Characteristics of Vacuum Tubes for Radar Intermediate Frequency Amplifiers

By G. T. FORD

1. Introduction

THE desired characteristics for vacuum tubes for use in broad-band intermediate frequency amplifiers are primarily high transconductance, low capacitances, high input resistance, and good noise figure. These characteristics determine the frequency bandwidth, amplification, and signal-to-noise ratio attainable with such an amplifier. The maximum operating frequency is generally limited by the input resistance of the tube which decreases as the frequency is increased, and, in some cases, by the tube noise which increases with increasing frequency. Three other characteristics which are also important are small physical size, low power consumption, and ruggedness. The present paper describes how these characteristics are related to the performance requirements for intermediate frequency (IF) amplifiers used in radar systems and shows how the requirements were met in the design of the Western Electric 6AK5 Vacuum Tube.

In a coaxial cable carrier telephone system of the type which was initially installed between Stevens Point, Wisconsin and Minneapolis, Minnesota[1] the upper frequency of the useful band is of the order of three megacycles per second (mc) and the bandwidth is of the same order. The Western Electric 386A tube was developed a number of years ago for amplifiers such as those used in this system. It is characterized by high transconductance and low capacitances. In radar receiving systems similar but more exacting requirements must be met for the IF amplifier. It is desirable to operate in many cases at a mid-band frequency of 60 mc, to have a bandwidth of the order of 2–10 mc, and to have as close to the ideal noise figure as possible. Additional considerations of great practical importance are low power consumption, small size, and ruggedness.

The choice of the mid-band frequency for the IF amplifier is influenced by considerations, a detailed discussion of which is outside the scope of this paper. For example, the characteristics of the beating oscillator and ite relation to the operation of the automatic frequency control (AFC) system, when AFC is used, are involved. The usual practice has been to

[1] "Stevens Point and Minneapolis Linked by Coaxial System," K. C. Black, *Bell Laboratories Record*, January 1942, pp. 127–132.
"Television Transmission Over Wire Lines," M. E. Strieby and J. F. Wentz, *Bell System Technical Journal*, January 1941, Vol. 20, p. 62.

standardize on a mid-band frequency of 30 mc or 60 mc. The latter frequency has been used in most radars developed at the Bell Telephone Laboratories.

In pulsed radar systems the transmitter is turned on and off by the modulator in such a way that radio frequency energy is generated for a pulse duration τ seconds at a repetition rate which is usually several hundred per second. When the transmitter is modulated by a pulse which approaches that shown in Fig. 1(a), the energy-frequency distribution in the transmitted signal is as shown in Fig. 1(b). Although the maximum of the distribution

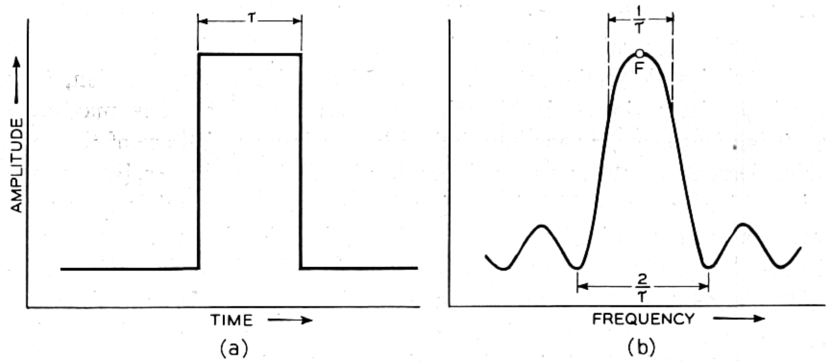

Fig. 1a—Modulating pulse
Fig. 1b—Amplitude—frequency distribution in transmitted RF pulse.

curve lies at the transmitter frequency F, there is considerable energy in the range $F \pm \frac{1}{2\tau}$ and the receiver usually has a bandwidth of at least $\frac{1}{\tau}$ in order to make as efficient use as possible of the energy in the echoes reflected by the target. For many radar applications, this requirement and the problems of transmitter and beating oscillator frequency stability result in the use of a bandwidth of as much as 10 mc for the IF amplifier.

The Western Electric 386A tube, Fig. 2(a), had characteristics which approached those needed to meet the IF amplifier requirements discussed above. It was therefore slightly modified in physical form for convenience of use and recoded the Western Electric 717A tube, Fig. 2(b). The 717A tube was used extensively in IF amplifiers in several radar systems. As the emphasis on small size and light weight for airborne radars increased, and, as the need for better characteristics became more pressing, further development was undertaken which resulted in the Western Electric 6AK5 tube, Fig. 3.

The importance of size and weight for radar systems to be used in airplanes is obvious. The use of miniature tubes in airborne equipment has

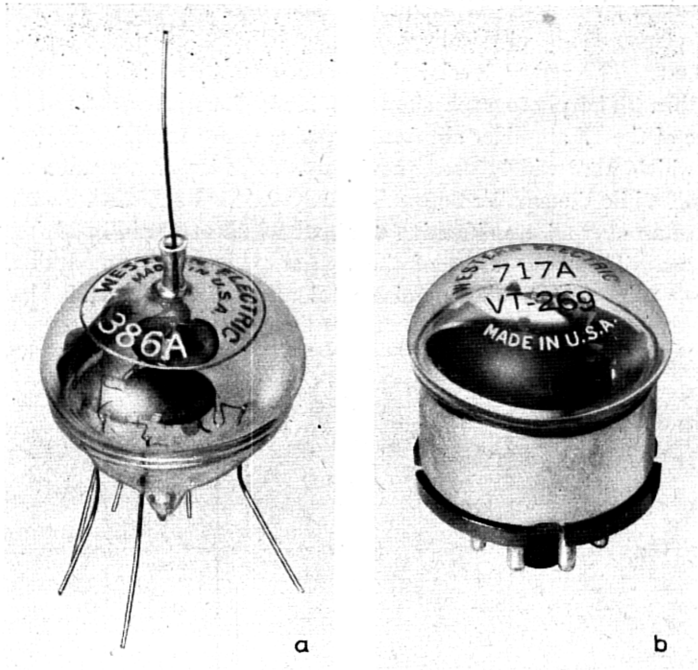

Fig. 2a—Western Electric 386A vacuum tube (full size).
Fig. 2b—Western Electric 717A vacuum tube (full size).

Fig. 3—Western Electric 6AK5 vacuum tube (full size).

been a consequence wherever their power handling capabilities are adequate to meet the performance requirements. The IF amplifier offers an ideal opportunity to effect substantial savings in both size and weight by using small tubes. When the IF frequency is as high as 30 mc or 60 mc, the circuit elements are physically small, so that the tube size is relatively important. The power level is low enough so that miniature tubes are applicable. The photograph shown in Fig. 4 illustrates the reduction in the size of the IF amplifier obtained by using 6AK5 tubes in place of 6AC7 tubes which were widely used previously. The larger amplifier weighs 2 lbs. 4 oz. while the smaller one weighs only 9 oz. Each of these amplifiers provides an over-all amplification of about 95 db at a mid-band frequency of 60 mc. The bandwidths of the two amplifiers are comparable. The amplifier using 6AC7 tubes requires 31.3 watts of power, while the 6AK5

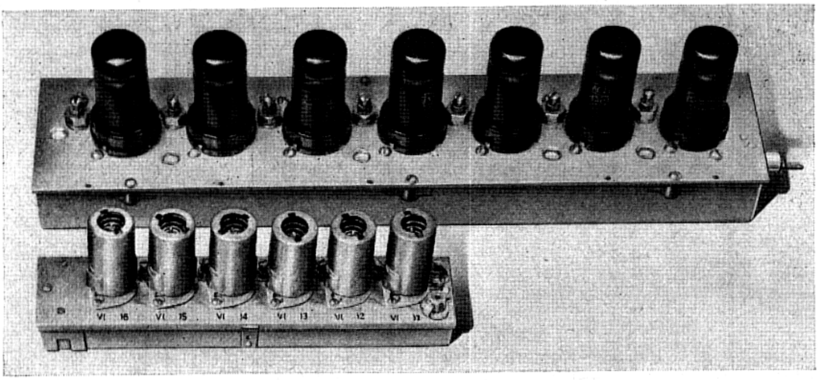

Fig. 4—60 Megacycle IF amplifiers (¼ full size).

amplifier requires 14.4 watts. The power savings possible with the 6AK5 tubes are not particularly important from the power economy standpoint, but rather because of the easier heat dissipation problem. In compact equipment such as airborne radar, the problem of keeping the operating temperatures of the various components within safe limits is formidable.

In the later years of the war the 6AK5 tube became the standard IF amplifier tube for radar systems and because of its superior properties was used for other applications in many other radio equipments.

2. Amplification and Bandwidth

The amplification that can be obtained with a given number of stages, and the useful bandwidth, are closely related. Within certain limits, one can be increased at the expense of the other. In fact, the product of the amplification per stage and the bandwidth is one important measure of the goodness of a particular tube and circuit design. A simple case will illustrate how this comes about. Assume the band-pass interstage shown in Fig. 5. L is the inductance of the coil, R is the shunt resistance equivalent to the load resistor and any loading effects due to high-frequency losses in the tubes or circuit elements, and C is the total shunt capacitance of the tubes, the circuit elements and the wiring. If it is assumed that R, L and C are

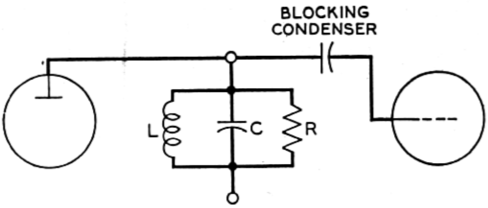

Fig. 5—Band-pass interstage.

independent of frequency, the magnitude of the impedance of this network is

$$|Z| = \frac{1}{\left[\left(\frac{1}{R}\right)^2 + \left(\omega C - \frac{1}{\omega L}\right)^2\right]^{1/2}} \qquad (1)$$

When $\omega = \omega_0 = \frac{1}{\sqrt{LC}}$ the impedance is a maximum and the frequency $f_0 = \frac{\omega_0}{2\pi}$ is the resonant frequency. The maximum value of the impedance is $|Z_0| = R$. If this type of interstage is used in a grounded-cathode type of circuit employing a pentode tube, the small-signal voltage amplification A_v from the control grid of the tube to the plate is

$$A_v = G_m |Z| \qquad (2)$$

where G_m is the grid-plate transconductance of the tube. It is assumed that $|Z|$ is small compared to the plate resistance of the pentode and that there is no feedback. The voltage amplification A_{v0} at the resonant frequency is then

$$A_{v0} = G_m R \qquad (3)$$

If the bandwidth is defined as $\Delta F = f_1 - f_2$, where f_1 and f_2 are the two frequencies where $A_v = \frac{A_{v0}\sqrt{2}}{2}$, it can be shown that

$$\Delta F = f_1 - f_2 = \frac{1}{2\pi RC} \qquad (4)$$

The product of the mid-band voltage gain times the bandwidth can be called the band merit, B_0, and we have

$$B_0 = (\Delta F)(A_{v0}) = \frac{1}{2\pi RC}(G_m R) = \frac{G_m}{2\pi C} \qquad (5)$$

This expression for band merit has been derived from the product of the bandwidth and the voltage amplification for a simple band-pass interstage. Higher band merits can be realized with a given tube by using more complicated coupling networks.

Another interpretation of band merit is to say that it is the frequency at which the voltage amplification is unity. This is the frequency at which the product of the transconductance and the reactance of the shunt capacitance is unity.

From the foregoing expression for band merit it is evident that, in general, the higher the band merit the fewer is the number of stages that are required to obtain a given gain and bandwidth. It is highly desirable to keep the number of stages small in order to save space, weight and power consumption and to avoid the use of unnecessary components which reduce the reliability in operation. In practice the total amplification in the receiver is made high enough so that the system noise, with no signal, produces a fair indication on the output device when the gain control is set for maximum gain. For a bandwidth of 5 mc, the equivalent *RF* input noise power level is of the order of 2×10^{-13} watt and the power level necessary for a suitable oscilloscope presentation in a radar is about 20 milliwatts. The net over-all gain needed is then about 110 *db*. Making an allowance of, say, 15 *db* for losses in the detectors and elsewhere, a total of about 125 *db* gain is required. A minimum of about 110 *db* of this is usually in the *IF* part of the receiver. With this amount of gain as a requirement it is easy to see the importance of high band merit, small size, and low power consumption in the *IF* tubes.

TABLE I

Type	Heater Power	Plate Current	Total Power Consumption	Nominal Transconductance	Transconductance per Unit Plate Current	Band Merit
	watts	*ma*	*watts*	*umhos*	*umhos/ma*	*mc*
6AC7	2.84	10	4.7	9,000	900	89.5
6AG7	4.10	30	9.6	11,000	367	85.3
6AG5	1.89	7.2	3.0	5,100	708	90.0
717A	1.10	7.5	2.3	4,000	533	71.4
6AK5	1.10	7.5	2.3	5,000	667	117.

Table I, above, compares the salient characteristics of the 6AK5 with those of other similar types of tubes.

The tube design factors which determine the band merit will now be examined by considering an idealized case. For an idealized plane parallel triode structure in which edge effects are assumed to be negligible, the plate current can be related to the tube geometry and the applied voltages approximately as follows:[2]

[2] "Fundamentals of Engineering Electronics," W. G. Dow, pp. 44, 102, et seq.

$$I_b = \frac{2.33 \times 10^{-6} A \left(E_{c1} + \frac{E_b}{\mu}\right)^{3/2}}{a^2\left(1 + \frac{1}{\mu}\frac{a+b}{b}\right)^{3/2}} \quad \text{.............(6)}$$

It is assumed that the electrons leave the cathode with zero initial velocities. The failure to take into account the effects of initial velocities makes this equation only a fair approximation for close-spaced tubes, but it is still instructive to assume it is approximately correct for the purposes of the present discussion. "A" is the active area of the structure, "a" is the distance from the cathode to the plane of the grid, "b" is the distance from the plane of the grid to the plate, E_b is the plate voltage, E_{c1} is the effective grid voltage (including the effect of contact potential) and μ is the amplification constant. Dimensions are in cms. This stipulation is unnecessary for equation (6) but is necessary for some of the later equations. For a plane parallel tetrode or pentode, equation (6) is a good approximation if E_b is replaced by the screen voltage E_{c2}, μ is replaced by the "triode mu" μ_{12} (*mu* of control grid with respect to screen grid) and the coefficient M introduced, where M is the ratio of the plate current to the cathode current. The reason this approximation is good is that the field at the cathode, and therefore the cathode current, is determined almost entirely by the potentials of the first two grids. Because of the screening effect of these grids the potential of the plate has little effect on the cathode current. We have then

$$I_b = \frac{2.33 \times 10^{-6} MA \left(E_{c1} + \frac{E_{c2}}{\mu_{12}}\right)^{3/2}}{a^2\left(1 + \frac{1}{\mu_{12}}\frac{a+b}{a}\right)^{3/2}} \quad \text{.............(7)}$$

The distance "b" is now the distance from the plane of the grid to the plane of the screen. This expression can be differentiated with respect to E_{c1} to get

$$G_m = \frac{dI_b}{dE_{c1}} = \frac{3}{2}\,\frac{2.33 \times 10^{-6} MA \left(E_{c1} + \frac{E_{c2}}{\mu_{12}}\right)^{1/2}}{a^2\left(1 + \frac{1}{\mu_{12}}\frac{a+b}{a}\right)^{3/2}} \quad \text{.......(8)}$$

It is assumed that μ_{12} and M are independent of E_{c1}. If we let I_0 be the cathode current density, then substitute MI_0A for I_b in (7), and eliminate the expression $\left(E_{c1} + \frac{E_{c2}}{\mu_{12}}\right)$ between (7) and (8), we have

$$G_m = \frac{3}{2}\,\frac{(2.33 \times 10^{-6})^{2/3} MAI_0^{1/3}}{a^{4/3}\left(1 + \frac{1}{\mu_{12}}\frac{a+b}{a}\right)} \quad \text{...............(9)}$$

Neglecting the stray capacitances between the lead wires and also the edge effects, the greater proportion of the cold capacitance (input plus output) in Farads can be approximated by

$$C_0 = .0885A\left(\frac{1}{a} + \frac{1}{b} + \frac{1}{c}\right) \times 10^{-12} \quad \text{(10)}$$

where "a" and "b" have the same meaning as in (7) and "c" is the spacing between the plate and the suppressor (or screen in the case of a tetrode). This is of course a highly idealized case. The cathode, the grids, and the plate, are each assumed to be plane conductors of infinitesimal thickness, each having an area equal to the active area of the structure.* The band merit then becomes.

$$B_0 = \frac{G_m}{2\pi C_0} = \frac{4.74\ M\ I_0^{1/3} \times 10^8}{a^{4/3}\left(\frac{1}{a} + \frac{1}{b} + \frac{1}{c}\right)\left(1 + \frac{1}{\mu_{12}}\frac{a+b}{a}\right)} \quad \text{(11)}$$

With a given cathode current, the factor M increases as the screen current is reduced. The use of small wires for the screen grid is an important factor in obtaining minimum screen current.

I_0, the useful cathode current density, is limited by the emission capabilities of the cathode. In practice it is necessary to operate in a region considerably below the maximum available emission to avoid excessive changes in transconductance which would result from variations in cathode activity with time. Also, the shot noise will begin to rise when the region of temperature-limited operation is approached. It should be noted that B_0 is independent of the area A.

Taking reasonable values for M, I_0 and μ_{12}, ($M = 0.75$, $I_0 = 50\ ma/cm^2$, $\mu_{12} = 25$), equation (11) becomes

$$B_0 = \frac{1.31 \times 10^8}{a^{4/3}\left(\frac{1}{a} + \frac{1}{b} + \frac{1}{c}\right)\left(1 + .04\ \frac{a+b}{a}\right)} \quad \text{(12)}$$

The curves in Figs. 6, 7 and 8 show how B_0 varies with each of the variables "a", "b" and "c" when a constant value is assigned to the other two. The band merit shown on these curves is considerably greater than that which can be realized in an actual circuit of the simple band-pass type assumed because the stray capacitances in the tube, the socket, the circuit elements, and the wiring increase the total interstage capacitance substantially. Also, the grid-cathode capacitance is substantially higher under normal operating

* This assumption is obviously not true, particularly in the case of the suppressor grid, but, by simply regarding the effective suppressor-plate spacing as somewhat greater than the actual spacing, the assumption becomes useful.

conditions, because of the presence of space charge, than when the tube is cold, as assumed in deriving equation (11). It has been assumed in making

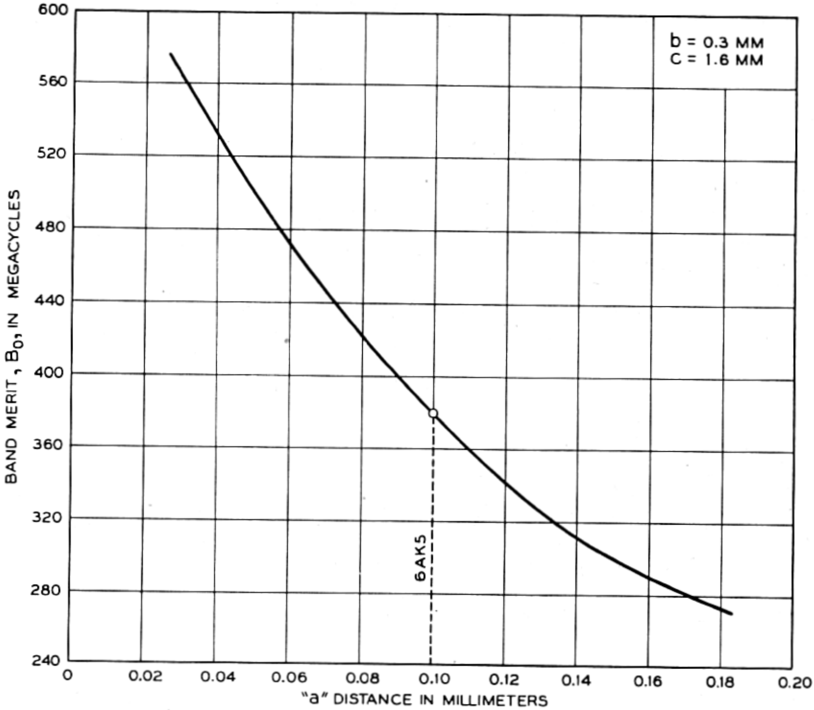

Fig. 6—Band merit vs. grid-cathode spacing.

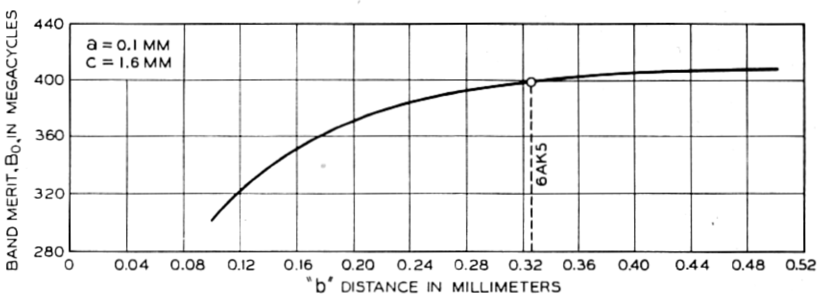

Fig. 7—Band merit vs. grid-screen spacing.

the calculations that the current density and the triode *mu* are held constant while "*a*", "*b*" and "*c*" are varied. In the cases of "*a*" and "*b*", this requires that the control grid and/or screen voltage be varied in such a fashion as to hold I_0 constant. The current density is nearly independent of "*c*" over a reasonable range provided "c" is not so large as to cause the formation of a virtual cathode in the screen-plate space.

Figure 6 shows that B_0 rises as "a" is reduced. Reducing "a" also has the advantage that a lower screen voltage is required with a given grid bias. The improvement in B_0 as "a" is reduced is quite rapid, but of course the mechanical difficulties involved in reducing "a" below about 0.10 mm become very great.

The curve in Fig. 7 shows that B_0 does not increase much if "b" is increased above about 0.30 mm, and increasing "b" has the disadvantage that higher screen voltage for a given grid bias is required.

The curve in Fig. 8 shows that B_0 increases very slowly if "c" is increased beyond about 1.50 mm, and increasing "c" has the disadvantage that the external dimensions of the structure become greater. Increasing "c" also means that the plate comes closer to any shield which is placed around the

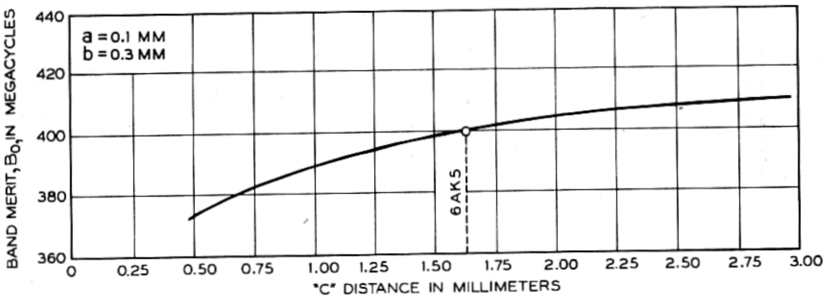

Fig. 8—Band merit vs. plate-suppressor spacing.

bulb of the tube, thus increasing the output capacitance and tending to cancel out any improvement obtained in the structure itself. Also, if "c" is made too large a virtual cathode may be formed in the screen-plate space under some conditions. This would interfere with the normal operation of the tube.

In order to take full advantage of the close grid-cathode spacing, the control-grid pitch should be no greater than about 1.5 times the spacing and the grid wires should be as small as possible. If the pitch is too large, the parts of the cathode directly opposite the grid wires will be cut off while space current is flowing from the sections opposite the spaces between grid wires. This state of affairs shows up in the characteristics as excessive variation of the triode amplification factor, μ_{12}, as the grid bias is varied, and results in a reduction of the transconductance. That is, when the grid is made more negative the amplification factor μ_{12} decreases so that the plate current does not decrease as much as it would if μ_{12} remained constant.

When the grid is made more positive the plate current rise is reduced because μ_{12} increases. The grid wires should be as small as possible so as to block off no more of the area of the structure than is necessary. An ideal grid would be an infinitesimally thin conducting plane which offered no resistance to the passage of electrons through it except that due to its electrostatic potential (sometimes called a "physicist's grid"). In the 6AK5 tube the grid-cathode clearance is 0.089 mm (0.0035 inch), the control-grid pitch is 0.0127 mm (0.0050 inch), and the wire size is 0.0010 inch. Experiments have shown that substantially higher transconductance could have been realized, with the same spacing, if smaller wires and smaller pitch had been used, but the mechanical difficulties would have been much greater.

The way to achieve a high band merit from the tube design standpoint is thus to use as close grid-cathode clearance as practicable, to operate the tube at as high a current density as the emission capabilities of the cathode will permit, and to keep the stray capacitances as low as possible. It was noted above from (11) that B_0 is independent of A. However, if it were possible to maintain the same grid-cathode clearance with a large tube as it is with a small one, the larger tube would have the advantage that the stray capacitances in the tube would be a smaller fraction of the total capacitance so that the band merit for the tube would be higher. It would also be closer to what can be realized when the tube is used in an actual circuit because the capacitances added by the socket, the wiring and the circuit elements would be less important. However, the practical mechanical limitations controlling the minimum grid-cathode clearance have been such that the band merit is roughly independent of the tube size over a moderate range of sizes of high transconductance receiving tubes.

3. Input Conductance

Two factors tend to make the input conductance of tubes higher at high frequencies than at low frequencies. One is the effect of lead inductances and the other is the effect of transit time. If the loading produced by these effects is no more than that required to get the desired bandwidth, it may be no particular disadvantage for stages other than the first one in the amplifier. As will be seen later, however, this effect in the input tube increases the noise figure. The practical result of a consideration of these effects is that the leads are made as short as possible and that small tubes are used in order to use close grid-cathode clearances when the frequency at which the tubes are to be used is above about 10 mc. The expression derived by North[3] for

[3] "Analysis of the Effects of Space Charge on Grid Impedance," D. O. North, I. R. E. Proceedings, Vol. 24, No. 1, January, 1936.

the input conductance of a tetrode or pentode can be re-written, neglecting the higher order terms, to give the approximate expression

$$G_{in} = \frac{5.0 \times 10^{-3} a^2 f^2 G_m}{V_1}\left[1 + 3.3\frac{b}{a\left(1 + \sqrt{\dfrac{V_2}{V_1}}\right)}\right] \quad \ldots (13)$$

where G_m is the triode-connected transconductance, a is grid-cathode spacing in cms, b is grid-screen spacing in cms, V_1 and V_2 are the effective grid-plane and screen-plane potentials in volts, and f is the frequency in mc. It is assumed that there are no lead inductances and that there is no potential minimum in the grid-cathode region. For a given transconductance, (13) shows that close spacings are necessary for minimum grid conductance.

If the lead inductances between the external circuit and the tube elements are appreciable, the input loading may be excessive even though the transit time through the tube structure is negligibly small. The general case taking account of the mutual and self inductances of all the leads of a pentode has been treated by Strutt and van der Ziel[4]. The equations are cumbersome even though only the first order terms in frequency are retained. If all of the lead inductances except that in series with the cathode are neglected, and transit time is assumed to be negligible, the input conductance of a pentode becomes approximately[5]

$$G_{in} = \omega^2 G_m L_k C_k \quad \ldots\ldots (14)$$

where L_k is the cathode lead inductance and C_k is the grid-cathode capacitance. It is further assumed that the plate-grid capacitance is negligible.

Work done at the Naval Research Laboratory includes data on the input conductance of 6AK5 tubes in the frequency range from 100 mc to 300 mc. Through the courtesy of the Naval Research Laboratory some of the data are reproduced in Fig. 9. It is of interest to check a point on this curve against equation (14). For the 6AK5, 0.02 micro-henry is the estimated* cathode lead inductance, $G_m = 5000 \times 10^{-6}$ mhos, and $C_k = 4 \times 10^{-12}$ farad. At a frequency of 250 mc we have a calculated conductance of 990 micromhos, which checks roughly with the value of 1110 micromhos from the curve in Fig. 9. Equation (13) can be used to obtain an approximate value for the loading due to transit time. Taking a = 0.0089 cm., b = 0.032 cm., $G_m = 6.7 \times 10^{-3}$ mhos, f = 250 mc, V_1 = +2.3 volts, and V_2 = +120 volts, a calculation gives G_{in} = 177 micromhos. These results

[4] "The Causes for the Increase of the Admittance of Modern High-Frequency Amplifier Tubes," M. J. O. Strutt and A. van der Ziel, *I. R. E. Proceedings*, Vol. 26, No. 8, August 1938.

[5] "Hyper and Ultra-High Frequency Engineering," Sarbacher and Edson, p. 435.

* This has been checked roughly by Q-meter measurements.

show that the performance of the 6AK5 near the upper end of its present useful frequency range is probably limited to a considerable degree by the lead inductances rather than by transit time effects in the structure.

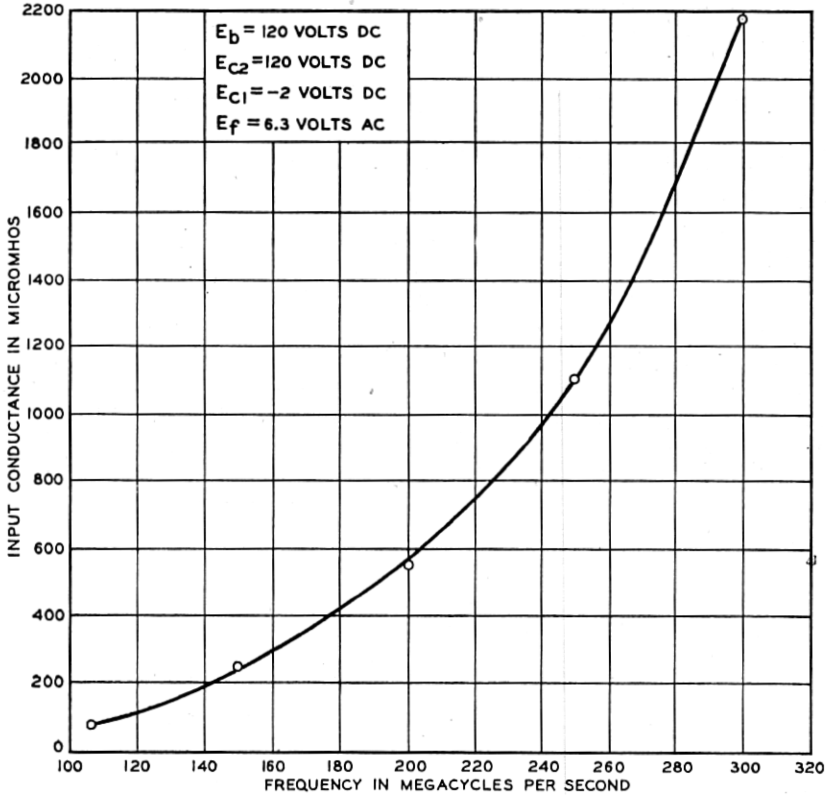

Fig. 9—Average input conductance vs. frequency for six 6AK5 tubes—courtesy of Naval Research Laboratory.

4. Noise

Although it may be possible to employ enough stages of IF amplification to provide the necessary gain and band-width we may still have a relatively insensitive receiver for weak signals. This comes about because there are inherent electrical disturbances in vacuum tubes and passive networks which give rise to random voltages. Since these disturbances may be of the order of the strength of the signal, they must be kept to a minimum in order to maintain a high signal-to-noise ratio. In a receiving system in which no RF amplification is used ahead of the first detector, the signal-to-noise ratio is limited to a large extent by the noisiness of the first detector and the first

IF tube. The noise performance of the first IF stage will be discussed in some detail.**

A convenient method of expressing the departure from ideal performance is the use of the "noise figure" proposed by Friis.[6] The noise figure of a network or amplifier may be defined as follows:

$$NF = \frac{\text{Available output noise power}}{GKT\Delta f} \qquad (15)$$

where G is the "available gain" which is defined as the ratio of the available signal power at the output terminals of the network or amplifier to the available signal power at the terminals of the signal generator. $KT\Delta f$ is the available noise power from a passive resistance, where K is Boltzmann's constant, T is absolute temperature and Δf is the incremental bandwidth. This follows from consideration of a noise generator of resistance R_0 working into a load resistance R_0. This is the condition for maximum power into the load, half of the noise voltage appearing across the source and half across the load. The open-circuit noise voltage appearing across the terminals of a resistance R_0 is

$$V^2 = 4KTR_0\Delta f \qquad (16)$$

The noise power delivered to the load by the source resistance will be

$$P_n = \frac{V^2}{4R_0} = KT\Delta f \qquad (17)$$

If there were no source of noise other than that of the resistance of the signal source itself, the noise figure would be unity. If the signal source works directly into a matched load resistance at the same temperature, the noise figure is 2 since the available output noise power is $KT\Delta f$ and the available gain is one-half.

The importance of the noise figure of the radar receiver is obvious since a reduction in the noise figure is equivalent to the same percentage increase in transmitter power. In recent radar systems the noise arising in the first IF stage constituted a substantial part of the total receiver noise.

If the first IF tube provides at least 15 db gain, noise introduced by its plate load impedance, and by anyother sources in the rest of the amplifier, is usually negligible. The departure of the noise figure of the IF amplifier from unity is then due to noise arising in the first tube and in its input circuit.

In the very high frequency (VHF) range, essentially all of the noise arising

** It will be assumed in all of the discussion about noise that there is no noise due to flicker effect, emission of positive ions, microphonics, sputter, ionization, secondary emission, or reflection ot electrons from charged insulators.

[6] "Noise Figure of Radio Receivers," H. T. Friis, *I. R. E. Proceedings*, July 1944.

in the tube itself is due to random fluctuations in the emission of electrons from the cathode. For a parallel plane diode in a circuit such as that shown in Fig. 10(*a*), the mean square noise current is[7]

$$\overline{i_k^2} = 2\Gamma^2 e I_k \Delta f \qquad (18)$$

where e is the electronic charge, I_k is the d-c cathode current, Δf is the incremental bandwidth, and Γ is a factor which takes into account the

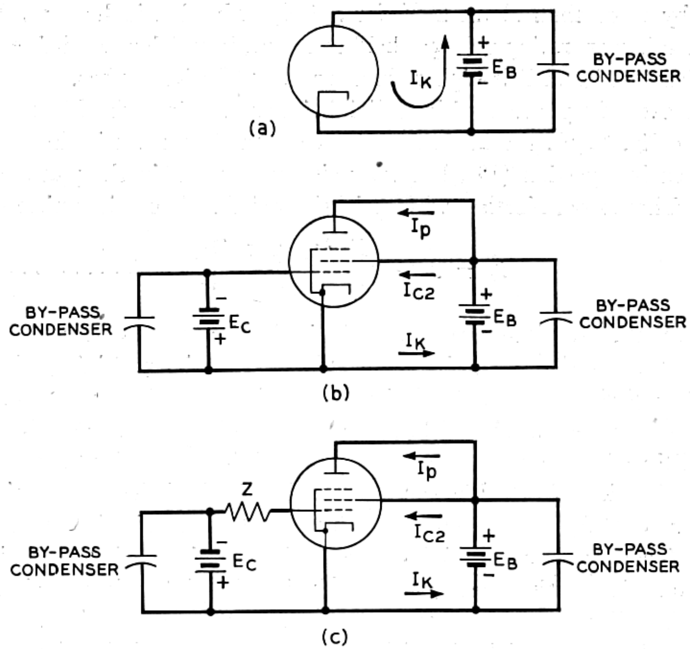

Fig. 10a—Diode, negligible circuit impedance.
Fig. 10b—Pentode, negligible circuit impedance.
Fig. 10c—Pentode, impedance in grid circuit.

"cushioning effect" of space charge. For anode-cathode spacings and operating conditions such that the transit time is not too large a faction of the period of the frequency involved, Γ is of the order of 0.20 when the zero-field emission is several times larger than I_k. Under temperature-limited conditions Γ is unity. That is, under favorable space-charge-limited conditions, the fluctuation noise component in I_k is only about 20% of the value for the same cathode current under temperature-limited conditions. Although the introduction of grids between the cathode and anode

[7] "Fluctuations in Space-Charge-Limited Currents at Moderately High Frequencies," B. J. Thompson, D. O. North, W. A. Harris, *R. C. A. Review*, October 1940, Vol. 5, p. 244.

of the simple diode complicates the noise problem, the factor Γ remains of great importance. It is therefore desirable to control the tube processing in manufacture so as to insure adequate available emission in every tube. An "activity test" is made on completed tubes for this purpose. This test consists of reducing the heater voltage by an arbitrary amount (usually 10%) and observing the change in one of the tube characteristics which is sensitive to changes in available emission. The characteristic used for tubes like the 717A and 6AK5 is the transconductance. If the transconductance decreases by more than about 20% for a 10% reduction in heater voltage, with the other operating voltages held constant, insufficient available emission is indicated and Γ is higher than for more "active" tubes.

In the case of a pentode with negligible impedance in each of its leads, as shown in Fig. 10(b), the fluctuation in the cathode current is the same as that given in equation (18), but the noise component in the plate lead is larger. Thompson, North, and Harris[7] showed that it can be written as

$$\overline{i_p^2} = 2eI_p\Delta f\left[\frac{\Gamma^2 I_p + I_{c2}}{I_k}\right] \qquad (19)$$

where I_p is the d-c plate current and I_{c2} is the d-c screen current. It was mentioned above that Γ can be made as low as 0.20 by providing adequate available emission. From the design standpoint, equation (19) also shows that the screen current should be as small as possible. For normal operating conditions in a pentode, the screen current is influenced by the screening fraction (fraction of area blocked off by grid wires) of the screen grid and by the amount of space current turned back to the screen in the screen-plate region. The way to get a minimum screening fraction and still obtain the desired function of the screen grid of reducing the plate-grid capacitance is to use wire of as small diameter as possible. The presence of a suppressor grid, at cathode potential, placed between the screen and the plate to prevent interchange of secondary electrons, causes a certain proportion of the space current which would otherwise go to the plate to be turned back to the screen. Here again, this effect is minimized by using as fine wire as possible for the suppressor grid. It was pointed out in an earlier section that fine wires are desired for the control grid. The ideal for each of the three grids in an IF pentode would be a conducting plane which offers no resistance to the passage of electrons other than the influence of its potential.

When impedances are connected in the various leads of a pentode the noise components discussed above will, in general, be different due to the influence of fluctuation voltages developed between the elements of the tube. In particular it is found experimentally in the VHF range of frequencies that when an impedance Z is introduced between the grid and cathode, as shown in Fig. 10(c), the noise component in the plate lead rises more than would be expected due to thermal noise from Z, because of the effect of grid noise. North and Ferris[8] showed that the grid noise can be taken into account by assuming that the input resistance of the tube is a resistance noise source whose absolute temperature is about 4.8 times ambient, if the input loading is due to transit time effects alone. One consequence of this input loading is that at high frequencies the best signal-to-noise ratio is usually obtained with an input circuit of lower impedance than that which would be used at low frequencies.

Actually, as was brought out in an earlier section, the loading in tubes like the 6AK5 is probably due largely to lead inductances between the active tube elements and the external circuit components up to a few hundred megacycles. According to Pierce[9] the effect of the cathode lead inductance feedback is to reduce the signal component in the output current while leaving the noise current due to screen interception noise unaffected. Input loading may be a limiting factor in tubes like the 717A and 6AK5 when the frequency is of the order of 100 mc or higher, both because of its effect on gain in some cases and because of its adverse effect on the signal-to-noise ratio for early stage use. There is good evidence that the 6AK5 structure would be useful at much higher frequencies than is the case at present if the circuit connections to the tube elements were improved by more advantageous mounting of the structure and the use of more suitable sockets or external connectors.

Noise measurements made by a number of workers at Bell Telephone Laboratories[10] indicate that an average noise figure of about 2.8 can be obtained with the 6AK5 at a midband frequency of 60 mc, with a well-designed input circuit, and bandwidths up to 10 mc. At a mid-band frequency of 30 mc the noise figure is about 2.4. At 100 mc it is about 3.6.

5. Description of the Design of the Western Electric 6AK5 Tube

In order that the reader may have a full appreciation of the dimensions and other requirements of design to meet the characteristics discussed above, a detailed description of the 6AK5 tube follows.

5.1 *Mechanical Description*

The 6AK5 tube is an indirectly heated cathode type pentode employing the 7-pin button stem and the T-5-1/2 size miniature bulb. The outline dimensions are shown in Fig. 11. A photograph of a mount ready to be sealed into a bulb is shown in Fig. 12. Figure 13 is a photograph of a transverse section through the tube at the middle of the structure in a plane

[8] "Fluctuations Induced in Vacuum Tube Grids at High Frequencies," D. O. North and W. R. Ferris, *I. R. E. Proceedings*, Vol. 29, No. 2, February 1941.
[9] Unpublished Technical Memorandum, J. R. Pierce.
[10] S. E. Miller, V. C. Rideout, R. S. Julian.

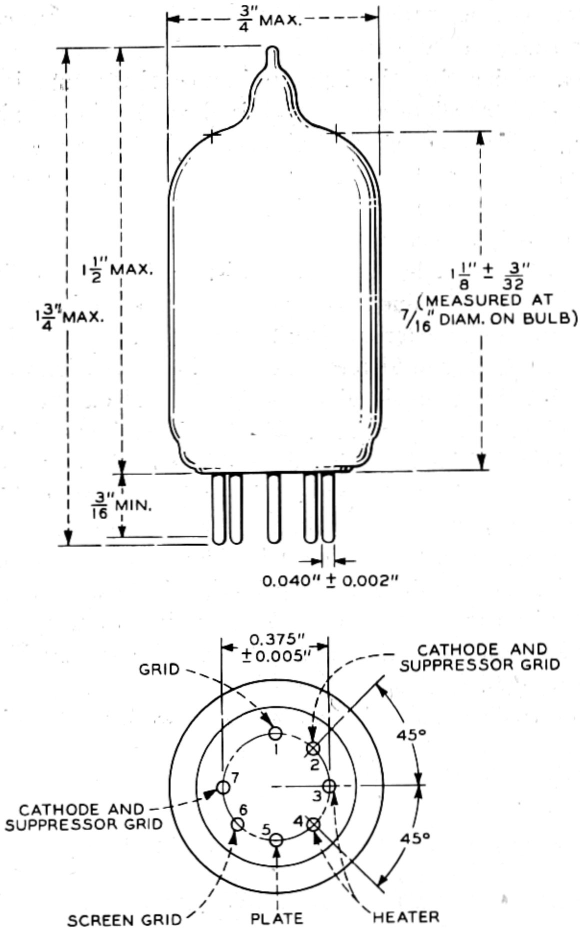

Fig. 11—6AK5 outline dimensions.

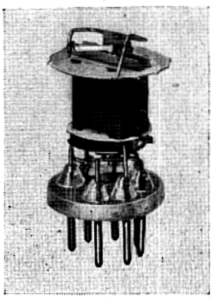

Fig. 12—6AK5 mount structure (full size).

parallel to the stem or "base" of the tube. The magnification in Fig. 13 as reproduced is about 5½ times.

The heater is a conventional folded type with eight legs of coated tungsten wire. The wire diameter is 0.0014 inches and the unfolded length is 3 inches. The insulating coating consists of fired aluminum oxide and is about 0.0025 inch thick.

The cathode is of oval cross-section with the contours of the longer sides shaped to conform to the shape of the control grid. The major axis of the

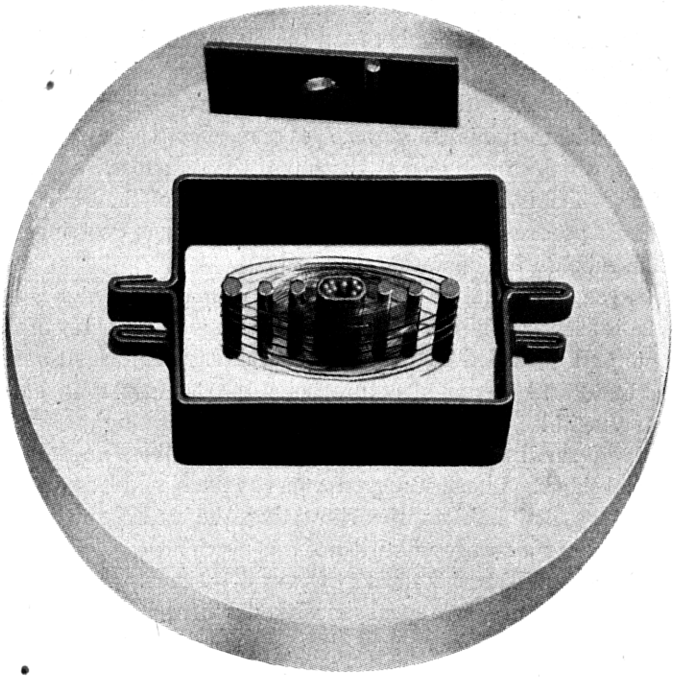

Fig. 13—Transverse section of 6AK5 tube (5½ times).

cross-section of the cathode sleeve is 0.048 inch. The minor axis is 0.025 inch. The length of the sleeve is 0.47 inch and it is coated over a centralized section which extends 0.28 inch along its length. The coating is the usual mixture of oxides of barium, calcium, and strontium. Before the tube is processed, the coating thickness is about 0.002 inch. After processing, it is about 0.001 inch thick.

The grids are oval-shaped and are wound with small diameter wires. The control-grid and the screen-grid lateral wires are 0.001 inch diameter tungsten. The suppressor-grid lateral wires are 0.002 inch diameter molybdenum. The control-grid wires are gold-plated in order to minimize primary emission of electrons from this grid.

The plate has a rectangular transverse cross-section and is made of 0.005 inch thickness carbonized nickel. The carbonization increases the thermal emissivity of the plate surface so that a reasonable amount of power dissipation in the plate can be tolerated.

The end shields are made of nickel-plated iron. The reason for using this material instead of nickel, as is more often the case, is to prevent overheating of these shields during the exhaust process. At the temperatures used, the iron shields pick up less energy in the induction field during the out-gassing of the plate than do nickel shields. The function of the end shields is to minimize the stray capacitance between the plate and the control grid.

The insulators or spacers which hold the tube elements in proper disposition with respect to each other are of high grade mica. They are coated with magnesium oxide to minimize surface leakage effects.

The getter, which can be seen at the top of the structure in Fig. 12, contains barium which is flashed onto the inside surface of the bulb at the end of the exhaust process in order to take up residual gas evolving from the parts of the tube during operation.

Although the individual parts are extremely small and fragile, the completed tube is surprisingly rugged. The short supporting wires in the stem and the support provided by the bulb-contacting top insulator result in the stem, bulb, and mount structure being a relatively rigid unit. The small parts assembled into the mount are very light in weight and therefore exert relatively small forces on their supporting members under conditions of mechanical shock. Shock tests performed at the Naval Research Laboratory and at the Bell Laboratories show that the 6AK5 tube stands up satisfactorily under a steady vibration of rms acceleration 2.5 times gravity and withstands 1 millisecond shocks of over 300 times gravity.

The most important single geometrical factor in the tube is the spacing between the cathode and the control grid. In the 6AK5 tube this clearance is 0.0035 inch after processing. Before processing it is 0.0025 inch. The manufacturing difficulties involved in assembling the structure and maintaining such small clearances are obviously very great. However, as was seen from the discussion above, this close spacing is essential in order to obtain the desired high-frequency performance.

It can be observed that the close mounting of the structure on the stem provides very short lead lengths between the tube elements and the external pins. This is of importance at frequencies where the inductances of the lead wires become comparable in magnitude to the circuit reactances.

5.2 *Low Frequency Electrical Charatceristics*

The usual static characteristics are shown by the curves in Fig. 14. The ratings, nominal characteristics, and cold capacitances are given in Table II

5.3 *Fixed Tuning*

It is highly desirable to design the IF amplifier with fixed tuning in order to minimize the number of adjustments that need to be made when tubes

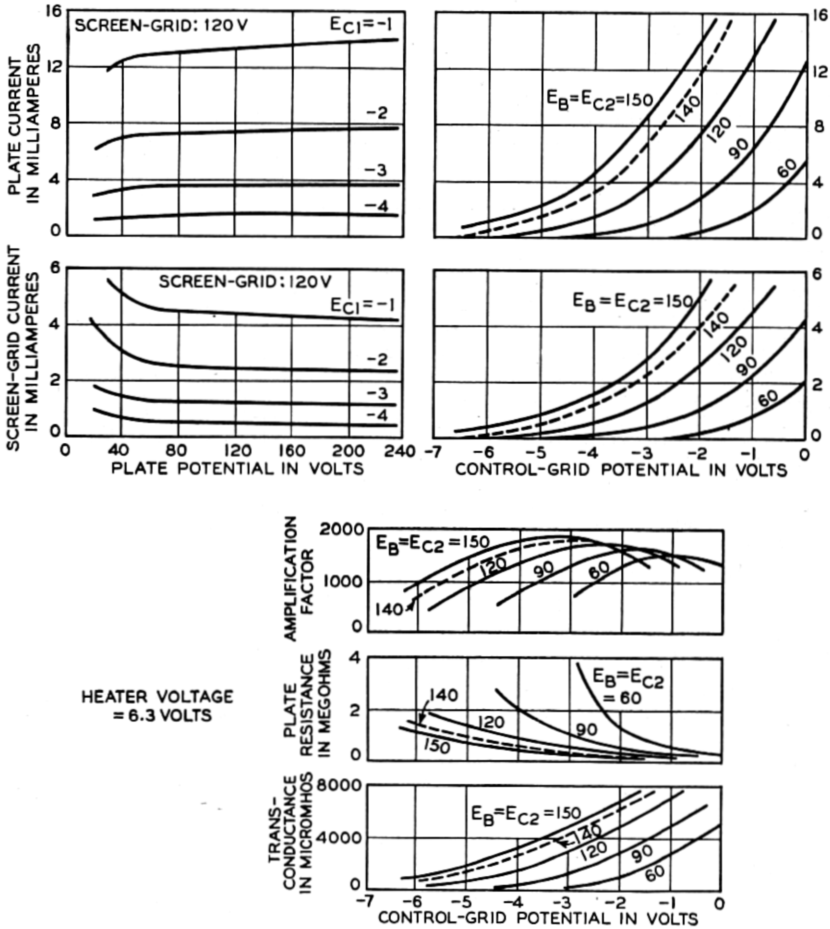

Fig. 14—Average 6AK5 characteristics.

are changed. The difficulty immediately arises that, since the tube capacitances are usually a large fraction of the total shunt capacitance of each coupling network, the variations in capacitance from one tube to another become very important. It is essential that, after the amplifier is lined up at the factory with standard tubes, any stock tubes taken at random shall give satisfactory performance in the amplifier. This requires close control of the capacitances. Not only must the capacitances be held within the limits of the test specification, but the product averages must be kept close to the design center values particularly in the cases of the input and output capacitances.

TABLE II

HEATER RATING			
Heater voltage			6.3 volts, a-c or d-c
Nominal heater current			0.175 ampere
Maximum Ratings (Design-center values)			
Maximum plate voltage			180 volts
Maximum screen voltage			140 volts
Maximum plate dissipation			1.7 watts
Maximum screen dissipation			0.5 watt
Maximum cathode current			18 milliamperes
Maximum heater-cathode voltage			90 volts
Maximum bulb temperature			120°C
OPERATING CONDITIONS AND CHARACTERISTICS			
Plate voltage	120	150	180 volts
Screen voltage	120	140	120 volts
Cathode-bias resistor	200	330	200 ohms
Plate current	7.5	7.0	7.7 milliamperes
Screen current	2.5	2.2	2.4 milliamperes
Amplification factor	1700	1800	3500
Plate resistance	0.34	0.42	0.69 megohms
Transconductance	5000	4300	5100 micromhos
INTERELECTRODE CAPACITANCES (*With JAN 1A No. 314 shield connected to cathode*)			
Control grid to heater, cathode, screen grid and suppressor grid			3.90 $\mu\mu f$
Plate to control grid			0.01 $\mu\mu f$
Plate to heater, cathode, screen grid and suppressor grid			2.85 $\mu\mu f$

6. CONCLUSION

The important factors to be considered in evaluating the suitability of a vacuum tube for broad band IF use in the VHF range are as follows:

1. Band merit
2. Noise figure
3. Input conductance
4. Power consumption
5. Physical size
6. Control of capacitances

We have seen that the tube design features which make a tube good on the basis of these requirements are close grid-cathode spacing, fine grid wires, short lead wires, and small elements. An important consideration from the manufacturing standpoint is the control of cathode emission for low noise. It has been possible to extend the useful frequency range of conventional receiving tubes up through the VHF range and somewhat higher by this line of attack.

The 6AK5 tube is an outgrowth of the many years of development in this general field by the Electronics group of the Bell Telephone Laboratories. Contributions to the success of the development have been made by chemists, physicists, mechanical engineers and electrical engineers too numerous to mention individually and to whom the author is indebted for much of the material presented.