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A REFLEX KLYSTRON OSCILLATOR FOR THE 8–9 MM BAND

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SUMMARY

The paper describes a reflex klystron oscillator, tunable over the wavelength range of 8–9 mm, which is suitable both for use in a superheterodyne receiver and as a source for laboratory measurements.

The cavity operates in its fundamental mode and tunes smoothly over the wavelength band. It is mounted in a metal envelope having a glass window through which the output power passes.

At an input of 2000 volts 10 mA, the output power of an average valve is about 30–50 mW, and the electronic tuning range to half-power is 60 Mc/s. Powers up to 160 mW have been obtained.

The factors affecting the performance of the valve are discussed, and consideration is given to the noise power generated by the oscillator when used in a superheterodyne receiver.

LIST OF PRINCIPAL SYMBOLS

The M.K.S. system of units is used throughout.

β = Beam coupling factor.
R = Shunt resistance of the unloaded resonator.
I = Effective resonator current.
N = Overall noise factor of superheterodyne receiver.
L = Conversion loss of crystal mixer.
t = Noise temperature ratio of crystal (n.t.r.).
t' = Excess n.t.r. due to local-oscillator noise.
$N_{i.f.}$ = Noise factor of intermediate-frequency amplifier.
P_n = Local oscillator noise power contained in the two noise sidebands.
T = Absolute temperature.
B = Bandwidth of receiver.
P = Power output from local oscillator.
f_0 = Natural resonant frequency of the oscillator.
ω = Angular frequency.
F' = A bunching factor which depends upon the potential distribution in the reflector space.
s = Total drift length.
V_0 = Beam potential at the resonator-gap centre.
Q_L = Loaded Q-factor.
ϕ = Difference of phase between the arrival of the electron bunches at the resonator and the maximum electric field across the gap.
C_0 = Equivalent capacitance at the resonator gap.

(1) INTRODUCTION

Since the 1939–45 War attention has been given to the development for radar purposes of still shorter wavelengths than those formerly used, with a view to obtaining, amongst other advantages, aerials of smaller size for a given resolution. One such band selected for exploitation was in the wavelength range of 8–9 mm, where there exists a region of minimum atmospheric absorption. A tunable low-power oscillator for this band, suitable for use in superheterodyne receivers and also for bench measurements, was therefore needed.

The first design to operate in this band was of disc-seal construction, the cavity oscillating in a harmonic mode, and was developed by workers at the Clarendon Laboratory, Oxford, from an earlier valve for the 1·25 cm band. Samples of this valve were made both at the Radar Research Establishment (formerly the Telecommunications Research Establishment) and at the Services Electronics Research Laboratory, but by late in 1947, difficulties had been encountered both in manufacture and performance (these were later much reduced) and there appeared to be a need for an alternative design. The paper describes the outcome of the development started at that time, the final valve being known as type VX5023. A magnetron for this band has also been developed and is described in a companion paper.[1]

Within about six months the basic design had been evolved by means of a series of experiments in a demountable system; thereafter attention was turned to making sealed-off valves incorporating a tuning mechanism, and the development was virtually complete by 1950.

It was decided at the outset to use a high-voltage input of about 2 kV; this decision was governed mainly by practical considerations. The value chosen was thought to be the minimum that would give the performance required, using a resonator having gridless apertures. The grids that would have been necessary in a low-voltage valve for this frequency would have been very difficult to make owing to the minute dimensions involved, and might well have been capable of dissipating only a very limited amount of power, thus limiting the input power and efficiency. Another advantage of the high-voltage valve, which has been found to be of importance in this band during the last few years, is that local-oscillator noise in a superheterodyne receiver is less.

With regard to the cavity, the choice lay between fundamental- and harmonic-mode operation. The harmonic type has the advantage of being bigger and therefore easier to tune and to handle generally. The fundamental type, however, has two advantages: first, the copper losses are less than for any other cavity and the efficiency is therefore greater, and secondly, the frequency may be varied smoothly over a 10% band without sudden variation of output, whereas this is not always possible with certain types of harmonic cavity.

The fundamental cavity turned out to be of about 4 mm diameter and 1 mm depth, the apertures being of about 0·5 mm diameter. These dimensions are very small, but experiment showed that, by using a thin diaphragm for one wall of the cavity, the capacitance could be varied sufficiently to tune the frequency over about 10%. A cavity of the fundamental type was therefore chosen. More recent developments of harmonic cavities show that the disadvantages are not necessarily as great as was believed in 1947, and such cavities can therefore be useful in oscillators of this type.

On considering possible methods of construction, the primary necessity was to ensure great accuracy of assembly, particularly with regard to axial alignment of the gun, resonator and reflector. It was therefore decided to machine the resonator from a copper block, and to locate the gun and reflector with respect to this block. The copper block was mounted inside a thin steel envelope, the oscillatory power being coupled to an external waveguide through a glass window of the type which had been

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used on magnetrons. The resonator was placed in good thermal contact with the metal envelope, thus ensuring adequate cooling and minimizing frequency drift.

(2) THE BASIC ELECTRICAL DESIGN

The method of arriving at the final design was to determine as many of the relevant dimensions as possible by a combination of calculation and previous experimental knowledge, and to find the remainder by empirical means.

Several difficulties stand in the way of designing more completely by calculation, but the principal one is concerned with electron optics. At the high current densities required in valves of the type under consideration, the electron paths are determined more by mutual repulsion between electrons than by the laws of geometrical electron optics. The reflex klystron is complicated by the fact that the electron beam reverses direction near the reflector, and under these conditions the electron trajectories cannot be calculated with any accuracy.

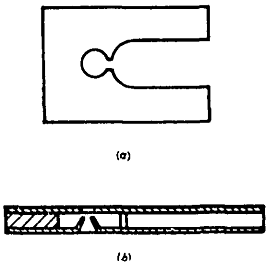


Fig. 1.—The basic resonator design.

It is important to keep the resonator apertures as small as possible, consistent with passing substantially the whole of the electron beam, since the resonator losses are otherwise too big (shunt resistance R too low) and the coupling factor (β) too low. The size chosen will of course depend upon the magnitude of the electron current which it is proposed to use (assuming that the input voltage has already been fixed). The current which would give the required output power was calculated by the method given in Section 5, taking the best estimates possible for quantities such as the shunt resistance R of the resonator. It appeared that a current in the region of 8–12 mA at 2 000 volts would be adequate, and for this current a resonator aperture of 0·020 in diameter was chosen.

It is advantageous in a reflex oscillator to have the resonator aperture on the reflector side slightly larger than that on the gun side. By this means a number of marginal electrons that would otherwise be lost are allowed to enter the resonator after reflection and to give up energy, which more than compensates for the lower coupling factor (β). For the larger aperture the diameter chosen was 0·025 in.

The best value for the separation of the two apertures—the "gap"—is given by the condition that β^2R should be a maximum.[2] For small gaps β is high and R low, whilst for large gaps the opposite obtains. In between these extremes a broad optimum exists. β is readily calculable,[3] but, as is shown later, R can be determined only in an approximate manner. Nevertheless the optimum β^2R, and hence the optimum gap, may be found by such calculations sufficiently accurately to enable experimental work to start.

The remaining resonator dimensions, namely the diameter and the depth, are easily determined, at least approximately, from existing data. Slight changes may be necessary to get the exact frequency required when the resonator is made up and tested. Some choice of the ratio of diameter to depth is available, and in this case a fairly large ratio was chosen so as to give as large a diameter as possible to the flexible diaphragm.

The first experiments were carried out with a continuously-pumped demountable valve. This was the simplest way to establish a suitable design for the electron gun and reflector, to check the frequency of the cavity, and to make adjustments if necessary. The resonator cavity and output section were machined from a copper plate [Fig. 1(*a*)], the cavity being completed by two plates in which the apertures had been formed. Good joints were ensured by silver-plating all three parts and

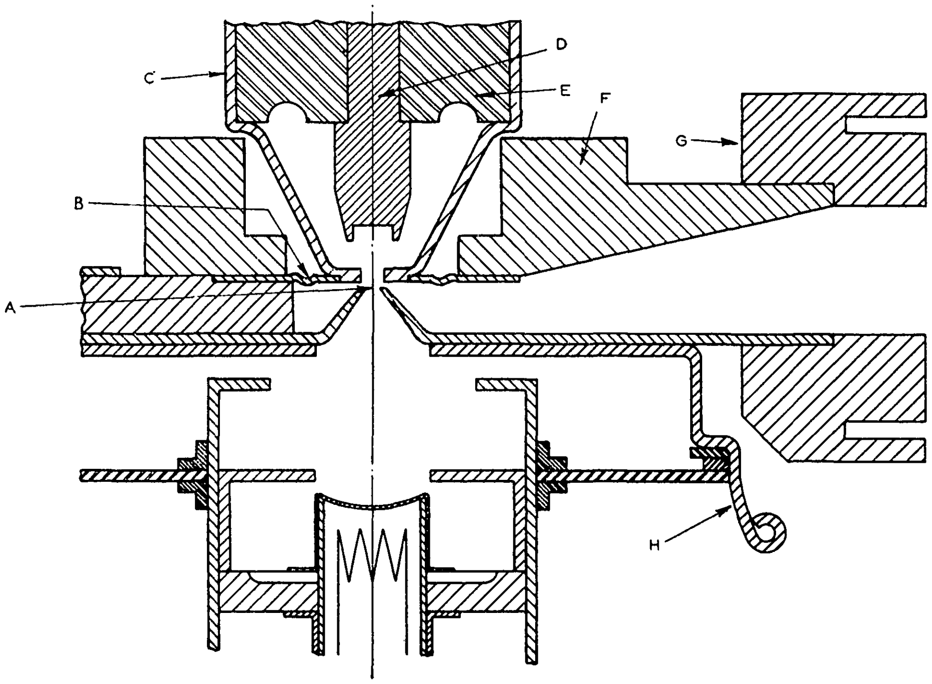


Fig. 2.—Section showing resonator and electrode assembly.

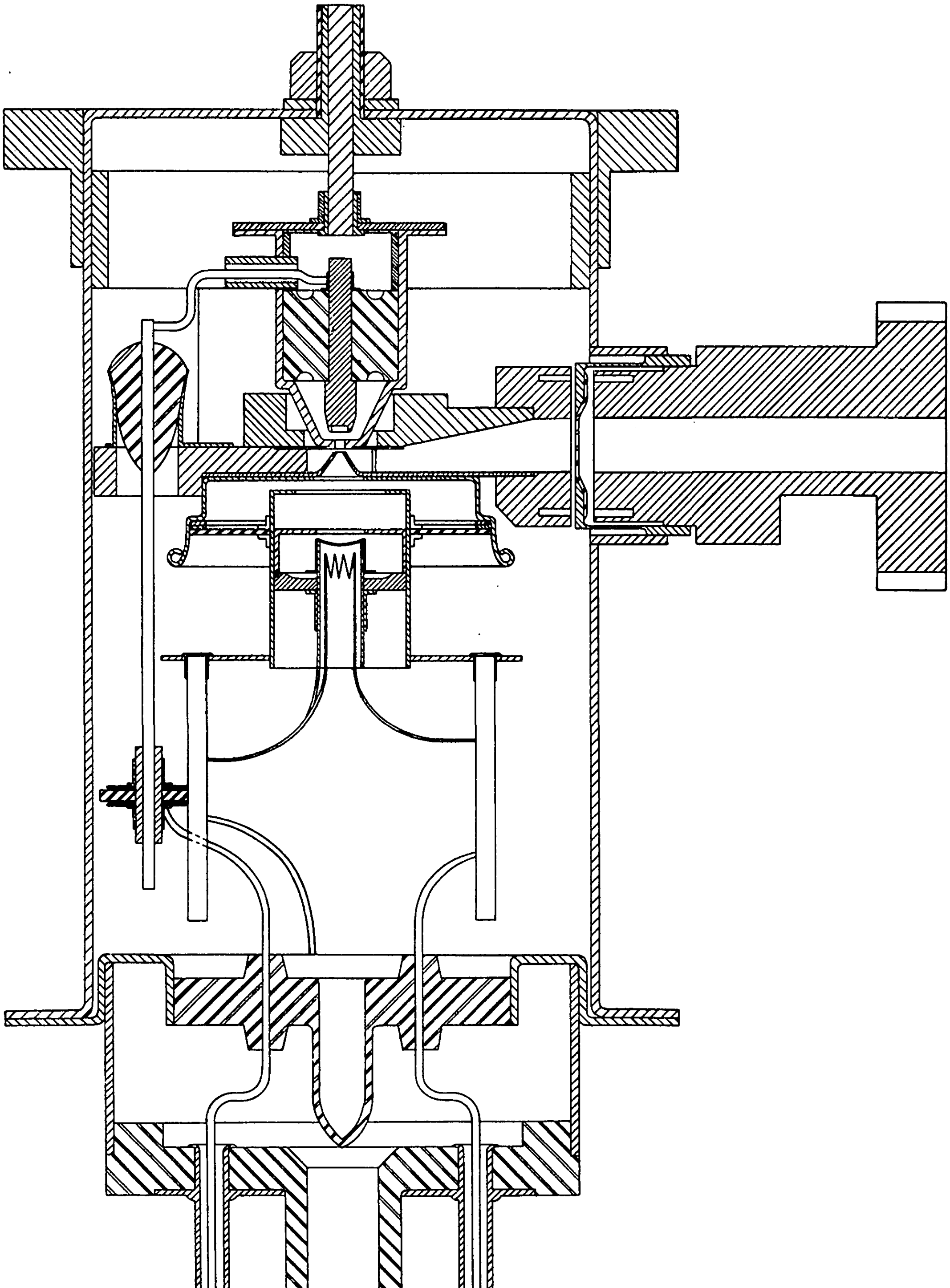


Fig. 3.—Sectional view.

brazing together after assembly in a jig. A sectional view of the assembly is shown in Fig. 1(*b*).

The design of the electron gun was based on an earlier design of an oscillator for the 3 cm band,[4] but was modified for this application. The oxide cathode was of 3 mm diameter and had a concave surface of radius of curvature $\frac{3}{16}$ in. This was surrounded by a focusing electrode, of the shape shown in Fig. 2, which normally operated at a negative potential with respect to the cathode. It was found that the efficiency of transmission through the resonator aperture was about 90%.

The reflector was of the conventional dish shape, and of 0·050 in diameter and 0·025 in depth. Other dimensions were tried in the demountable system before these values were decided.

Whilst these experiments were being carried out it was found that very accurate alignment of the electrodes and apertures was necessary. If the reflector were as little as 0·002 in off the position for maximum power, the output dropped to one-half. The gun position was not so critical.

After a series of demountable-system experiments, a performance fairly close to that expected was obtained, and the decision to start making sealed-off valves was taken. During this stage the technique of assembling the valve and mounting it within the metal envelope in a vacuum-tight manner was studied. Sundry small changes were incorporated; these included slight variation of the output coupling slot and variation of the gap for maximum output until a satisfactory performance was obtained.

(3) CONSTRUCTION

(3.1) The Resonator Assembly

The resonator assembly previously described was adapted for use in the sealed-off valve, in particular to provide means of tuning. Various ways of tuning were considered before it was decided to adopt the method of capacitance variation by flexing one wall of the resonator. The tuning diaphragm B (Fig. 2) was corrugated to improve flexibility and joined to a nickel tube C, which housed the reflector D, and could be rigidly connected by brazing to the top of the valve envelope. This was sufficiently flexible to be capable of being moved from outside by means of the mechanism described later. The reflector was a metallic rod with a cup-shaped portion machined concentrically in the end facing the resonator diaphragm, and was held in place by an accurately ground ceramic insulator E. With this construction the reflector can be located on the axis to within 0·001 in. A taper from the resonator output coupling slot up to the full depth of the standard waveguide was formed in the copper block F; the latter also acted as a stop to prevent the diaphragms from being damaged during tuning.

(3.2) The General Assembly

A sectional view of the valve (less tuner) is shown in Fig. 3. The output window consisted of a disc of glass sealed across an aperture in a cup-shaped pressing made of iron-nickel-cobalt alloy, which was brazed to a tubular metal side-arm on the envelope. On the outside, a short length of waveguide was soldered into the window cup, and was terminated at its far end in a standard coupler. The waveguides facing the window on either side were fitted with ditch chokes—so as to avoid loss of output power. The resonator assembly was attached by means of screws to a metal rim welded to the envelope. By using a jig when joining the rim to the envelope, accurate location of the resonator assembly and choke G (Fig. 2) with respect to the window was ensured. When the dimensions of the window iris, glass thickness and spacings between the window and two ditch chokes were correctly maintained, the voltage standing-wave ratio of the output run was less than 1·2 over the wavelength range of 8–9 mm. A photograph of the valve complete with tuning mechanism is shown in Fig. 4.

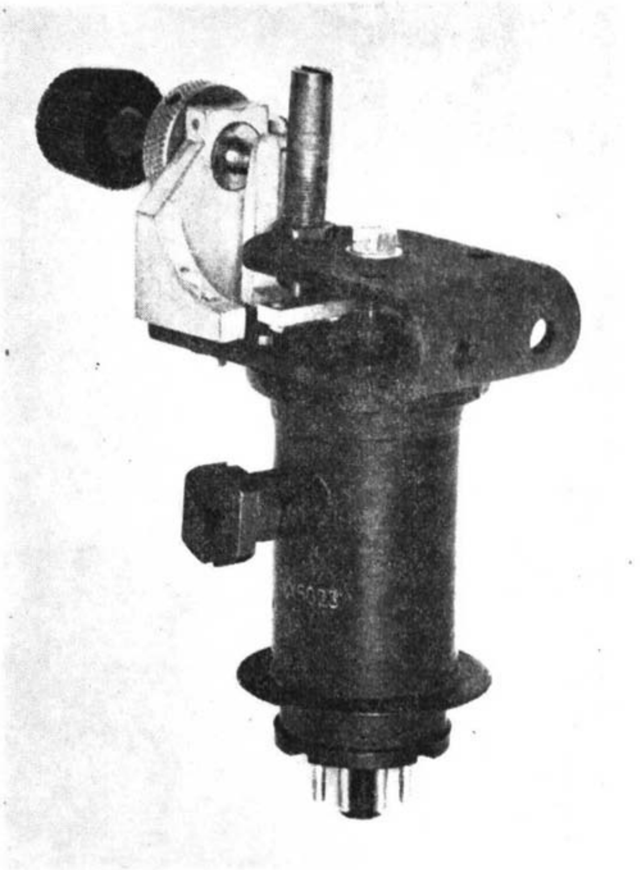


Fig. 4.—Photograph of valve type VX.5023.

The remaining technical problems which arose in the design of the sealed-off valve were concerned with the attachment of the gun and electrical connecting wires to the resonator, and the technique of making the final vacuum seal between the pinch and valve envelope.

The pinch consisted of a seal between a lead-glass disc, through which six dumet leads were sealed, and a nickel-chrome-iron alloy surround having a flat flange to mate with a similar one on the steel envelope. These were joined together by arc welding, and this method was found to be convenient and satisfactory for small numbers of valves.

A solution to the difficulty of attaching the gun and leads to the resonator was found by flexibly mounting the gun on the pinch so that it could be sprung into the nickel dish, shown at H in Fig. 2, the latter being brazed concentrically on to the base of the resonator assembly. The reflector lead from the pinch is at the same time joined by means of a sliding contact to an insulated lead passing through the resonator block. This arrangement is satisfactory since the reflector takes no current.

(3.3) The Mechanical Tuning Mechanism

The natural frequency of the cavity is a very sensitive function of the setting of the resonator gap, a change of 0·001 in giving rise to a frequency shift of 500 Mc/s. It is necessary, therefore, that the external tuning mechanism should provide a large velocity ratio for ease in tuning. This has been obtained by use of a screw-and-lever mechanism in combination with a C spring, the latter being connected to the centre of the top of the valve envelope, which flexes during tuning. The mechanism, which can be seen mounted on top of the valve in Fig. 4, is similar to that originally used on the 1¼cm oscillator (type VX302).

(4) CHARACTERISTICS

Table 1 gives the operating conditions of the valve.

Table 1

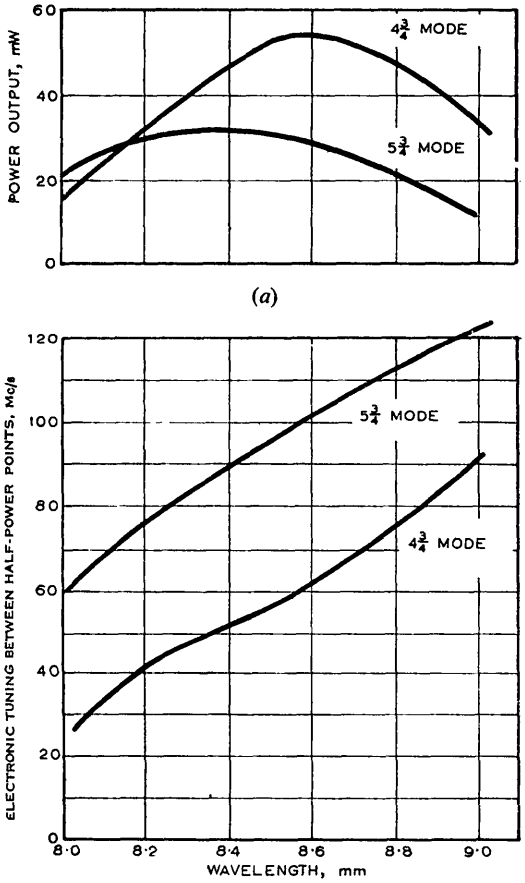
Operating Conditions

Heater voltage	6·3V
Heater current	0·9A
Cathode–resonator voltage ..	2·0kV
Resonator current	8 to 12mA
Screen–cathode voltage range ..	0 to −200V
Reflector–cathode voltage range	−100 to −500V

The resonator current is adjusted for maximum output by means of the cathode-screen voltage.

(4.1) Power Output and Electronic Tuning Range

The variation of power output and electronic tuning range with wavelength is shown in Figs. 5(*a*) and 5(*b*) for two different electronic modes. The curves refer to a valve where the output slot coupling the cavity to the waveguide is 0·060in wide. The performance varies somewhat over the wavelength band, but at the centre a power output of 50mW and an electronic tuning range of 60Mc/s is obtained for the 4¾ mode.



(*a*)

(*b*)

Fig. 5.—Characteristics.

A series of demountable-system experiments was performed in order to determine the output-slot width for optimum power and electronic tuning range. The results are shown in Fig. 6, where curves (*a*), (*b*) and (*c*) show the variation of loaded *Q*, electronic tuning range and power output, respectively, with slot width. It will be seen that the slot size of 0·060in is undercoupled, maximum power and electronic tuning range occurring at a slot width of between 0·080in and 0·090in; the loaded *Q* is then in the neighbourhood of 400. Another consideration that must be taken into account in determining the best slot size is that of noise output, dealt with in more detail in Section 4.2. It is known that noise output increases with slot width (decreasing Q_L), and for some applications an oscillator having the full slot width might generate too much noise. A slot width of 0·060in, and more recently 0·070in, has therefore been used. It may be found possible to increase this to the optimum value when more is known about the cause of the wide fluctuations in noise output from valve to valve mentioned later, and about what level of local oscillator noise can be tolerated in practice in a receiver for the 8–9mm band.

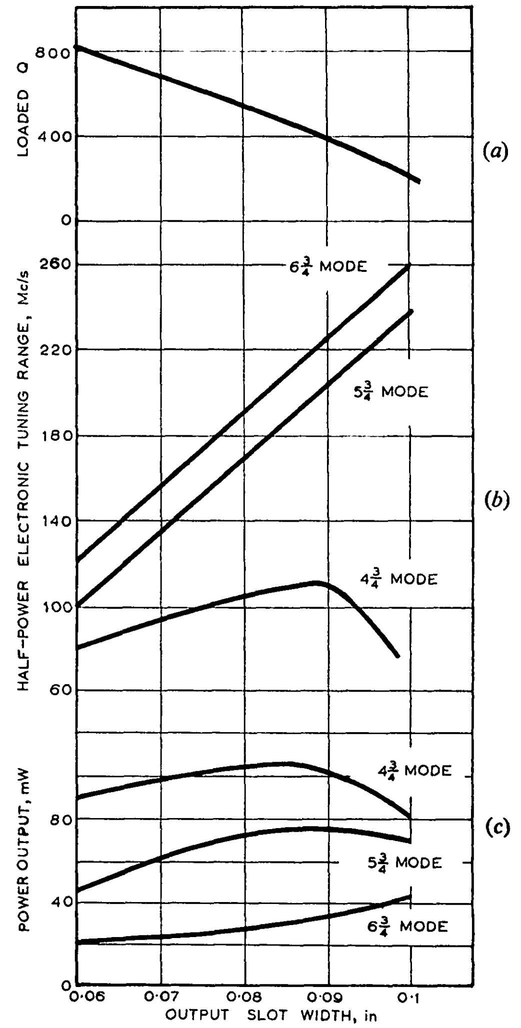


Fig. 6.—Variation of characteristics with cavity loading.

For the measurement of power an enthrakometer bridge[5] was used. Some results on early valves were obtained with a thermistor bridge, but this was replaced after it had been shown[6] that it is subject to an absolute error which becomes appreciable at wavelengths below 3cm.

Table 2

VARIATION OF RECEIVER NOISE FACTOR WITH EXCESS NOISE TEMPERATURE RATIO

t'	0	0·2	0·3	0·5	1	2	3	4	5	10
N(dB)	13·2	13·5	13·6	13·9	14·5	15·4	16·2	16·9	17·5	19·6

The electronic tuning range was measured using a sine-wave modulation on the reflector electrode, since it was found that unreliable results were obtained if frequencies corresponding to the half-power points were measured by manual control of the reflector voltage. This was due to a slow heating effect caused by a small change in resonator current as the reflector voltage was varied, and resulted in a frequency change in addition to that due to normal electronic tuning.

(4.2) Local-Oscillator Noise

The local oscillator in a superheterodyne receiver generates a certain amount of noise. This is of little importance at wavelengths of 3 cm and above but can be significant in the 8–9 mm band. The local-oscillator contribution to receiver noise is conveniently expressed as an increase t' in the noise temperature ratio of a crystal mixer. The overall noise factor of a superheterodyne receiver at centimetric frequencies is given by

$$N = L(t + t' + N_{i.f.} - 1) \quad . \quad . \quad . \quad . \quad (1)$$

All these factors have to be reduced as far as possible for good receiver performance.

The envelope of the noise spectrum generated by the valve is shown in Fig. 7, the two noise sidebands which contribute to receiver noise being spaced at the intermediate frequency from that of the local oscillator. Clearly the magnitude of the noise power depends on the intermediate frequency and the amplifier bandwidth, as well as on the local-oscillator noise spectrum, the latter being dependent on the loaded Q of the resonator.

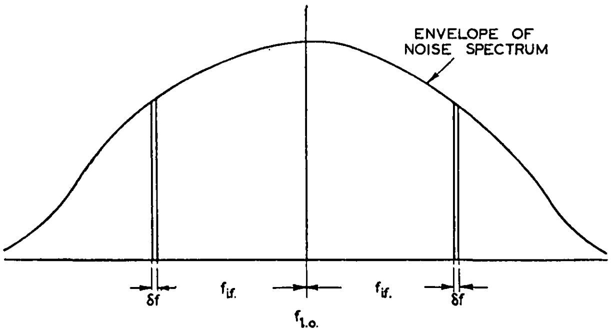


Fig. 7.—The local-oscillator noise spectrum.

From elementary considerations it may be shown that

$$t' = \frac{a}{kT}\frac{P_n}{BP} \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

where a is a constant and k is Boltzmann's constant.

Since P_n is proportional to the bandwidth, t' is a measure of the noise to signal-power output from the local oscillator per unit bandwidth. It will increase as the loaded Q of the cavity is decreased and with decrease of intermediate frequency.

It is instructive to determine how the noise factor of a receiver for use in the 8–9 mm band varies with t' if reasonable values of the other receiver parameters are inserted in eqn. (1). Some values are given in Table 2 for a single crystal mixer making the following assumptions: $L = 8{\cdot}5\,\mathrm{dB}$; $t = 2{\cdot}0$; $N_{i.f.} = 3\,\mathrm{dB}$ (for 45 Mc/s amplifier).

Table 2 shows that the deterioration in receiver performance using a simple mixer would begin to become excessive if t' were allowed to exceed 0·5. The effect of local-oscillator noise can, however, be reduced considerably by the use of a balanced mixer, and if the crystals are perfectly matched, local-oscillator noise is completely cancelled. In practice, however, some mismatch occurs, and so it remains important to maintain as low a value of local-oscillator noise as possible. Just what level can be accepted when using a balanced mixer has not yet been determined, but it is thought that a maximum excess noise temperature ratio of about 10 is permissible. If the match between the crystals is poor, the noise-suppression ratio, which is the amount by which the local-oscillator noise is reduced by use of the balanced mixer, might fall to 20. In these circumstances the deterioration in receiver performance, due to a local oscillator where $t' = 10$, is about 0·7 dB. Normally the deterioration would be considerably less.

No apparatus was available to measure the excess noise temperature ratio at the time of the main development of the valve, but subsequently measurements were made by C. R. Ditchfield of R.R.E., Malvern. Average results for different reflector modes are given in Table 3 for the valve with the 0·060 in output coupling slot.

Table 3

MEASUREMENTS OF EXCESS NOISE TEMPERATURE RATIO

Reflector mode	4¾	5¾	6¾
t' at mode centre	1·7	4·1	10
t' at the low-frequency half-power point	4·3	10·4	>20
t' at the high-frequency half-power point	5·3	16·8	>20

The increase in t' with increasing electronic mode number is due to the fact that the noise power remains sensibly constant with mode, whereas the power output decreases with increasing mode number over the range considered.

Kuper and Knipp (Reference 7, Chapter 17) have discussed the general noise behaviour of the reflex klystron using the shot effect in the injected current as a basis for noise generation. They showed that the asymmetry of the noise with regard to the half-power tuning points may be explained by the change of relative phase between the injected and reflected noise currents as the reflector voltage is varied. This effect is demonstrated in Table 3.

While the general noise behaviour of the VX5023 can be explained in terms of known theory, wide fluctuations in noise output have been observed. For instance, in a batch of valves where the cavity loadings were all approximately the same, it was found that individual values of t' ranged from 1·5 to over 10 for the centre of the 5¾ electronic modes, the mean value being 4·1. The precise explanation for this is not yet known, but experiments now in progress show that the variation is due partly to displacement of the electrodes.

(5) THEORETICAL

The efficiency and electronic tuning range of a reflex klystron may be calculated in a straightforward manner using a small-signal approximation, if the values of several parameters are known. Some of these, however, are known only approximately; moreover, the theory used ignores some effects that may be appreciable, such as the effect of space charge on bunching of the beam. Consequently the numerical values obtained are somewhat approximate; nevertheless the calculations were of considerable assistance in deciding certain features of the design.

The calculations were based on the theory of Barford and Manifold.[2]

(5.1) Efficiency

The efficiency is given in terms of two parameters C and k, where

$$C = 1\cdot68 \times 10^{-6} \times \omega F' s V_0^{-1/2} \quad . \quad . \quad . \quad . \quad (3)$$

and

$$k = \beta^2 RICV_0^{-1} \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

The efficiency is inversely proportional to C, and increases with k from zero at $k = 1$, steeply at first, but flattening off for large values of k (see Reference 2, p. 305). For high efficiency C must be small and k large.

In these equations F', s and V_0 may be obtained from a field plot using an electrolytic trough, although the value of F' found in this way may be to some extent in error because of the neglect of space charge. β can be calculated with fair accuracy,[3] the value of 0·6 being obtained for a resonator gap spacing of 0·006 in. The effective current is another factor which is difficult to assess accurately. It has been taken here as 80% of the total current, i.e. 8 mA.

The shunt resistance of the resonator, which is the resultant of contributions from copper losses and beam damping, is difficult to assess with any accuracy. The part due to beam damping can be calculated, and may be neglected since it is greater than 1 megohm. Published data (Reference 7, p. 78) for a resonator of approximately the shape in use in this oscillator but without apertures gives a shunt resistance due to copper losses of 50 000 ohms. The presence of apertures would perhaps raise this value, but on the other hand the surface conductivity of copper rarely equals the low-frequency value, so it is likely that a somewhat lower value than 50 000 ohms would be obtained in practice.

If, for the moment, the shunt resistance is regarded as a variable, the efficiency calculated from eqns. (3) and (4) can be plotted as a function of R. This has been done in Fig. 8 for the 4¾ and 5¾ modes. Then, if the theoretical efficiency is to agree with the maximum experimental value obtained (0·8% for the 4¾ mode), R must be in the region of 30 000 ohms, which is in reasonable agreement with the estimate made above. Using this value of shunt resistance a value of theoretical efficiency for the 5¾ mode is obtained which is greater than the measured one. This appears to be a general characteristic of the reflex klystron, and has been explained[8] in terms of the increased drift time for the higher mode. This is liable to cause an increased loss of electrons after reflection, and to increase any phase difference that may exist between the axial and the marginal electrons.

(5.2) Electronic Tuning Range

It is well known that the frequency of oscillation f, when the reflector voltage is slightly different from that giving maximum power, is given by

$$f - f_0 = -\frac{f_0}{2Q_L}\tan\phi \quad . \quad . \quad . \quad . \quad . \quad (5)$$

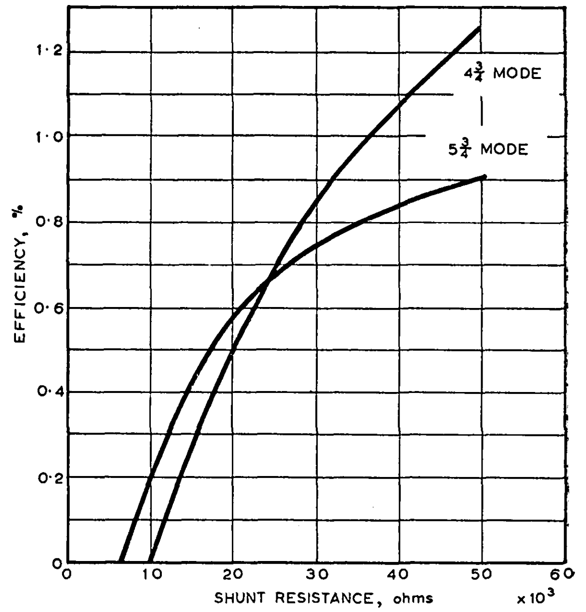


Fig. 8.—Valve efficiency as a function of shunt resistance.

It has been shown[8] that

$$Q_L = \frac{C_0}{1\cdot68 \times 10^{-6} \times \beta^2 IF' s V_0^{-3/2}\dfrac{2J_1(x)}{x}} \quad . \quad . \quad (6)$$

$\dfrac{2J_1(x)}{x}$ is a function depending solely upon k, and can therefore be computed from the known value of k. C_0 may be calculated from

$$Q_0 = \omega_0 C_0 R$$

Hence Q_L may be calculated from eqn. (6), and finally the electronic tuning range from eqn. (5), since $\tan\phi$ is a known function of k.[8]

Table 4

ELECTRONIC TUNING-RANGE CALCULATIONS

Mode	Half-power electronic tuning range		
	$R = 20\,000$	$R = 30\,000$	$R = 50\,000$
	Mc/s	Mc/s	Mc/s
3¾	43	53	54
4¾	72	77	81
5¾	125	130	138

The results given in Table 4 apply to optimum loading conditions and should therefore be compared with experimental values obtained for a slot size of about 0·090 in. It will be seen that the calculated electronic tuning range is not very sensitive to what value is chosen for the shunt resistance over the range 20 000 to 50 000 ohms. Experimental values obtained, namely 110 Mc/s and 200 Mc/s, are rather greater than the computed ones for the 4¾ and 5¾ modes. It is likely that the effect of space charge on some of the parameters involved, such as the bunching factor F', contributes to the discrepancies.

Considerable variation in electronic tuning range is found between valves. Average values are about 60 Mc/s for the 4¾ mode and about 80 Mc/s for the 5¾ mode for an output-slot width of 0·06 in. These values are smaller than the calculated ones because the load coupling is less than optimum.

(6) ACKNOWLEDGMENTS

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