ELECTRON TUBES

RCA Application Note

AN - 118
April 15, 1947

Input Admittance of Receiving Tubes

This Note gives values of short-circuit input conductance and capacitance for a number of rf pentodes at frequencies of 50 to 125 megacycles. It also describes how these values were measured and describes their application in design problems.

THEORETICAL CONSIDERATIONS

In determining tube admittances at frequencies higher than approximately 10 megacycles, it is not practicable to introduce voltages or measure them directly at the electrodes of a tube. The lead inductances and inter-electrode capacitances form a network too complex for exact analysis. The most practical method of avoiding such difficulties is to consider the tube, the socket, and the associated bypass or filter circuits as a unit, and to select a pair of accessible input terminals and a pair of accessible output terminals as points of reference for measurements. When such a unit is considered as a linear amplifier, it is possible to calculate performance in terms of four admittance coefficients. These are:

(1) The short-circuit input admittance, measured at the input terminals with the output terminals shorted for the signal frequency;

(2) The short-circuit output admittance, measured at the output terminals with the input terminals shorted for the signal frequency;

(3) The short-circuit forward admittance, which is the quotient of current at the output terminals divided by the voltage between the input terminals, with the output terminals shorted for the signal frequency; and,

(4) The short-circuit feedback admittance, which is the quotient

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			Single-Ended Metal Types						Miniature Types					
			6SJ7	6SK7	6SH7	6SG7	6AB7	6AC7	9001	9003	6AU6	6BA6	6AG5	6AK5
1.	Plate Voltage	Volts	250	250	250	250	300	300	250	250	250	250	250	120
2.	Screen Voltage	Volts	100	100	150	125	200	150	100	100	150	100	150	120
3.	Grid Voltage	Volts	-3.2	-2.8	-1.0	-1.0	-2.8	-2.2	-2.9	-2.9	-1.2	-1.3	-1.8	-2.0
4.	Plate Current	Ma.	3.0	9.2	12.2	11.8	12.5	10.0	2.0	6.7	10.8	11.0	7.4	7.5
5.	Screen Current	Ma.	0.9	2.7	4.2	4.5	3.1	2.4	0.8	2.5	4.4	4.4	2.4	2.5
6.	Transconductance	μmhos	1490	1980	5500	4950	4700	9450	1450	1900	5250	4300	4950	4950
SHORT-CIRCUIT INPUT CAPACITANCE:														
7.	Tube Operating as in lines 1 to 6	μμf	9.5	9.4	13.9	13.6	12.4	18.0	5.5	5.0	10.0	9.6	9.3	6.4
8.	Tube Cutoff	μμf	8.5	8.2	11.6	11.3	10.6	15.6	5.0	4.5	7.5	7.4	7.9	5.3
9.	Tube Cold	μμf	8.1	8.1	11.0	11.0	10.2	14.5	4.9	4.4	7.1	6.9	7.4	5.0
10.	Capacitance Increase (Cold to Cutoff)	μμf	0.4	0.1	0.6	0.3	0.4	1.1	0.1	0.1	0.4	0.5	0.5	0.3
11.	Capacitance Increase (Cutoff to Operating)	μμf	1.0	1.2	2.3	2.3	1.8	2.4	0.5	0.5	2.5	2.2	1.4	1.1
SHORT-CIRCUIT INPUT CONDUCTANCE:														
12.	Tube Operating as in lines 1 to 6	μmhos	528	503	632	604	792	1970	61.7	66.0	759	603	326	134
13.	Tube Cutoff	μmhos	101	72	98	99	112	240	11.6	9.4	25	28	40	13
14.	Tube Cold	μmhos	66	66	95	96	86	162	11.1	8.4	16	21	28	11
15.	Conductance Increase (Cold to Cutoff)	μmhos	35	6	3	3	26	78	0.5	1.0	9	7	12	2
16.	Conductance Increase (Cutoff to Operating)	μmhos	427	431	534	505	680	1730	50.1	56.6	734	575	286	121
17.	Socket Capacitance	μμf			1.6							0.8		
18.	Socket Conductance	μmhos			26.6							2.3		
19.	Grid-to-Cathode Capacitance*	μμf	2.14	2.01	3.59	3.42	3.15	5.26	1.64	1.31	3.10	3.02	3.35	2.31

* Measured with tubes cold at low frequency.

Table I - Short-Circuit Input Admittance Data at 100 Megacycles of current at the input terminals divided by the voltage between the output terminals, with the input terminals shorted for the signal frequency.

Each of these admittances can be considered as the sum of a real conductance component and an imaginary susceptance component. In the cases of the input and output admittances, the susceptance components are nearly always positive (unless the tube is used above its resonant frequency) and it is, therefore, common practice to present the susceptance data in terms of equivalent capacitance values. The short-circuit input capacitance is the quotient of the short-circuit input susceptance divided by 2π times the frequency. The capacitance values are more convenient to work with than the susceptance values because they vary less rapidly with frequency and because they are directly additive to the capacitances used in the circuits ordinarily connected to the input and output terminals. However, when frequencies higher than those considered in this Note and resonant lines used as tuning elements are involved, the use of susceptance values may be preferable.

An equivalent circuit for the system discussed can be drawn as follows:

Represent the short-circuit input admittance by a resistor and capacitor in parallel across the input terminals. The resistor is equal to the reciprocal of the short-circuit input conductance and the capacitor is equal to the short-circuit input capacitance. Represent the short-circuit output admittance by a similar combination across the output terminals. Draw a constant-current generator at the output terminals producing a current equal to the product of the short-circuit forward admittance and the input voltage. Draw a similar generator at the input terminals producing a current equal to the product of the short-circuit feedback admittance and the output voltage. This circuit differs in one respect from the equivalent circuit usually drawn at low frequencies: the low-frequency circuit generally shows the feedback capacitance as a passive element between input and output terminals with a single constant-current generator representing the transconductance. The latter circuit simplifies the drawing but complicates the algebra.

The principal differences in the performance of receiving tubes at high and low frequencies can be attributed to the variations of the short-circuit input conductance with frequency. The other short-circuit admittance coefficients, however, contribute to the input admittance actually observed in an operating circuit as follows:

The voltage gain from the input to the output terminals is the ratio of the short-circuit forward admittance to the sum of the short-circuit output admittance and the admittance of the load connected between the output terminals.

The added current at the input terminals due to the presence of the load is the product of the input voltage, the voltage gain, and the short-circuit feedback admittance.

The total input admittance is therefore the sum of the short-circuit input admittance and the product of voltage gain and short-circuit feedback admittance. The phase angle of the added component is the sum of the phase angle for the voltage gain and the phase angle for the feedback admittance.

MEASUREMENT OF SHORT-CIRCUIT INPUT ADMITTANCE

The circuit diagram for the equipment used in obtaining high-frequency admittance data is shown in Fig. 1. Fig. 2 is a cutaway view showing the physical arrangement of the circuit elements. The tube under test is used as a part of a resonant circuit which includes a continuously-variable inductor and a small concentric-cylinder capacitor built on a micrometer head. The high-potential end of the inductor is connected to the high-potential electrode of the micrometer capacitor inside a cylindrical cavity open at the top. Fig. 2 also shows the positions of some of the bypass capacitors used with the octal socket. These are button-type, silver-mica capacitors of approximately 500-μμf capacitance. The socket is of the molded phenolic type. Terminals 1, 3, 5, and 7 are connected directly to the mounting plate at a point directly below the terminal in each instance. Terminals 2 (heater) and 6 (screen grid) are bypassed to ground, and a lead is brought from each of these terminals through the mounting plate to a small rf choke and a second bypass capacitor. Terminal 8 (plate) is bypassed and fitted with a spring contacting the terminal for one of the circuits. Terminal 4 (grid) has only the contact spring. A similar arrangement is used with

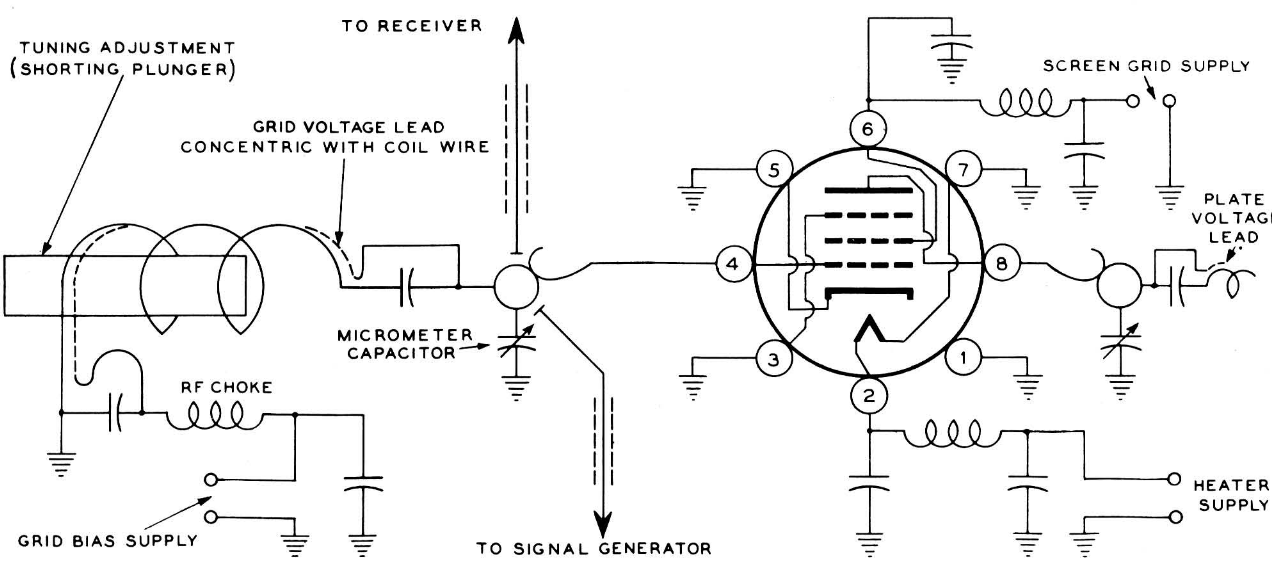

Fig. 1 - Circuit Diagram of Equipment Used for Measurement of Short-Circuit Input Admittance

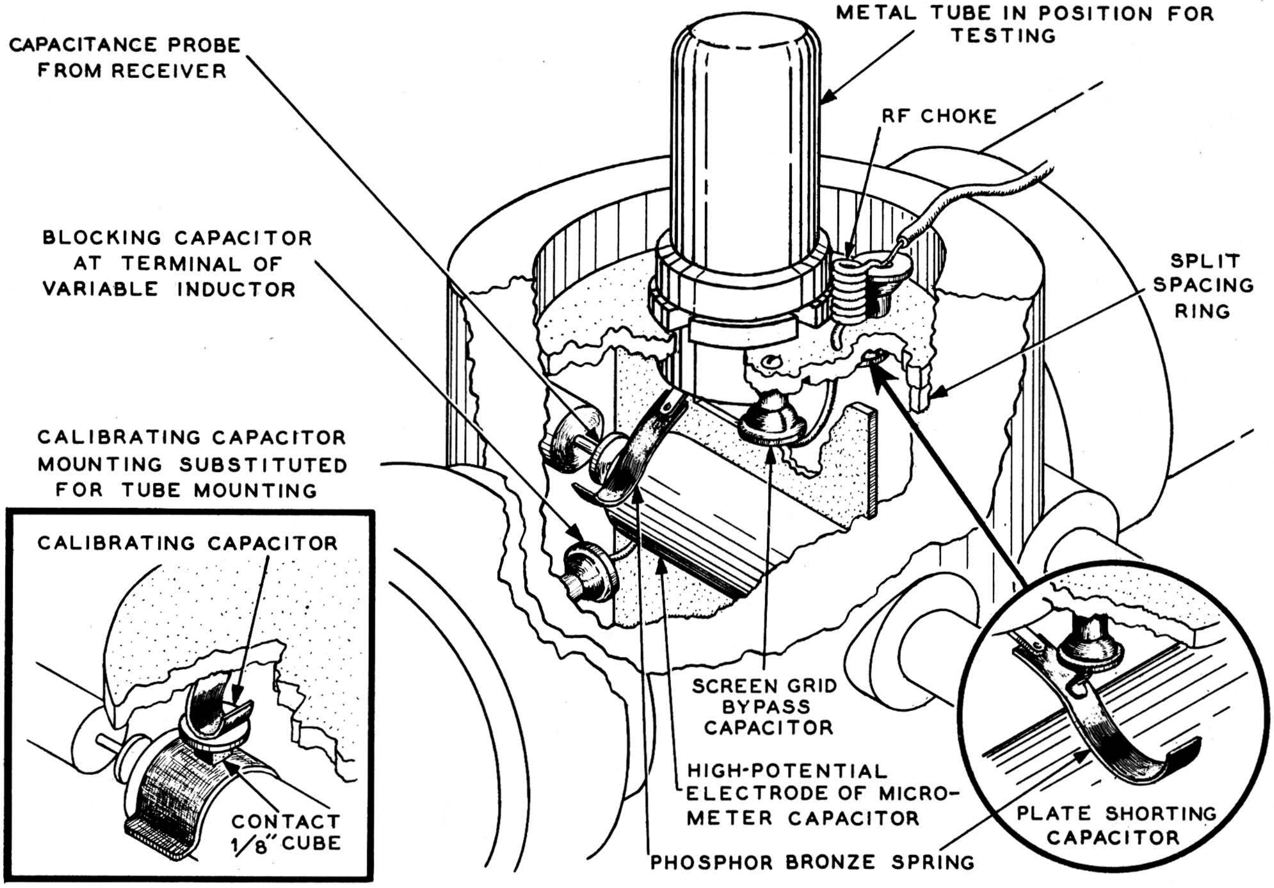

Fig. 2 - Physical Arrangement of Circuit Elements

miniature tubes. In this case, the socket is of the wafer type with mica-filled rubber insulation. Terminals 2, 3, and 7 are grounded and terminals 4, 5, and 6 are bypassed to ground. Terminals 1 (grid) and 5 (plate) have contact springs. The bypass capacitors are closer to the mounting plate than in the case of the octal socket; the capacitors at terminals 4 and 6 overlap the capacitor to terminal 5. A 10-ohm resistor, mounted inside a cylindrical shield to minimize lead inductance, is connected between socket terminal 5 (plate) and the bypass capacitor. This component was added to suppress a parasitic oscillation observed with certain type 6AK5 tubes. The resistor was found to have no measurable effect on input-admittance readings obtained with tubes, either of the 6AK5 type or of other types, not subject to the parasitic oscillation.

In order to obtain susceptance values, the circuit must first be calibrated for the capacitance required for resonance at each test frequency. The circuit is calibrated by determining the inductor settings for resonance with each of a number of small, disc-shaped, calibrating capacitors substituted for the tube. The insert in the lower left corner of Fig. 2 shows a cutaway view of the cavity with one of the calibrating capacitors in place. The length of the phosphor-bronze contact spring used with the calibrating capacitors is approximately the same as that used with the tube. Thus, the inductance of this lead is accounted for in the calibrating procedure. The reference terminals for the tube are the socket plate and the grid terminal of the socket or, possibly, a point on the grid terminal a little inside the body of the socket. The calculated inductance of the contact spring is 4.5 milli-microhenries per centimeter of length within about ± 25 per cent, but the difference in effective lengths of the springs for the socket connection and the calibrating connection is not more than 2 or 3 millimeters.

Conductance values for the tubes are obtained by the susceptance-variation method. In this method, the circuit is detuned with either the capacitor or the inductor adjustment to a point giving half the power output observed at resonance. The increment in susceptance, determined from the capacitance calibration curves, is then equal to the circuit conductance. In practice, the mean value obtained by using the half-power points on each side of resonance is used. The range of the micrometer capacitor is sufficient for measurements of the circuit with the calibrating capacitors, with most cold tubes, and with some tubes under operating conditions. For other cases, adjustment of the inductor is required. The calibration curves used at each test frequency are:

(1) Capacitance for resonance

(2) Slope of the capacitance curve

(3) Conductance at resonance of the circuit with the calibrating capacitors.

These three quantities are plotted against the inductor adjustment readings. Since the conductance values for the calibrating capacitors themselves are too small to affect the calibration appreciably, the conductance curve corresponds, essentially, to the equipment.

MEASUREMENT RESULTS

The results of measurements on twelve types of rf pentodes are given in Table I and Figs. 3 to 8. Table I gives data obtained at 100 megacycles under a typical operating condition, a hot cutoff condition, and for the cold tube. A representative sample of each type was chosen. Lines 1 through 6 give voltages, currents, and transconductance values for the tubes tested. Generally, the control-grid bias was adjusted to give a typical plate-current value. This

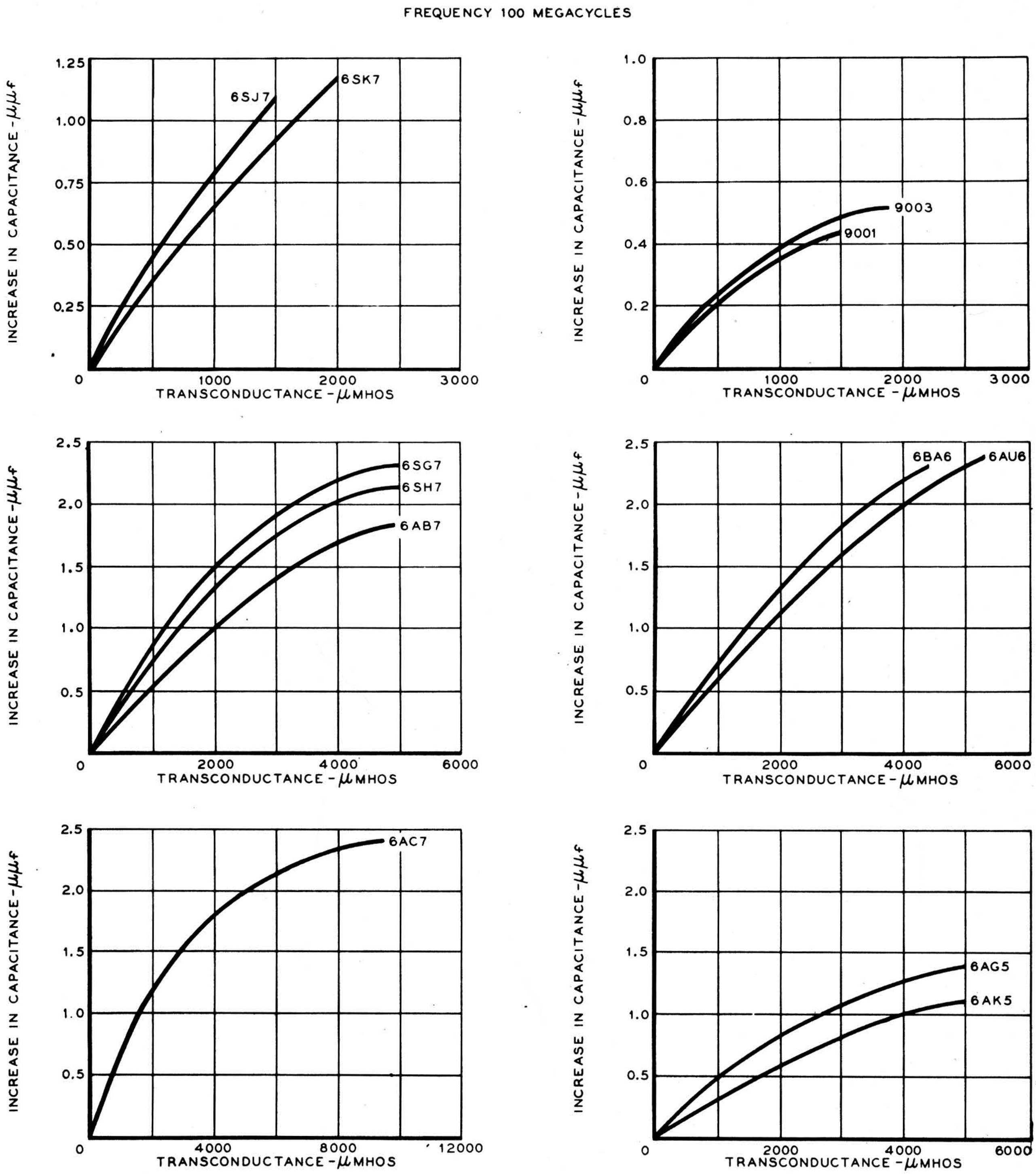

Fig. 3 - Change of Short-Circuit Input Capacitance with Transconductance.

procedure tends to minimize the effects of contact-potential changes which might occur during the series of tests made on each tube.

The short-circuit input capacitance values at 100 megacycles for the twelve tube types are given in lines 7, 8, and 9 of Table I. The increase (line 10) observed from the cold condition to the hot cutoff condition is low enough in each instance so that it may reasonably be attributed to thermal expansion of the cathode. The increase in capacitance from cutoff to the chosen operating condition is caused primarily by the space charge between cathode and control grid. The curves of Fig. 3 show this increase as a function of transconductance. The increase in capacitance with transconductance is not a high-frequency effect and the curves and data apply to lower frequencies except for minor corrections due to inductances in the tube leads.

Calculations based on capacitance data obtained in the frequency range of 50 to 125 megacycles indicate an increase in capacitance between zero and 100 megacycles of 1.2 μμf for type 6AC7, 0.8 μμf for type 6SH7, and 0.2 μμf for type 6BA6. The apparent series-inductance values corresponding to these increases are 9, 9, and 4 milli-microhenries. The increase of capacitance with frequency for cold tubes is found to be a little greater than the increase for tubes under operating conditions, so the change in capacitance with transconductance is generally a little greater at low frequencies than at 100 megacycles. The increases in capacitance and apparent series-inductance values for the other tube types are inside the ranges of the values given above. Line 19 of Table 1 gives the cold grid-to-cathode capacitances. The calculated increase in capacitance for a tube of ideal parallel-plane structure, under the assumption that the emitted electrons have zero initial velocity, is one-third of the grid-to-cathode capacitance. The observed increases (line 11 Table I) are much greater, but the curves for increase in capacitance with transconductance show that even these increases do not represent saturation values.

The short-circuit input conductance values observed at 100 megacycles are given in lines 12, 13, and 14 of Table 1. Fig. 4 shows the variation of input conductance with transconductance. The increases in input conductance values observed from the cold to the hot cutoff conditions (line 15, Table 1) are erratic, but the magnitudes of these differences are small in comparison with values under operating conditions. The curves of input conductance versus transconductance show a wide variety of shapes and indicate that in some cases it is very poor approximation to assume that the input conductance is proportional to the transconductance. Curves of short-circuit input conductance versus frequency for hot and cold tubes are shown in Figs. 5 and 6. In Fig. 5, the curves for most of the types tested indicate that, to a satisfactory degree of approximation, the conductance is proportional to the square of the frequency. Type 6AC7 was the only tube showing marked increase in capacitance due to lead inductance, and when the appropriate correction to the input conductance data is applied, the slope of the curve for this type corresponds more closely to the square-law relation. The most conspicuous exceptions to a square-law relation between conductance and frequency were observed for types 6SG7 and 6SH7. Curves for two type 6SG7 tubes and one type 6SH7 tube are shown; the curves for the 6SH7 and one of the 6SG7 tubes are coincident. The shapes of the curves for these tubes suggest the possibility of a resonance effect involving the internal leads and capacitances. The curve for type 6SJ7 also shows some irregularity. In spite of these irregularities,

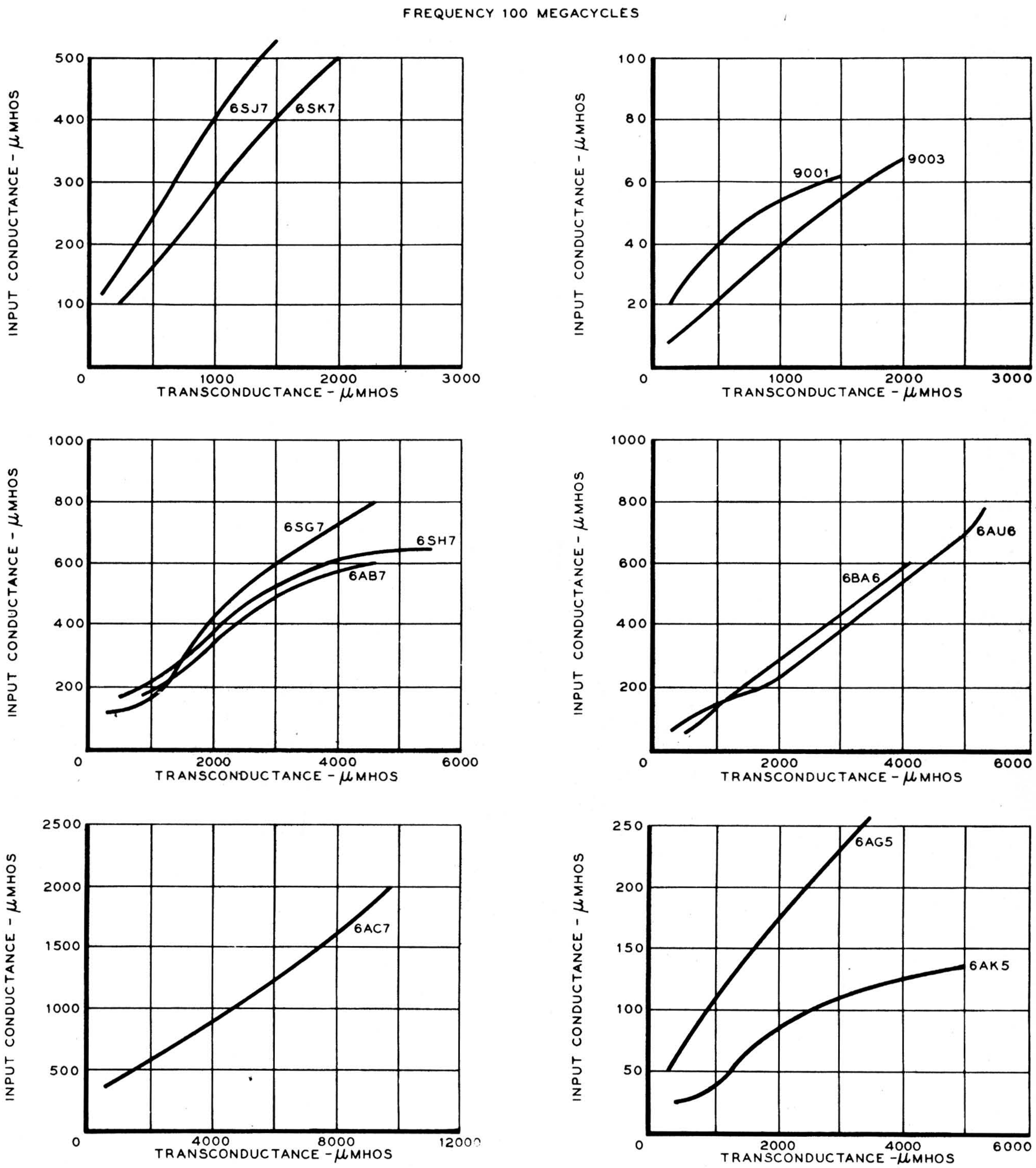

Fig. 4 - Change of Short-Circuit Input Conductance with Transconductance

however, the use of the values in line 12 of Table I and an assumed square-law variation with frequency can be recommended for situations in which high accuracy is not required.

The curves of Fig. 6 show that the conductance of a cold tube cannot be represented as a linear function of frequency in the 50-to-125-megacycle range. The shapes of the observed curves can be accounted for by assuming that the series resistance of the tube leads is an important factor in determining the conductance in this range. The resistance of the leads will increase with the square root of the frequency because of skin effect; the combination of this resistance in series with the tube input capacitance gives an effective conductance value across the terminals proportional to the five-halves power of the frequency. The series-resistance values appear to be of the order of one ohm for all the tube types. In the case of the metal types, there is evidence of some dielectric loss that is linear with frequency. The octal socket, of molded phenolic construction, shows a predominance of dielectric losses. The readings for the miniature socket could be accounted for by assuming a ten-ohm series resistance, but this is not reasonable in view of the results for tubes in the socket. The readings of conductance with the miniature socket are too low for accurate determination with the equipment used. The general conclusions that may be drawn from the curves of Fig. 6 are that the cold input conductance values for the tubes tested are almost negligible in comparison with the operating values in the 50-to-125-megacycle range, and that dielectric losses are responsible for the smaller part of even the cold input conductance values.

The principle components of the short-circuit input conductance at the high frequencies are:

(1) A component due to the time required for electrons to pass from the cathode through the control and screen grids.

(2) A component due to the inductance of the cathode lead.

A qualitative explanation of the first effect can be given as follows:

1. The grid of the tube can be considered as one electrode of a capacitor. When no current is flowing, the cathode and the screen grid, both at zero potential to signal frequency, form the other electrode.

2. When current is flowing through the tube, some of the lines of force from the grid terminate on electrons, and the effect is similar to a movement of the cathode toward the grid.

3. When the potential of the grid is varied slowly, the charge increases or decreases in phase with the variation of potential in the same manner as it would in the case of a capacitor. The charging current is proportional to the rate of change of the charge. For a sinusoidal variation of potential, the charging current leads the voltage by 90°.

4. When the potential of the grid is varied rapidly, the time of maximum charge is a little later than the time of maximum voltage. This occurs because the electrons starting from the cathode at the time the voltage reaches its maximum require an appreciable time to traverse the spaces between cathode and con-

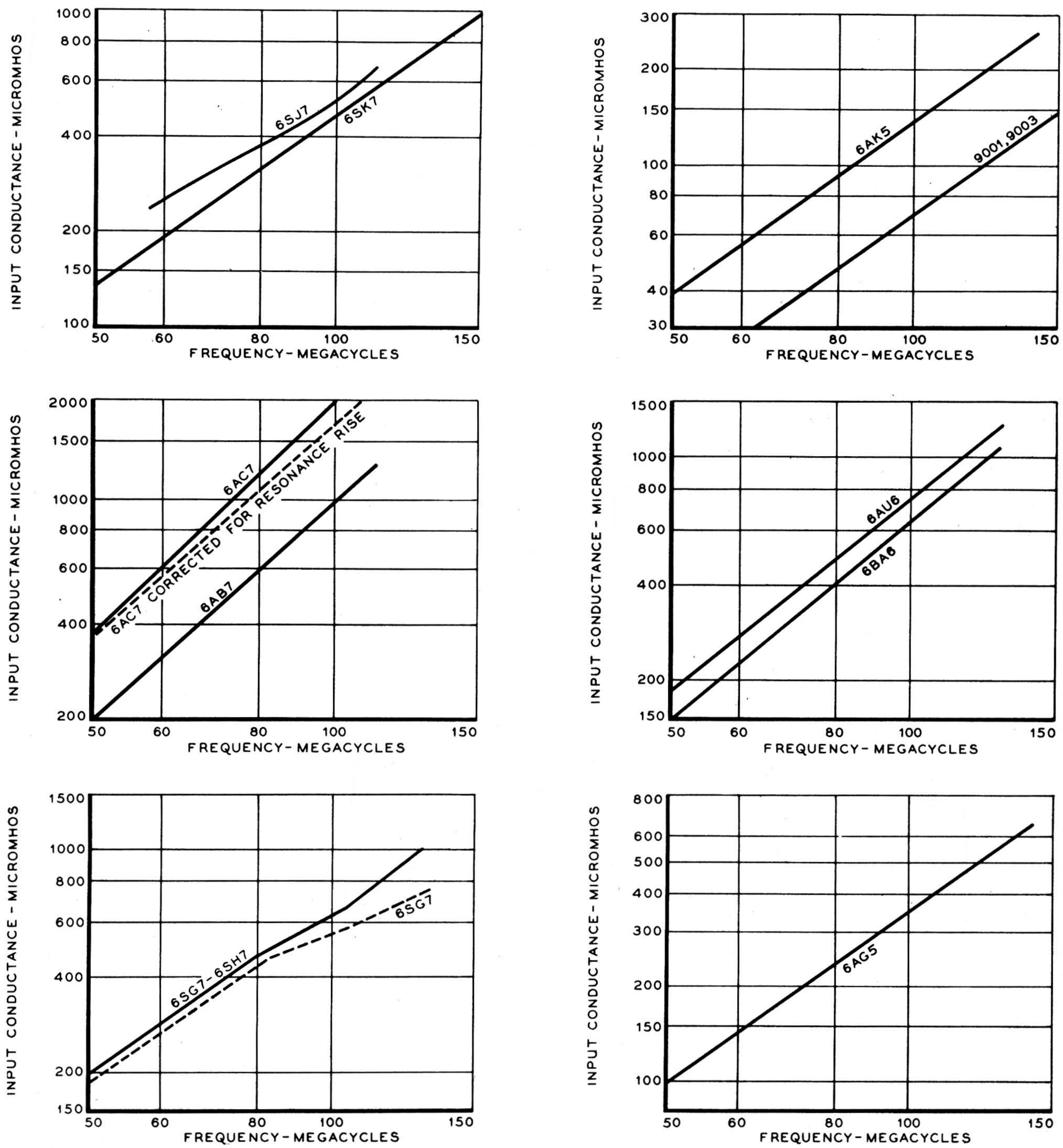

Fig. 5 - Change of Short-Circuit Input Conductance with Frequency (Typical Operating Conditions)

trol grid and between control grid and screen grid. These electrons constitute part of the charge so long as they are in either of these two regions.

5. Since the charge now lags behind the voltage, the charging current will lead the voltage by less than 90°. This means that there will be a component of current in phase with the voltage. The corresponding component of input conductance is the quotient of this component of current divided by the voltage.

6. The magnitude of the charging current is proportional to the frequency. The phase angle representing the lag of the charging current with respect to ninety degrees is also proportional to frequency. Consequently, the in-phase component of current (and, of course, the component of conductance derived from this source) is proportional to the square of the frequency.

7. These considerations indicate a linear dependence of input conductance on both the increment of capacitance due to space charge and (for small transit angles) on the transit time.

The second principle component of the short-circuit input conductance at high frequencies, the component due to the inductance of the cathode lead, can be evaluated as follows:

Let C_{gk} = capacitance, grid to cathode
L_k = inductance of cathode lead (to ground)
g_{kg} = transconductance, grid to cathode
E_1 = total input voltage
E_{gk} = voltage, grid to cathode
E_k = voltage, cathode to ground
i_{gk} = current, grid to cathode
i_k = current in cathode lead
f = frequency
$\omega = 2\pi f$

Then $E_1 = E_{gk} + E_k$
$E_k = j\omega L_k i_k$
$i_k = E_{gk}\, g_{kg}$
so $E_1 = E_{gk}\,(1 + j\omega L_k g_{kg})$
also $i_{gk} = j\omega C_{gk} E_{gk}$
Let Y_a = the component of input admittance corresponding to i_{gk}
g_a = the real (conductance) component of Y_a
then $Y_a = j\omega C_{gk} / (1 + j\omega L_k g_{kg})$
$= (j\omega C_{gk} + \omega^2 L_k C_{gk} g_{kg}) \,/\, (1 + \omega^2 L_k{}^2 g_{kg}{}^2)$
When the term $(\omega^2 L_k{}^2 g_{kg}{}^2)$ is small compared with unity,
$g_a = \omega^2 L_k C_{gk} g_{kg}$, approximately

The conductance g_a results from the shift in phase of the voltage E_{gk} with respect to the total input voltage E_1, and the consequent shift in phase of the current i_{gk}

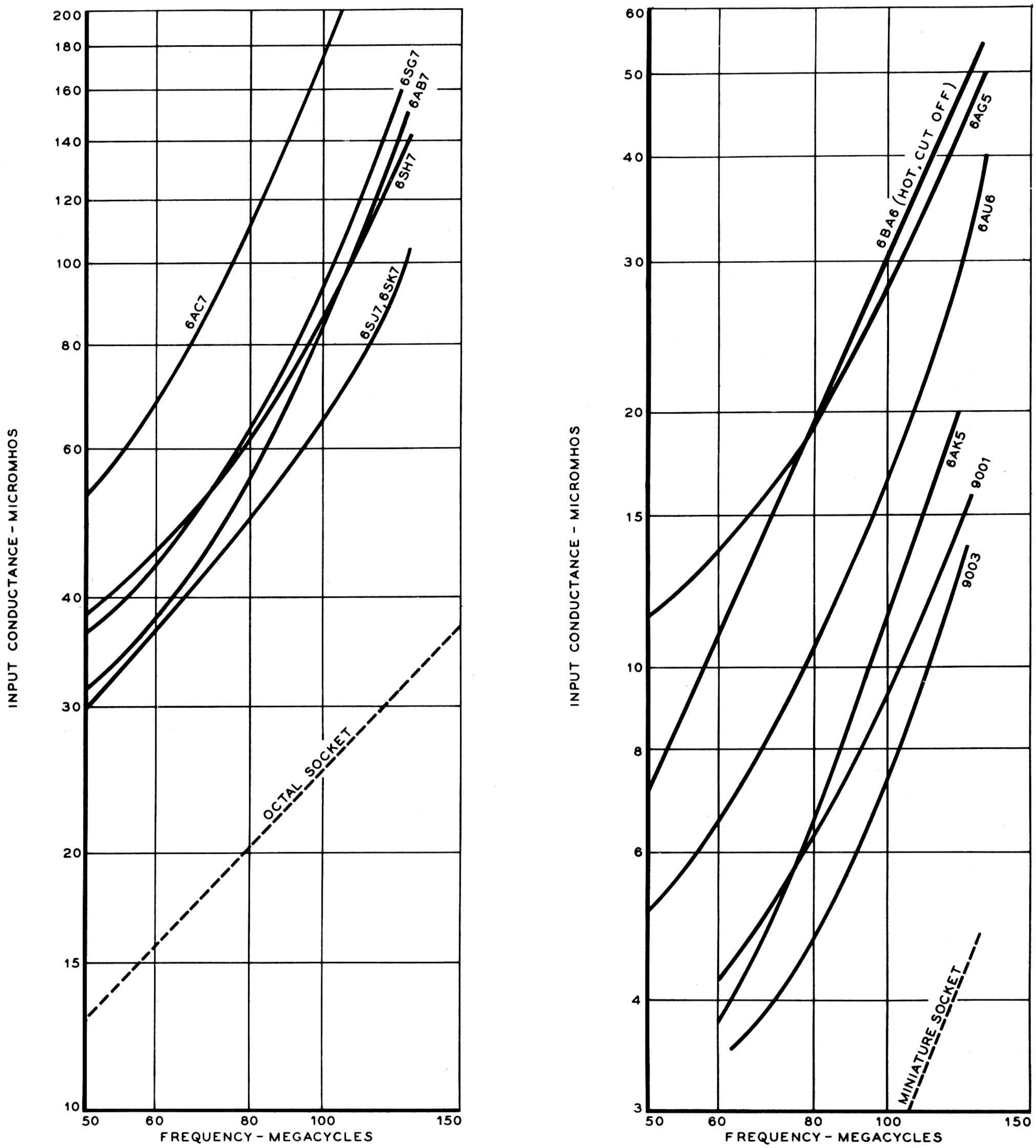

Fig. 6 - Change of Short-Circuit Input Conductance with Frequency (Cold Tubes and Sockets)

with respect to a quadrature relation to E_1. Currents flowing through the grid-to-screen capacitance or any other grid-to-ground capacitance are not affected by the presence of the voltage at the cathode, so the grid-to-cathode capacitance is the only component of input capacitance involved. A negative component of input conductance results from the inductance of the screen-grid lead in a pentode. The value of this component is proportional to the grid-to-screen capacitance and is equal to:

$$g_b = -\omega^2 L_2 C_{12} g_{21}$$

g_b = a component of input conductance

L_2 = inductance, screen grid to ground

C_{12} = capacitance, control grid to screen grid

g_{21} = transconductance, control grid to screen grid

Since the components of input conductance derived from cathode-lead inductance and from screen-grid lead inductance both vary with the square of the frequency, it is very difficult to distinguish them from the component resulting from electron transit time. For the tube types considered in this Note, calculated values of the input conductance components due to lead inductances at 100 megacycles range from about 10 per cent to 30 per cent or 40 per cent of the differences between operating-condition and cutoff values of short-circuit input conductance given in line 16, Table I. The measured values of grid-to-cathode capacitance (Line 19, Table I), increased by the cold-to-cutoff and cutoff-to-operating increments (lines 10 and 11, Table I), are used. Cathode-lead inductance values of 5 to 10 milli-microhenries for miniature tubes and 12 to 16 milli-microhenries for metal tubes are estimated. The lower values are used for tubes having two cathode leads. Allowance is made for the inductance of the screen-grid lead by using the grid-plate transconductance rather than the grid-cathode transconductance in the calculations. This is equivalent to an assumption that the two LC products involved are equal.

Example: Type 6AK5

Let g = the net input conductance component due to the cathode and screen-grid lead inductances.

C_{gk} = 2.3+0.3+1.1 = 3.7×10^{-12} farads

L_k (estimated) = 5×10^{-9} henries

g_m = 4950×10^{-6} mhos

f = 100×10^{6} cycles

then ω = 628×10^{6}

$L_k C_{gk}$ = 18.5×10^{-21}

ω^2 = 0.396×10^{18}

$\omega^2 L_k C_{gk}$ = 0.0076

and $g = \omega^2 L_k C_{gk} G_m$ = 36.1×10^{-6} mhos

The cutoff-to-operating increase in short-circuit input conductance for this type was observed to be 113 micromhos, so the lead-inductance components, as determined above, account for 32 per cent of this quantity. The assumed cathode—lead inductance, five milli-microhenries, is equivalent to the inductance of a 20-mil wire one centimeter long, with its center one-tenth of an inch from the center of a conductor carrying the return current. The total length of a lead from the socket mounting plate to the mount structure of the tube is more than one centimeter, but two separate socket leads and base pins are used in making connections to the cathode of type 6AK5.

APPLICATION OF INPUT CAPACITANCE MEASUREMENTS

The change in input capacitance with transconductance, observed when the control-grid bias is changed, has the effect of changing the resonant frequency of a circuit connected to the input terminals of a tube. The relative effect can be limited by using a large amount of fixed capacitance in parallel with the input, or the change in capacitance can be reduced by the use of an unbypassed cathode resistor in series with the cathode. The change in input capacitance with transconductance is most likely to cause difficulty in an intermediate-frequency amplifier. The frequencies of most immediate concern for this purpose are approximately 10 and 20 megacycles.

The data from Table I can be used to determine the approximate amount of unbypassed cathode resistance needed to compensate for the change in capacitance with transconductance. The voltage drop across an unbypassed cathode resistor is the product of its resistance, the grid-to-cathode transconductance, and the grid-to-cathode voltage. The total input voltage is the sum of the voltage at the cathode and the voltage between grid and cathode. The current flowing into the grid-to-cathode capacitance is the product of grid-to-cathode voltage and the susceptance of the grid-to-cathode capacitance. The increase in capacitance with transconductance applies to the grid-to-cathode component. To obtain full compensation for capacitance change it is necessary to make the product of cathode resistance and grid-cathode transconductance equal to the ratio of the increase in capacitance to the grid-to-cathode capacitance. The grid-to-cathode transconductance is approximately the product of grid-to-plate transconductance and the ratio of total cathode current to plate current. The resistance values ascertained by this method range from 32 ohms for type 6AC7 to 230 ohms for type 6SK7. The unbypassed cathode resistor reduces the effective transconductance by the ratio:

$$1/\,(1 + r_k\, g_{kg})$$

Values for this ratio range from 0.78 for type 9001 to 0.58 for type 6AU6. The bias voltages developed across the cathode resistors required for compensation with types 6SH7, 6SG7, 6AU6, and 6BA6 exceed the typical operation values given in Table I, so the transconductance would be reduced some more on that account.

Since the ratio of the increment in capacitance to the grid-to-cathode capacitance determines the resistor value and the consequent reduction in gain, an increase in the cold grid-to-cathode capacitance can lead to improved overall results. This can be accomplished by connecting a small capacitor between the grid and cathode terminals at the tube socket. The use of too large a value of capacitance for this purpose would have an unfavorable effect on the input conductance. For type 6BA6 the hot cutoff grid-to-cathode capacitance is 3.5 μμf and the increment from cutoff to operating conditions is 2.2 μμf. The ratio of these two values is 0.63. Since the grid-cathode transconductance is 6000 micromhos, a 105-ohm resistor is required. With this resistor, the grid-plate transconductance would be reduced from 4300 to 2560 micromhos. To obtain even this value it would be necessary to return the resistor to a point about 0.3 volts negative with respect to the grid. If the grid-to-cathode capacitance is doubled by adding 3.5 μμf between the grid and cathode terminals of the socket, the ratio of the increment to the capacitance at cutoff becomes 0.31, the resistor required is 52 ohms, and the effective grid-plate transconductance with this resistor is 3300 micromhos. The 105-ohm resistor in series with 3.5 μμf results in an input conductance component at cutoff of 5.1 micromhos at 10 megacycles or 20.4 micromhos at 20 megacycles; the 52-ohm resistance in series with the 7.0-μμf capacitance results in an input conductance component of 10.2 micromhos at 10 megacycles or 40.8 micromhos at 20 megacycles. Application of square-law extrapolation to the 100-megacycle value of input conductance for type 6BA6 leads to estimated conductances of 6 micromhos at 10 megacycles and 24 micromhos at 20 megacycles under operating conditions without the cathode resistor. The circuit conductances in 10-megacycle or 20-megacycle if amplifiers are generally considerably higher than these values. The increase in input conductance at cutoff limits the amount of capacitance which can be added between grid and cathode.

The curves of Fig. 7 show the results obtained with the 6BA6 at a frequency of 100 megacycles by using an unbypassed cathode resistor and adding grid-to-cathode capacitance. The conductance values to be expected at 10 megacycles and 20 megacycles can be estimated by dividing the values from Fig. 7 by 100 and by 25. The agreement between these measured results and the calculations given above is very good. Types 6SG7, 6SH7, 9001, 9003, 6AG5, and 6AK5 have relatively high plate-to-cathode capacitances because of internal connections between their suppressor grids or beam-confining electrodes and their cathodes. The use of an unbypassed cathode resistor results in an increase in the short-circuit feedback admittance when the plate-to-cathode capacitance is high. Consequently, circuits using tubes of the above-mentioned types with unbypassed cathode resistors should be carefully checked for oscillation.

APPLICATION OF INPUT CONDUCTANCE MEASUREMENTS

When an rf amplifier stage is connected between an antenna transformer and a converter (or a second rf stage), maximum voltage at the grid of the tube will be obtained with the transformer adjusted for matched impedances. If circuit losses are negligible, the maximum voltage gain from the antenna terminal to the grid terminal will be the square root of the ratio of antenna conductance to grid conductance. The gain from grid to plate will be the ratio of the transconductance to the conductance of the interstage circuit (including the short-circuit output conductance of the tube.) The gain from antenna to plate, therefore, is proportional to the quotient of transconductance divided by the square root of the input conductance.

The following example illustrates the method by which a specified selectivity requirement can be met when the input conductance and the input capacitance for a tube are given. It is assumed that the unloaded Q value for the circuit added to obtain selectivity is independent of the L/C ratio of the circuit. The quantity Q is commonly thought of as the ratio of the reactance of the inductor to the equivalent series resistance in a circuit; however, it is also the ratio of the susceptance of the capacitance to the

CURVES	R_K	C_{gK}	INPUT CAPACITANCE AT CUTOFF
CURVE I	0	0	7.35 μμf
CURVE II	47 OHMS	2.5 μμf	9.40 μμf
CURVE III	47 OHMS	3.8 μμf	10.8 μμf

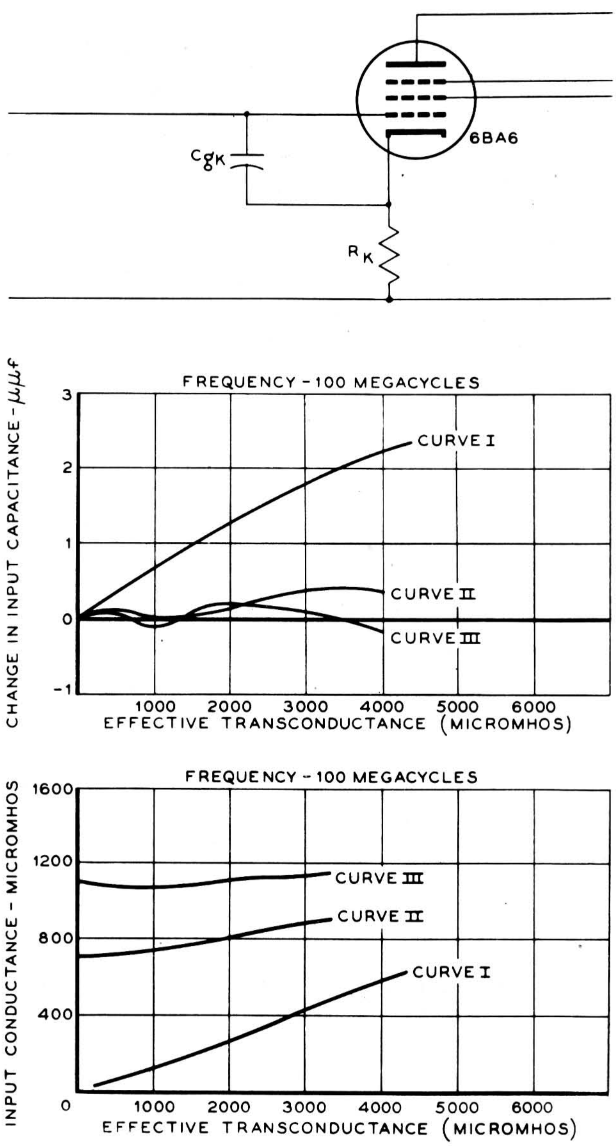

Fig.7 - Input Admittance Data for Type 6BA6 with Unbypassed Cathode Resistor

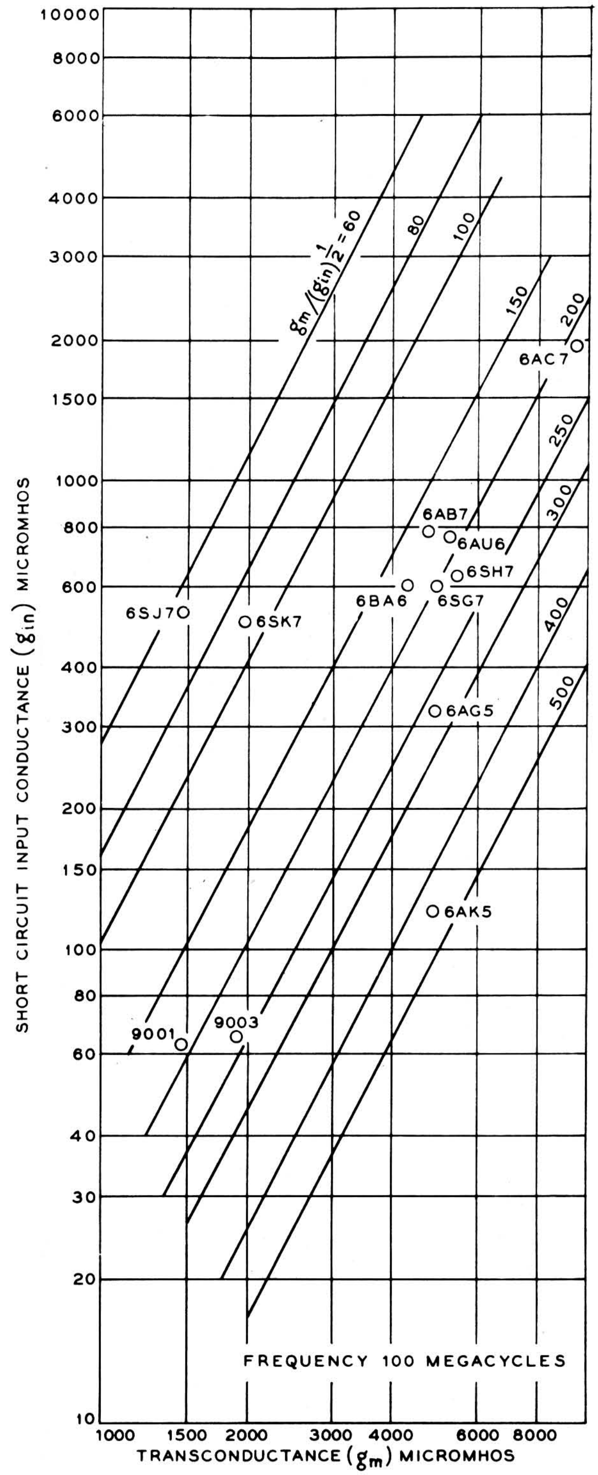

Fig.8 - Short Circuit Input Conductance Data for Several Tube Types

equivalent shunt conductance. A 100-megacycle circuit with a Q of 50 will provide voltage attenuation ratios of approximately 20 to 1 at frequencies 20 megacycles away from resonance.

EXAMPLE

Given:

Frequency, f	100 megacycles
Tube input conductance, g_1	500 micromhos
Tube input capacitance, C_1	8 micro-microfarads
Circuit Q, unloaded.	200

Required:

Loaded circuit Q = 50, for selectivity

Antenna coupling = 80% of value for maximum gain, to limit the dependence of the circuit on antenna characteristics.

Let g_1 = tube conductance = 500 micromhos

g_2 = circuit conductance

B_1 = tube susceptance = $2\pi fC$ = 5030 micromhos

B_2 = circuit susceptance = 200 g_2 (because Q = 200)

For 80% of critical coupling,

Conductance reflected from antenna = $(0.8)^2 x\ (g_1+g_2)$

Total conductance = $1+(0.8)^2 x(g_1+g_2)$ = $1.64\ (g_1+g_2)$

Required susceptance = 50 x total conductance (for Q = 50) = $82\ (g_1+g_2)$

Tube susceptance = 5030 micromhos = $10.06g_1$

Required circuit susceptance = Required susceptance - Tube susceptance

= $72\ g_1+82\ g_2 = 200g_2$

Solve for g_2: $g_2 = 0.61g_1$ = 305 micromhos

B_2 = 61000 micromhos

$C_2 = B_2/2\pi f$ = 97 micro-microfarads

Check: Total conductance = 1.64 (500+305) = 1320 micromhos

Total susceptance = 105 x $2\pi f$ = 66000 micromhos

Therefore, Q = 50

When considerations such as the range of a variable tuning capacitor call for a lower value of circuit capacitance than that determined from selectivity considerations, the same result may be obtained by connecting the grid of the tube to a tap on the inductor.

The gain for this circuit may be calculated as follows:

Given: Dummy antenna resistance = 300 ohms

Transformer secondary load = g_1+g_2 = 805 micromhos = 1240 ohms

Coupling = 80% of coupling for matched impedances

Transformer voltage ratio for matched impedances = $(1240/300)^{\frac{1}{2}}$ = 2.03

Ratio for 80% of this coupling = 0.8 x 2.03 = 1.62

Reflected resistance at primary terminals = $1240/(1.62)^2$ = 471 ohms

Fraction of signal-generator voltage across primary terminals = 471/(471+300) = 0.61

Gain, signal generator to transformer secondary terminals = 0.61x1.62 = 0.99

In the above example, the circuit conductance g_2 is found as a quantity proportional to the tube conductance g_1, and the total conductance, therefore, is proportional to g_1. Since the gain is inversely proportional to the square root of the total conductance, it follows that the gain is inversely proportional to the square root of the tube input conductance.

The use of more than one rf stage at 100 megacycles is not common practice, but interstage circuit considerations are similar to those for an antenna coupling circuit. When a tube having a high input conductance is used as the second stage in an rf amplifier, it is generally desirable to arrange the equivalent of a step-down transformer between the plate of the first tube and the grid of the second. One simple way of accomplishing this is to use a series blocking capacitor which is small in comparison with the input capacitance of the second tube. Then, for a specified conductance reflected to the plate of the first tube, the gain to the grid of the second tube becomes inversely proportional to the square root of its input conductance.

RELATIVE FIGURE OF MERIT OF TUBE TYPES

Fig. 8 shows the short-circuit input conductance values plotted against transconductance values for the twelve types discussed in this Note. Since the lines drawn on the chart represent constant values of the quotient of the transconductance divided by the square root of the input conductance, the relations among the types with respect to this figure of merit can be readily seen. Type 6AK5 occupies the most favorable position on the chart. Types 9001 and 9003 are good in spite of their low transconductance values. Type 6AC7 occupies a favorable position on the chart because of its high transconductance even though it has the highest input conductance of any of the types tested.
