref. 102) and Green (Refs. 103, 105). Design charts are available in the references for the cases of arithmetical-mean and geometrical-mean tracking. The charts shown in Figs. 25.35 to 25.39 are for geometrical-mean tracking only.

For the circuit arrangements used in the derivation of these equations see Fig. 25.33.

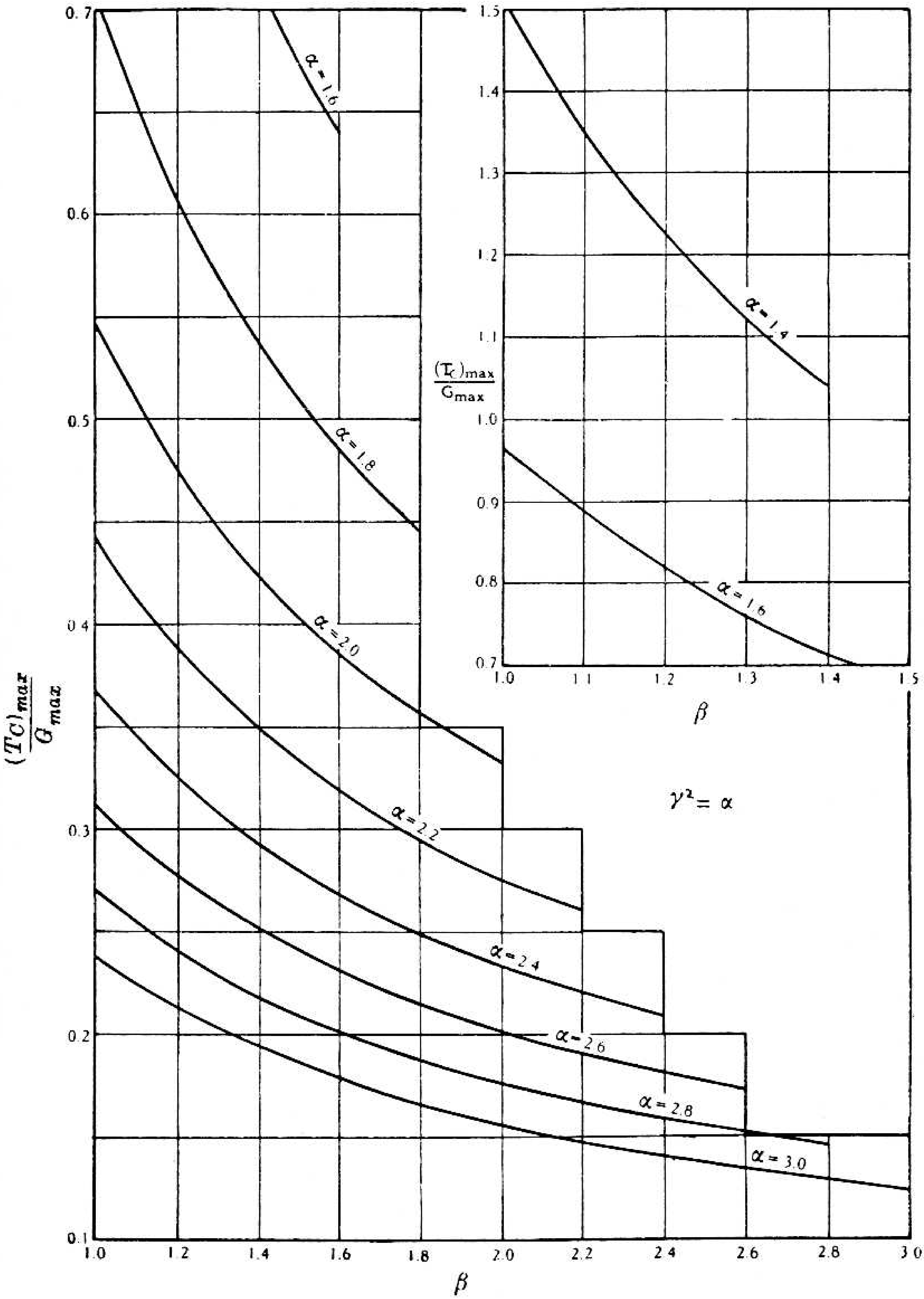


FIG. 25.35

Fig. 25.35. Charts giving $T_{c_{max}}$ for geometrical-mean tracking.

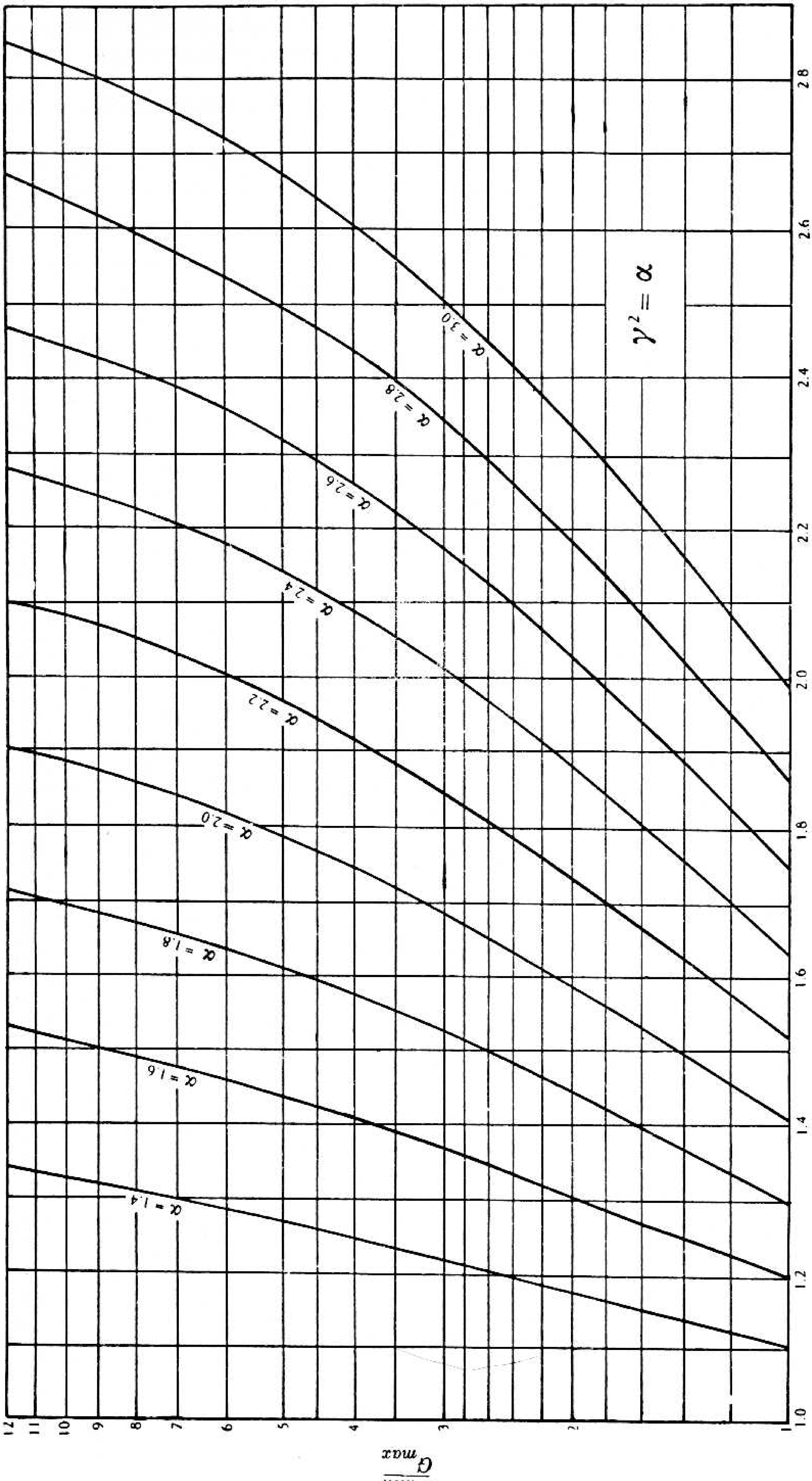


FIG. 25.36

Fig. 25.36. *Charts giving high values of* P_{min} *for geometrical-mean tracking.*

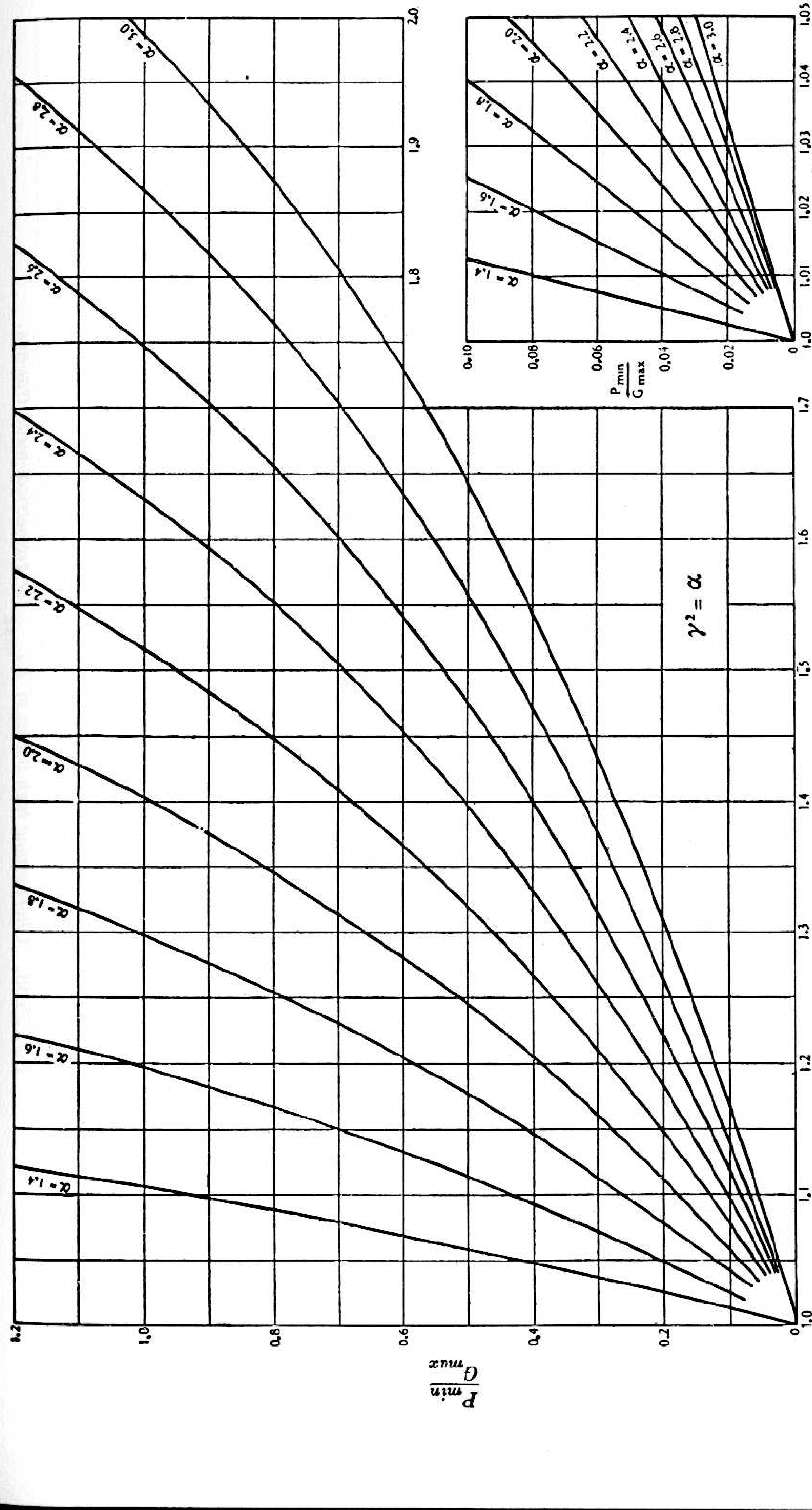


FIG. 25.37

Fig. 25.37. Chart giving low values of P_{min} for geometrical-mean tracking.

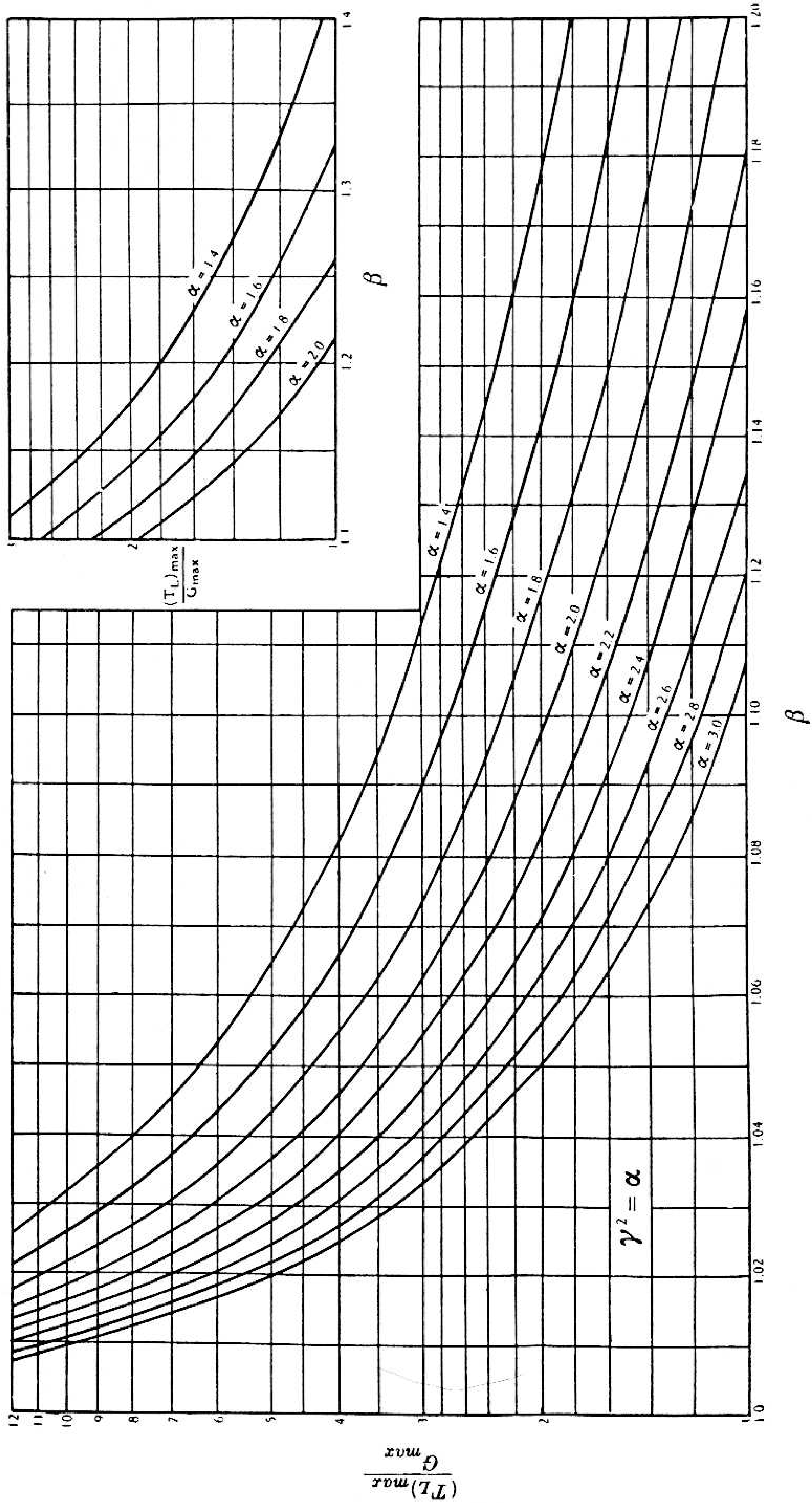


FIG. 25.38

Fig. 25.38. Chart giving high values of $T_{L_{max}}$ for geometrical-mean tracking.

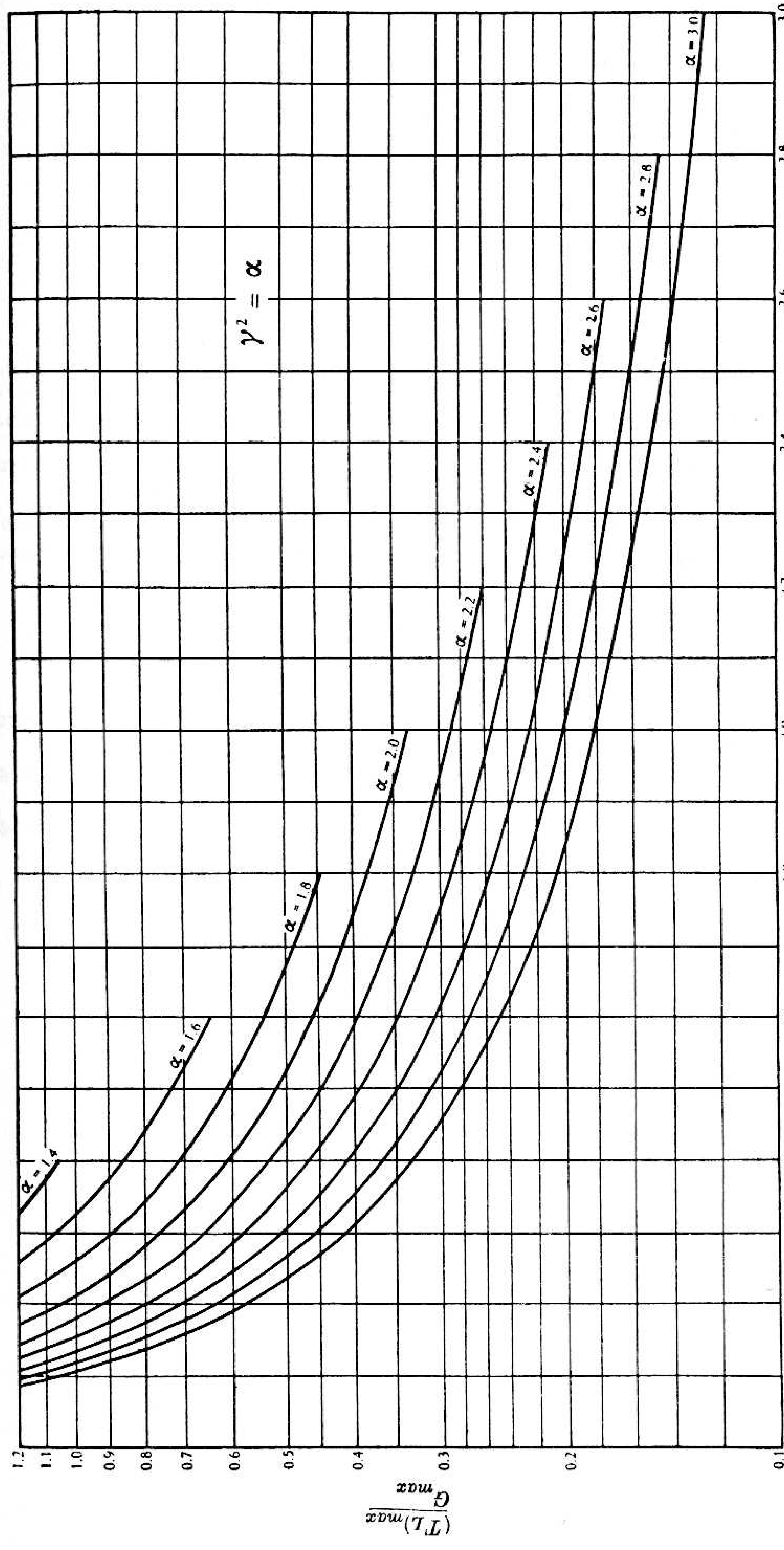


FIG. 25.39

Fig. 25.39. Chart giving low values of $T_{L_{max}}$ *for geometrical-mean tracking.*

Notation for the equations

$\omega_1 = 2\pi \times$ low-frequency tracking point of the signal circuit (f_1)
$\omega_2 = 2\pi \times$ high-frequency tracking point of the signal circuit (f_2)
$\omega_3 = 2\pi \times$ third tracking frequency for the signal circuit (f_3)
$\omega_i = 2\pi \times$ intermediate frequency (f_i)

$$\alpha = \frac{\omega_2}{\omega_1} = \frac{f_2}{f_1}$$

$$\beta = \frac{\omega_2 + \omega_i}{\omega_1 + \omega_i} = \frac{f_2 + f_i}{f_1 + f_i}$$

$$\gamma = \frac{\omega_3}{\omega_1} = \frac{f_3}{f_1}$$

L, L_0 = signal and oscillator inductances
G_{max} = incremental capacitance of each section of the ganged capacitors (i.e. difference between the maximum and minimum values of the capacitor)
T = capacitance in signal circuit at signal frequency f_2 (includes gang min. cap.)
T_L = fixed capacitance in parallel with L_0
T_c = fixed capacitance in parallel with G in the oscillator circuit
P = oscillator padding capacitance.

When the extreme tracking points do not coincide with the band limits, the following additional notation is required.

$\omega_1' = 2\pi \times$ low-frequency limit of signal band (f_1')
$\omega_2' = 2\pi \times$ high-frequency limit of signal band (f_2')

$$\alpha' = \frac{\omega_2'}{\omega_1'} = \frac{f_2'}{f_1'}$$

G'_{max} = total incremental capacitance of each section of the ganged capacitors between signal frequencies f_1' and f_2'
T' = T less the incremental capacitance of each section of the ganged tuning capacitors between signal frequencies f_2 and f_2'
T_c' = T_c less the incremental capacitance of each section of the ganged tuning capacitors between signal frequencies f_2 and f_2'.

Signal circuits :

For the tracking points coincident with the band limits.

$$T = \frac{G_{max}}{\alpha^2 - 1};$$

$$L = \frac{1}{T\omega_2^2} \quad \left(\text{or } L = \frac{25\,330}{Tf_2^2}\ \mu\text{H where } T \text{ is in } \mu\mu\text{F and } f \text{ is in Mc/s}\right)$$

Oscillator circuit :

For the tracking points coincident with the band limits.

$$P_{max} = \frac{G_{max}}{r - 1}$$

$$T_{c_{max}} = \frac{G_{max}}{r\beta^2 - 1}$$

and
$$r = \frac{\alpha^2}{\beta^2} \cdot \frac{1 + \dfrac{2(\alpha - \beta)}{(\beta - 1)(\alpha + \gamma)}}{1 + \dfrac{2(\alpha - \beta)}{(\beta - 1)(1 + \gamma)}}$$

For the case of arithmetical-mean tracking

$$\left(\text{i.e. } f_3 = \frac{f_1 + f_2}{2}\right)$$

$$r = \rho = \frac{\alpha^2}{\beta^2} \cdot \frac{3 + \alpha}{3 + \beta} \cdot \frac{1 + 3\beta}{1 + 3\alpha}$$

For geometrical-mean tracking (i.e. $f_3 = \sqrt{f_1 f_2}$)

$$r = R = \frac{\alpha^2}{\beta^2} \cdot \frac{2\beta + (1 + \beta)\alpha^{\frac{1}{2}}}{2\alpha + (1 + \beta)\alpha^{\frac{1}{2}}}$$

(a) General formulae for padding and trimming capacitances.

(1) T_c given.

$$P = P_{max} - T_c = P_{min} + T_{c_{max}} - T_c$$

$$T_L = \frac{P}{P_{min}}\left[T_{c_{max}} - T_c\right]$$

(2) T_L given.

$$P = \frac{P_{min}}{2}\left(1 + \sqrt{\frac{P_{min} + 4T_L}{P_{min}}}\right)$$

$$T_c = P_{max} - P = P_{min} + T_{c_{max}} - P.$$

Formulae when $T_L \ll P$

$$P = P_{min} + T_L$$

$$T_c = T_{c_{max}} - T_L$$

error in P is given by $-\left(\frac{T_L}{P}\right)^2 \times 100$ per cent.

error in T_c is given by $\frac{T_L^2}{P \times T_c} \times 100$ per cent.

(b) Formulae for inductance of oscillator coil

$$L_0(\omega_2 + \omega_i)^2 = \left(\frac{P_{max}}{P}\right)^2 \cdot \frac{1}{T_{L_{max}}} = \frac{P_{min}P_{max}}{T_{c_{max}}} \cdot \frac{1}{P^2}$$

$$\text{or } L_0 = \frac{25\,330 \times P_{min}P_{max}}{T_{c_{max}} P^2(f_2 + f_i)^2}$$

(capacitances in $\mu\mu$F, frequencies in Mc/s)

(c) Relations between P_{max}, P_{min}, $T_{c_{max}}$, and $T_{L_{max}}$

$$P_{max} = P_{min} + T_{c_{max}}$$

$$\frac{P_{max}}{P_{min}} = \frac{T_{L_{max}}}{T_{c_{max}}}$$

(d) Relations between oscillator-circuit components.

The effect of a change in value of one component of the oscillator circuit on the values required for the other components can be found from the following expressions :

$$\delta T_c = -\delta P$$

$$\delta T_L = \left(1 + \frac{2T_L}{P}\right)\delta P$$

$$\delta L_0 = -\frac{2L_0}{P}\delta P$$

where δL_0, etc. means a small change in L_0, etc. These changes must give values of P, T_c etc. which lie between the limits of the maximum and minimum values (P_{max}, P_{min}, etc.) of the components considered.

(e) Auxiliary formulae when the outer tracking frequencies do not coincide with the band limits

$$G_{max} = \left(\frac{\omega_2'}{\omega_2}\right)^2 \cdot \frac{\alpha^2 - 1}{(\alpha')^2 - 1} \cdot G'_{max}$$

$$\frac{T'}{T} = \left(\frac{\omega_2}{\omega_2'}\right)^2$$

$$T_c - T_c' = T - T'$$

(B) Worked examples

To illustrate the application of the equations and charts, worked examples are appended.

A tuning capacitor is available having a capacitance range of 12-432 $\mu\mu$F. It is required to tune over the range of frequencies 530 to 1620 Kc/s, and the i-f of the receiver is 455 Kc/s. Determine the component values foı the signal and oscillator circuits.

For ease of working the outer tracking frequencies will be made coincident with the band limits, and the third tracking frequency (f_3) will be taken as the geometrical-mean of the outer frequencies ($\sqrt{f_1 f_2}$ = 926.6 Kc/s).

Signal circuits

$$G_{max} = 432 - 12 = 420 \ \mu\mu\text{F}$$

$$\alpha^2 = \left(\frac{1620}{530}\right)^2 = 9.342$$

$$T = \frac{420}{9.342 - 1} = 50.24 \ \mu\mu\text{F}$$ (includes 12 $\mu\mu$F gang min. cap. so that actual trimmer would be 38.24 $\mu\mu$F)

$$L = \frac{25\,330}{50.24 \times 1.62^2} = 192.2 \ \mu\text{H}$$

Oscillator circuit

$$\alpha^2 = \left(\frac{1620}{530}\right)^2 = 9.342 \ ; \ \alpha^{\frac{1}{2}} = 1.748 \ ;$$

$$\beta = \frac{1620 + 455}{530 + 455} = 2.107 \ ; \ \beta^2 = 4.440 \ ;$$

$$(1 + \beta)\alpha^{\frac{1}{2}} = 5.431$$

$$r = R = \frac{9.342}{4.44}\left[\frac{(2 \times 2.107) + 5.431}{(2 \times 3.056) + 5.431}\right] = 1.758$$

$$P_{max} = \frac{432 - 12}{1.758 - 1} = 554 \ \mu\mu\text{F}$$

$$T_{c_{max}} = \frac{(432 - 12)}{(1.758 \times 4.44) - 1} = 61.7 \ \mu\mu\text{F}$$

(a) A value for T_L (about 8 $\mu\mu$F) will be assumed, after preliminary calculation, as this is the usual case.

As obviously $T_L \leqslant P$ (Compare 8 $\mu\mu$F with P_{max})

$$P_{min} = P_{max} - T_{c_{max}} = 492.3 \ \mu\mu\text{F}$$

$$T_{L_{max}} = \frac{554 \times 61.7}{492.3} = 69.42 \ \mu\mu\text{F}.$$

From this it is permissible to estimate T_L as 8 $\mu\mu$F since this is about the order of stray capacitances likely to be across L_0. Any small error here is unimportant, as long as the value chosen for T_L is less than $T_{L_{max}}$, since this is taken up during circuit alignment.

Hence $P = 492.3 + 8 = 500.3 \ \mu\mu\text{F}$

$T_c = 61.7 - 8 = 53.7 \ \mu\mu\text{F}$ (includes gang min. cap. of 12 $\mu\mu$F. Actual trimmer would be 41.7 $\mu\mu$F)

$$\text{(b) } L_0 = \frac{25\,330 \times 492.3 \times 554}{61.7 \times 500.3^2 \times (1.62 + 0.455)^2} = 103.9 \ \mu\text{H}$$

(c) and (d)

Suppose now that it is required to make a change in the value of one of the oscillator components. What should be the new values of the other components?

A good example is the padder; a more suitable value may be 510 $\mu\mu$F (which lies between the maximum and minimum values of 554 $\mu\mu$F and 492.3 $\mu\mu$F).

Then

$$\delta T_c = -9.7 \ \mu\mu\text{F}$$

$$\delta T_L = \left(1 + \frac{2 \times 8}{500.3}\right) 9.7 = 10 \ \mu\mu\text{F}$$

$$\delta L_0 = -\frac{2 \times 103.9 \times 9.7}{500.3} = -4.03 \ \mu\text{H}.$$

So that the new values for the oscillator components are $P = 510 \ \mu\mu\text{F}$

$T_c = 53.7 - 9.7 = 44 \ \mu\mu\text{F}$

$T_L = 8 + 10 = 18 \ \mu\mu\text{F}$

$$L_0 = 103.9 - 4.03 = 99.87\ \mu\text{H}$$

and these values all lie within the permissible range of values, set by the maximum and minimum values calculated for P, T_L and T_c; denoted by P_{max}, P_{min} etc.

A table showing a series of values between the calculated maxima and minima is often very useful. In this way the most suitable component values can be selected. Also, three point tracking is still maintained at the selected frequencies with the range of component values determined, as the change in values is independent of the tracking and intermediate frequencies provided these remain unaltered.

The same design problem, previously solved algebraically, will now be carried out using the tracking charts.

Signal circuits

Procedure exactly as before.

Oscillator circuit

$$\alpha = \frac{1620}{530} = 3.056$$

$$\beta = \frac{1620 + 455}{530 + 455} = 2.107$$

$$G_{max} = 432 - 12 = 420\ \mu\mu\text{F}$$

(a) From the charts:

$$\frac{T_{c_{max}}}{G_{max}} = 0.15\text{; therefore } T_{c_{max}} = 63\ \mu\mu\text{F}$$

$$\frac{P_{min}}{G_{max}} = 1.2\text{, therefore } P_{min} = 504\ \mu\mu\text{F}$$

$$\frac{T_{L_{max}}}{G_{max}} = 0.17\text{, therefore } T_{L_{max}} = 71\ \mu\mu\text{F}$$

$$P_{max} = 504 + 63 = 567\ \mu\mu\text{F}$$

Using $T_L = 8\ \mu\mu\text{F}$ (as previously)

$$P = 504 + 8 = 512\ \mu\mu\text{F}$$

$$T_c = 63 - 8 = 55\ \mu\mu\text{F}$$

$$\text{(b) } L_0 = \frac{25\,330 \times 504 \times 567}{63 \times 512^2 \times (1.62 + 0.455)^2} = 102\ \mu\text{H}$$

(c) and (d) Any circuit component changes are made exactly as before.

Suitable practical values are then selected after allowing for strays.

(iii) (A) Padded signal circuits

The case of the multi-band all-wave receiver is of interest to designers. It can hardly be said that the receiver covering the complete short wave band (6 − 18 Mc/s) in one step gives very satisfactory performance, particularly in the hands of an unskilled operator.

Consider first the case where bandspreading is obtained merely by loading additional parallel capacitance across the tuned circuits. The scale calibration will become non-linear with frequency and the scale will be crowded towards the low-frequency end of the band. If, instead of using parallel capacitance, a small capacitance is inserted in series with the tuning capacitor so as to restrict the tuning range, the scale will be crowded towards the high frequency end. From this it should be clear that a combination of the two methods can probably be made to give very much improved scale linearity. The method is not restricted, however, to producing a linear scale calibration.

The circuit arrangement for the padded signal circuit takes the same form as the oscillator circuit. The analysis of these circuits, due to Green (Ref. 103), is based on tracking the padded signal and oscillator circuits to a (non-existent) pilot circuit (which also determines the form of scale calibration) at three points, and thereby making the padded circuits track with one another at the same three points. The error at frequencies between tracking points is usually not serious as the range of frequencies to be covered is generally fairly small.

The design problem is to select values for the padder P and the two trimming capacitances T_L and T_c such that simultaneously the following conditions are fulfilled :—

(a) The desired frequency ratio is attained.

(b) The scale calibration is linear, or, more generally, corresponds to a desired form.

(c) The L/C ratio is maintained at a high value in order to achieve adequate gain in the signal circuits.

To simplify notation the virtual pilot circuit takes the same form as the original signal circuit shown in Fig. 25.33 and the same notation is retained, for both the padded signal and oscillator circuits, as was previously used for the oscillator circuit alone. The design equations are limited to the cases of arithmetical-mean and geometrical-mean tracking. Charts are available (in Ref. 103) covering the arithmetical-mean case.

Summary of formulae

Pilot circuit :

$$T = \frac{G_{max}}{\alpha^2 - 1}$$

$$\left(\frac{\omega_2}{\omega}\right)^2 = 1 + (\alpha^2 - 1)\frac{G}{G_{max}}$$

Padded circuit :

$$P + T_c = \frac{G_{max}}{r - 1}$$

$$T_{c_{max}} = \frac{G_{max}}{r\beta^2 - 1}$$

$$T_{L_{max}} = \frac{G_{max}}{r(\beta^2 - 1)}$$

$$P + T_c = P_{max} = P_{min} + T_{c_{max}}$$

where r is limited to the values given by ρ and R.

T_c specified :

$$P = P_{max} - T_c$$

$$T_L = \frac{P}{P_{min}} T_{c_{max}} - T_c$$

where P is first evaluated.

T_L specified :

$$P = \frac{P_{min}}{2}\left[1 + \sqrt{1 + \frac{4T_L}{P_{min}}}\right]$$

$$T_c = T_{c_{max}} - T_L \cdot \frac{P_{min}}{P}$$

where P is first evaluated.

When $T_L \ll P$

$$T_c = T_{c_{max}} - T_L$$

$$P = P_{min} + T_L$$

Gain is proportional to the dynamic impedance Z of the circuit and

$$Z = \frac{Q}{\omega C};$$

$$C = \frac{P^2}{P_{min}} \cdot \frac{G + T_{c_{max}}}{G + P_{max}} \text{ (effective capacitance across inductance).}$$

When $G = G_{max}$ then $C = C_{max}$. The value of P is obtained from the appropriate equations, according to whether T_c or T_L is specified.

Inductance

$$L_0 = \frac{1}{(\omega_1 + \omega_j)^2} \cdot \frac{P_{min}}{P^2} \cdot \frac{1 + \dfrac{P_{max}}{G_{max}}}{1 + \dfrac{T_{c_{max}}}{G_{max}}}$$

or for numerical computation $L_0 = \dfrac{25\,330\ P_{min}(G_{max} + P_{max})}{(f_1 + f_j)^2 P^2 (G_{max} + T_{c_{max}})}$

using μH, $\mu\mu$F, Mc/s.

in which $(\omega_1 + \omega_j)$ is the known low frequency tracking point of the padded signal circuit. For a padded oscillator circuit replace $(\omega_1 + \omega_j)$ by $(\omega_1 + \omega_j \pm \omega_i)$. The value for P is derived from the appropriate equation according as to whether T_c or T_L is specified.

For an unpadded signal circuit

$$L = \frac{\alpha^2 - 1}{\alpha^2 \omega_1^2\ G_{max}}$$

The notation is identical with that given previously, with the addition of

$\omega = 2\pi \times$ any frequency in the tuning range

$\omega_1 = 2\pi \times$ low frequency tracking point of pilot circuit

$\omega_2 = 2\pi \times$ high frequency tracking point of pilot circuit

$\omega_3 =$ arithmetical- or geometrical-mean tracking point of pilot circuit

$\omega_i = 2\pi \times$ true intermediate frequency of receiver

$\omega_j = 2\pi \times$ virtual intermediate frequency for the combination of a padded signal circuit with a virtual pilot circuit

$\alpha = \dfrac{\omega_2}{\omega_1} = \dfrac{f_2}{f_1} =$ frequency ratio of pilot circuit

$\beta = \dfrac{\omega_2 + \omega_j}{\omega_1 + \omega_j} =$ frequency ratio of padded signal circuit

$G =$ incremental capacitance of each section of the ganged tuning capacitors, measured from its value at ω_2

$G_{max} =$ value of G at ω_1

$C =$ total effective capacitance across L_0

$C_{max} =$ value of C when $G = G_{max}$

$Q = \omega L_0/R$ where L_0 and R apply to the signal circuit, and this R is the r-f resistance of the circuit.

(B) Worked example

The method will be applied to the design of a receiver tuning from 6 to 9 Mc/s, using an i-f of 455 Kc/s, with a capacitor having a straight line frequency characteristic and a capacitance range of 12-432 $\mu\mu$F. Arithmetical-mean tracking will be used.

Pilot circuit

For the pilot circuit assume that the required scale calibration is given by

$\alpha = 2$ (A sufficient range of values for most receiver calibration curves is from about 3 to 1.5)

$\alpha^2 = 4$

Signal circuits

$\beta = 9/6 = 1.5$

$\beta^2 = 2.25$

$$r = \rho = \frac{4}{2.25} \cdot \frac{3 + 2}{3 + 1.5} \cdot \frac{1 + (3 \times 1.5)}{1 + (3 \times 2)} = 1.552$$

$$P_{max} = P + T_c = \frac{432 - 12}{1.552 - 1} = 760.8\ \mu\mu\text{F}$$

$$T_{c_{max}} = \frac{432 - 12}{(1.552 \times 2.25) - 1} = 301.7\ \mu\mu\text{F}$$

$$T_{L_{max}} = \frac{432 - 12}{1.552\ (2.25 - 1)} = 216.5\ \mu\mu\text{F}$$

$P_{min} = 760.8 - 301.7 = 459.1 \ \mu\mu F$

Take the case where T_c is specified. Draw up a table of component values, and select the most suitable.

T_c	T_L	P	L_0
50		710.8	
100		660.8	
150		610.8	
200		560.8	
250	57.53	510.8	2.024
300		460.8	

From this it seems that a convenient padder value is 510.8 $\mu\mu$F. Then find T_L and L_0 to complete the table. If the values so found, using the first selection for P, are considered as being suitable, only those need be computed

$$T_L = \frac{510.8}{459.1}[301.7 - 250] = 57.3 \ \mu\mu F$$

$$L_0 = \frac{25\,330 \times 459.1 \times (420 + 760.8)}{6^2 \times 510.8^2 \times (420 + 301.7)} = 2.024 \ \mu H$$

So that the signal circuit components are

$L_0 = 2.024 \ \mu H$

$T_L = 57.3 \ \mu\mu F$

$P = 510.8 \ \mu\mu F$

$T_c = 250 \ \mu\mu F$ (which includes 12 $\mu\mu$F gang min. cap.)

(Suitable practical values are selected after due allowance for strays).

An estimate can now be made of the circuit dynamic impedance (Z), as this will serve as a guide to the possible circuit gain. Improvement can often be effected by a redistribution of the component values, and the most favourable L/C ratios can be determined by completing the table set out earlier.

However the L/C ratio and the component values selected are usually a compromise forced on the designer by considerations of practical convenience.

Oscillator circuit

It will be taken that the oscillator frequency is higher than the signal frequency.

$$\beta = \frac{9 + 0.455}{6 + 0.455} = 1.464$$

$$\beta^2 = 2.145$$

$$r = \rho = \frac{4}{2.145} \cdot \frac{3 + 2}{3 + 1.464} \cdot \frac{1 + (3 \times 1.464)}{1 + (3 \times 2)} = 1.609$$

$$P_{max} = \frac{432 - 12}{1.609 - 1} = 689.6 \ \mu\mu F$$

$$T_{c_{max}} = \frac{432 - 12}{(1.609 \times 2.145) - 1} = 171.4 \ \mu\mu F$$

$$T_{L_{max}} = \frac{432 - 12}{1.609\,(2.145 - 1)} = 228 \ \mu\mu F$$

$$P_{min} = 689.6 - 171.4 = 518.2 \ \mu\mu F.$$

Take the case, this time, where T_L is specified and draw up a table of component values. T_L is not necessarily very much smaller than P.

T_L	T_c	P	L_0
50		564.1	
100	85.6	604	1.619
150		639.7	
200		1072	

Then for the tabulated values of T_L we have

$$P = \frac{518.2}{2}\left[1 + \sqrt{1 + \frac{4T_L}{518.2}}\right]$$

These values are listed above.

Taking the padder of 604 $\mu\mu$F as being suitable, then T_L is 100 $\mu\mu$F, and now find T_c and L_0.

$$T_c = 171.4 - 100\left(\frac{518.2}{604}\right) = 85.6 \ \mu\mu\text{F}$$

$$L_0 = \frac{25\,330 \times 518.2\ (420 + 689.6)}{(6 + 0.455)^2 \times 604^2\ (420 + 171.4)} = 1.619 \ \mu\text{H}$$

So that one suitable set of oscillator component values would be

$T_L = 100$ $\mu\mu$F, $P = 604$ $\mu\mu$F, $L_0 = 1.619$ μH

and $T_c = 85.6$ $\mu\mu$F (including 12 $\mu\mu$F gang min. cap.).

The signal and oscillator circuits should track at 6, 9 and 7.5 Mc/s.

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