THE USE OF POWDERED IRON IN TELEVISION DEFLECTING CIRCUITS

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Abstract.—The horizontal deflection of electron beams in television systems has required excessive energy dissipation and expensive circuit components. For deflection by magnetic means, the deflection transformer and yoke have presented serious problems in the economical design of television receivers.

Low-loss systems have been constructed with energy recovery arrangements. Such systems, requiring no additional electrical energy, provide large increases in deflection capability. Simultaneously there are reductions in transformer core costs to less than one fourth those of equivalent laminated sheet or strip metal types.

Transformer and yoke cores have been molded from powdered iron materials especially prepared for these applications. Very low-cost materials have been developed to produce useful effective a-c permeabilities of between 40 and 230. The precise value depends upon the molding pressure, the degree of d-c saturation and the peak amplitude of the a-c flux density.

Small particle thicknesses available, at low cost, in powdered-iron materials make high-Q systems possible. Increased efficiency eliminates the necessity for dissipating large amounts of energy from the transformer and deflecting yoke structures.

Molded core structures, in comparison with laminated core structures, produce negligible acoustic radiation.

A low-cost system has been constructed to provide full deflection and 17 kilovolts second anode potential for a sixty degree kinescope driven by two type 807 or 6BG6G beam-tetrodes. The present pulse voltage ratings of available tubes limit the second anode voltage to approximately 17 kilovolts for a kinescope which is to be scanned from a circuit driven by a *single* 6BG6G tube. This second anode voltage may be derived from windings on the same deflection transformer via a voltage doubling rectifier.

Design theory for low-loss scanning systems has progressed so that now the resultant transformer designs may be relied upon to produce the expected results within the limits of the other component part tolerances and tube tolerances. Design equations and curves are given.

I. INTRODUCTION

THE horizontal deflection of electron beams and the high-voltage power supply for the electron beam acceleration in television receivers has accounted for a large portion of the required total energy dissipation, physical volume and cost. The advent of modern short-length, sharp-focus, kinescopes of wide deflection angle increased the energy consumption and the cost of the required component parts. Means have been found for decreasing the cost appreciably, and at the same time providing an equally efficient system.

The major component parts are the horizontal deflection transformer and its associated deflection yoke. These devices have been responsible for at least half the energy loss in many systems. The introduction of very thinly laminated cores in both units resulted in a considerable reduction in energy loss and a marked increase in the deflecting capability of any standard driver tube.

The cost of cores with very thin laminations is quite high. To reduce this expense, a special type, low-cost, high-permeability molded iron-dust core has been developed. Although its characteristics are different from those of the laminated cores, the operating results are similar and the cost of the cores is a small fraction of that of other adequate cores. In addition, the acoustic radiation from iron-dust cores is negligible.

As stated, low-cost sponge-iron and electrolytic-iron powders have been used. In each case the mean particle thickness was about 0.0005 inches. The cost of sponge iron is about half that of the electrolytic iron, and it requires no additional intraparticle insulation.

II. MOLDED CORES

In the manufacture of molded iron-dust cores, the iron powders, after being mixed with very small amounts of binders, are pressed into core shapes under very high molding pressures. These materials do not flow appreciably under compression, so the molds must be of the straight thrust type. The minimum satisfactory molding pressure is about 15 tons per square inch. Higher values up to 60 tons or more are desirable magnetically. The cost of higher pressures requires careful balancing against the technical advantages.

In Fig. 1 the apparent permeability of the powdered sponge-iron core material is plotted relative to molding pressure at a number of values of flux density, in low frequency operation. The increase in apparent permeability with increased molding pressure is obvious.

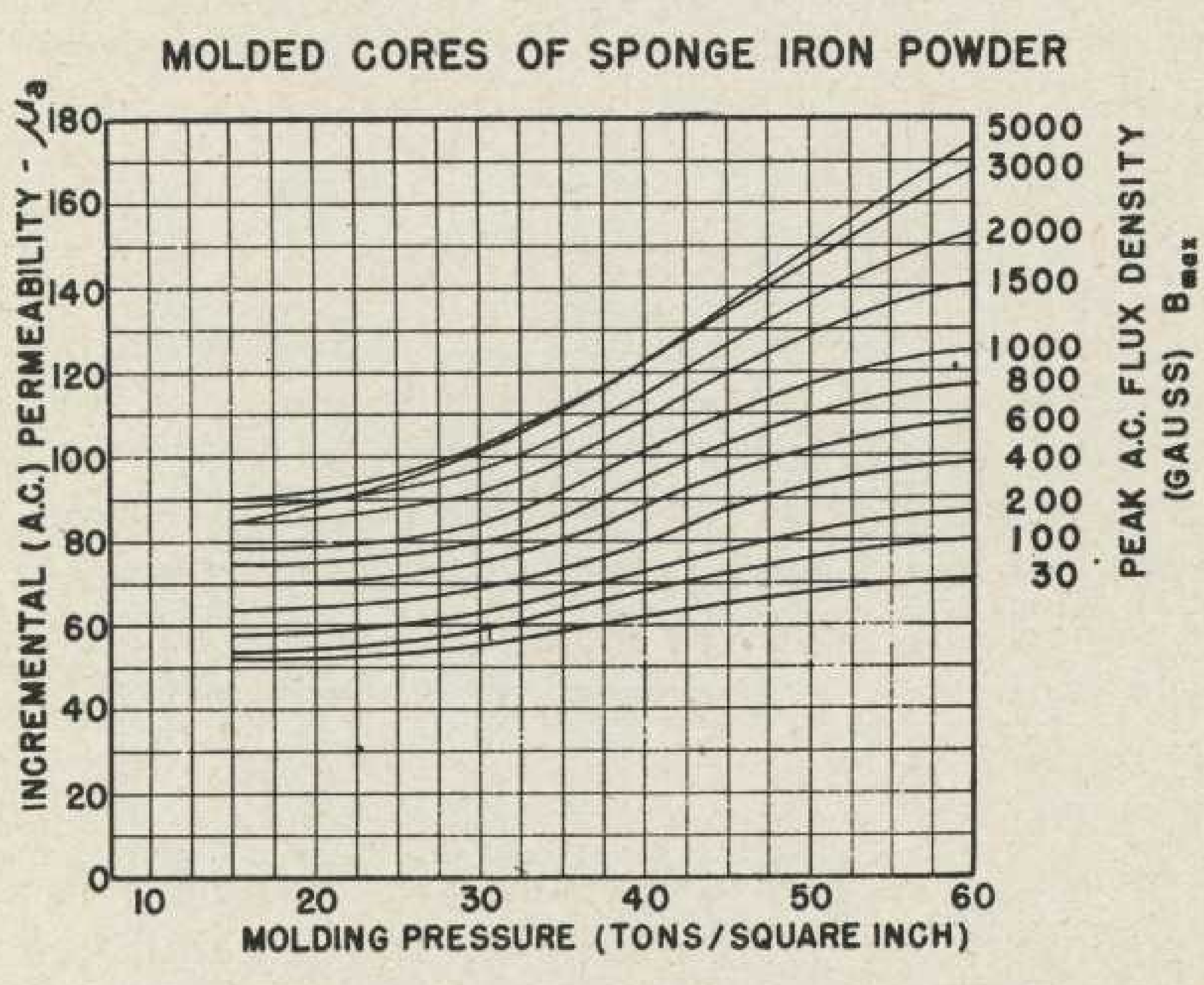


FIG. 1.—Apparent incremental a-c permeability μ_a of molded sponge-iron-powder core materials as a function of molding pressure, plotted for various values of peak a-c flux density B_{max} in gauss, at 60 cycles per second.

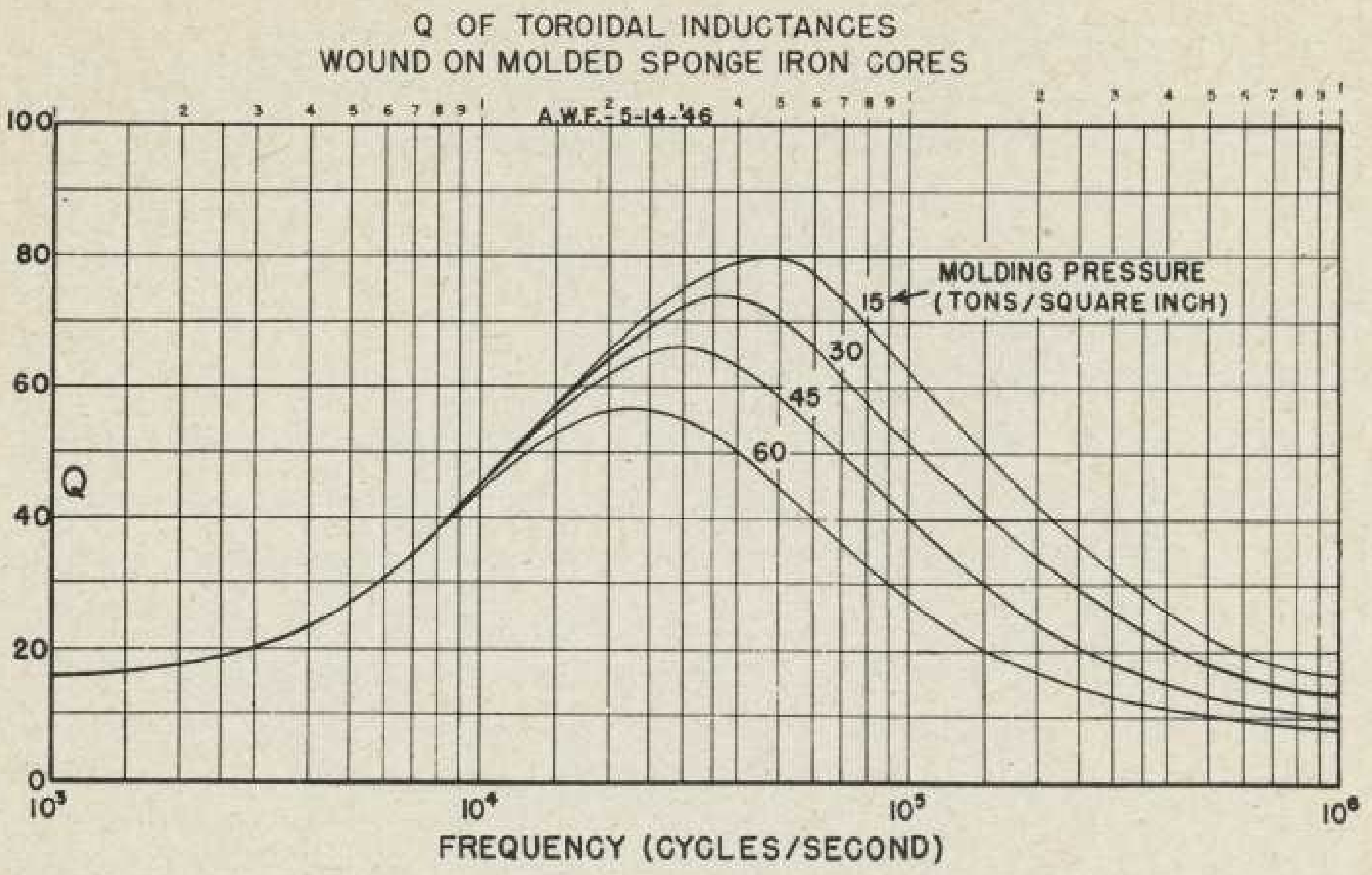


FIG. 2.—Plot of Q as a function of frequency and molding pressure for samples of molded sponge-iron-powder cores with identical toroidal windings of litz wire.

The relative Q values of identical coils on toroidal sponge-iron cores made at various molding pressures are plotted in Fig. 2. At frequencies in the vicinity of 70 kc, the losses are due chiefly to eddy currents in the core material particles. It is noteworthy that while increasing the molding pressure produced a reduction in Q, the reduction is not sufficiently great to be particularly disturbing for television horizontal deflection uses. The values of Q for electrolytic-iron cores are similar to those of 30 tons per square inch sponge-iron cores. They do not vary with pressure changes between 15 and 60 tons per square inch. This powder is especially treated to produce a superior insulating surface on each particle.

It is interesting to observe the effect of changing frequency upon the apparent permeability of the various samples of materials at the different molding pressures. Figures 3 and 4 illustrate these results for low flux density operation with sponge and electrolytic iron respectively. The particle sizes were chosen so that the apparent permeability remained very nearly constant over the range of frequencies up to 100 kc. The spectrum between 15 and 100 kc contains all the useful horizontal scanning frequencies, insofar as the present monochrome television is concerned.

The effects of variation of the maximum a-c flux density B_{max} (at several frequencies) upon the apparent incremental permeabilities of the various cores are illustrated by Figs. 5, 6, and 7. It is interesting to note that as the material is more closely compacted the variation in apparent permeability with changing magnetomotive force becomes progressively greater. The maximum value indicated for any of the tested materials is 230, for electrolytic iron molded at 60 tons per square inch. At the same pressure, sponge iron yields a permeability of 175.

Various sample moldings are shown in the photographs of Figs. 8 and 9. Figure 8 is an exploded view of a transformer core molded with a two piece

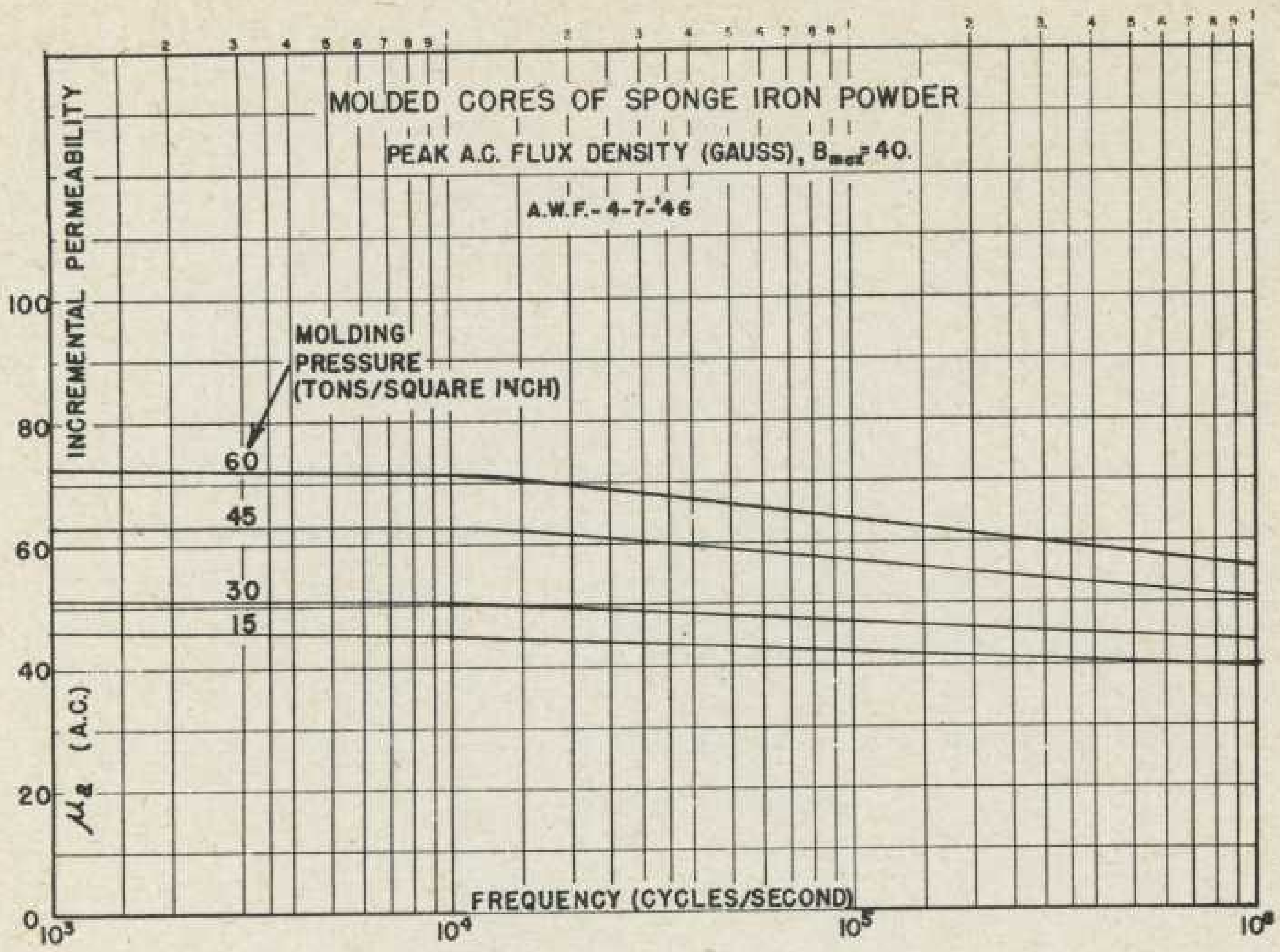


FIG. 3.—Incremental a-c permeability μ_a of sponge-iron-powder cores, molded at various pressures, as a function of frequency, at a peak a-c flux density B_{max} of 40 gauss.

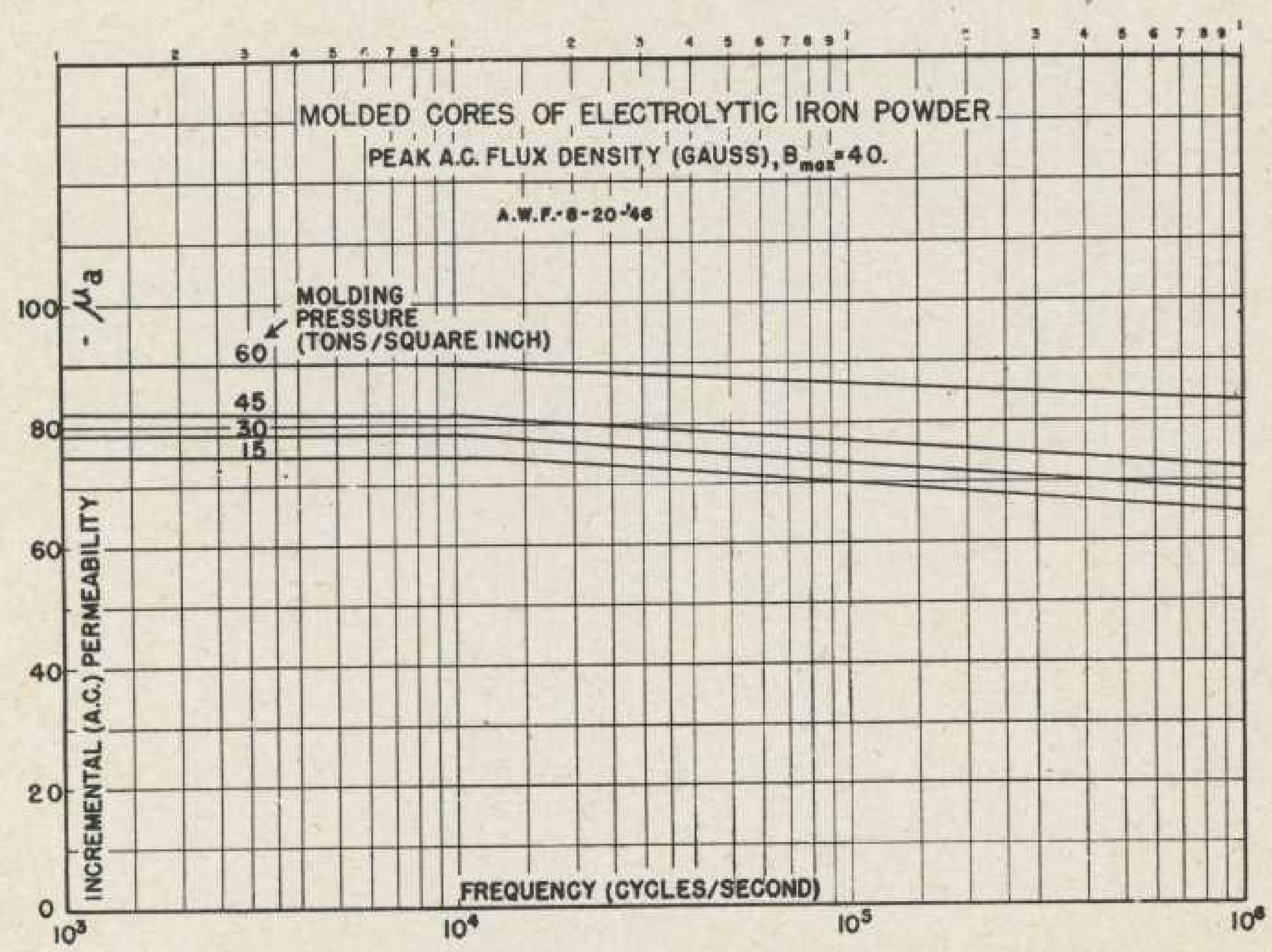


FIG. 4—Incremental a-c permeability μ_a of electrolytic-iron-powder cores, molded at various pressures, as a function of frequency, at a peak a-c flux density B_{max} of 40 guass.

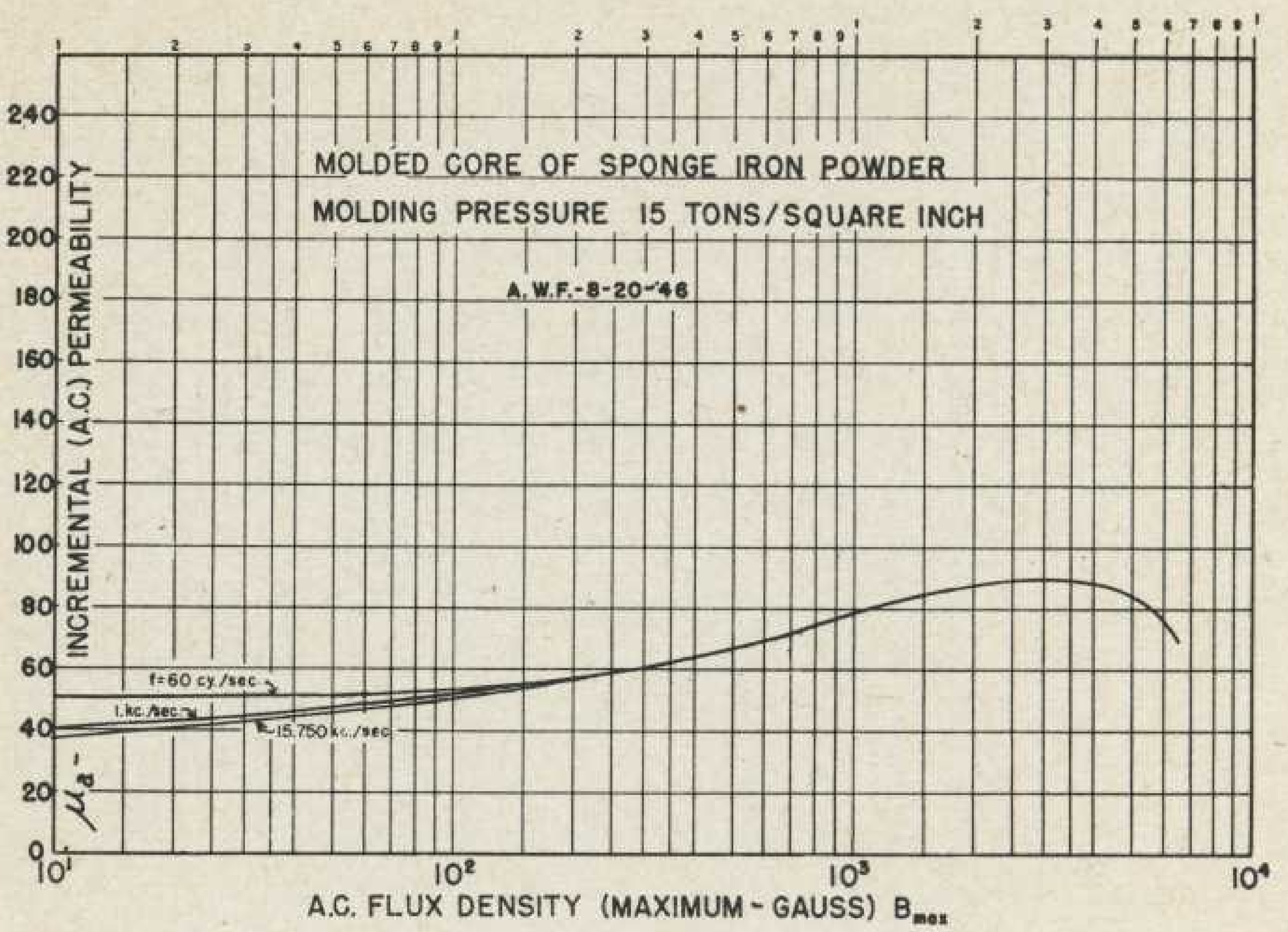


FIG. 5—Incremental a-c permeability μ_a of a sponge-iron-powder core, molded at a pressure of 15 tons per square inch, as a function of peak a-c flux density B_{max} in gauss, at frequencies of 60, 1,000 and 15,750 cycles per second.

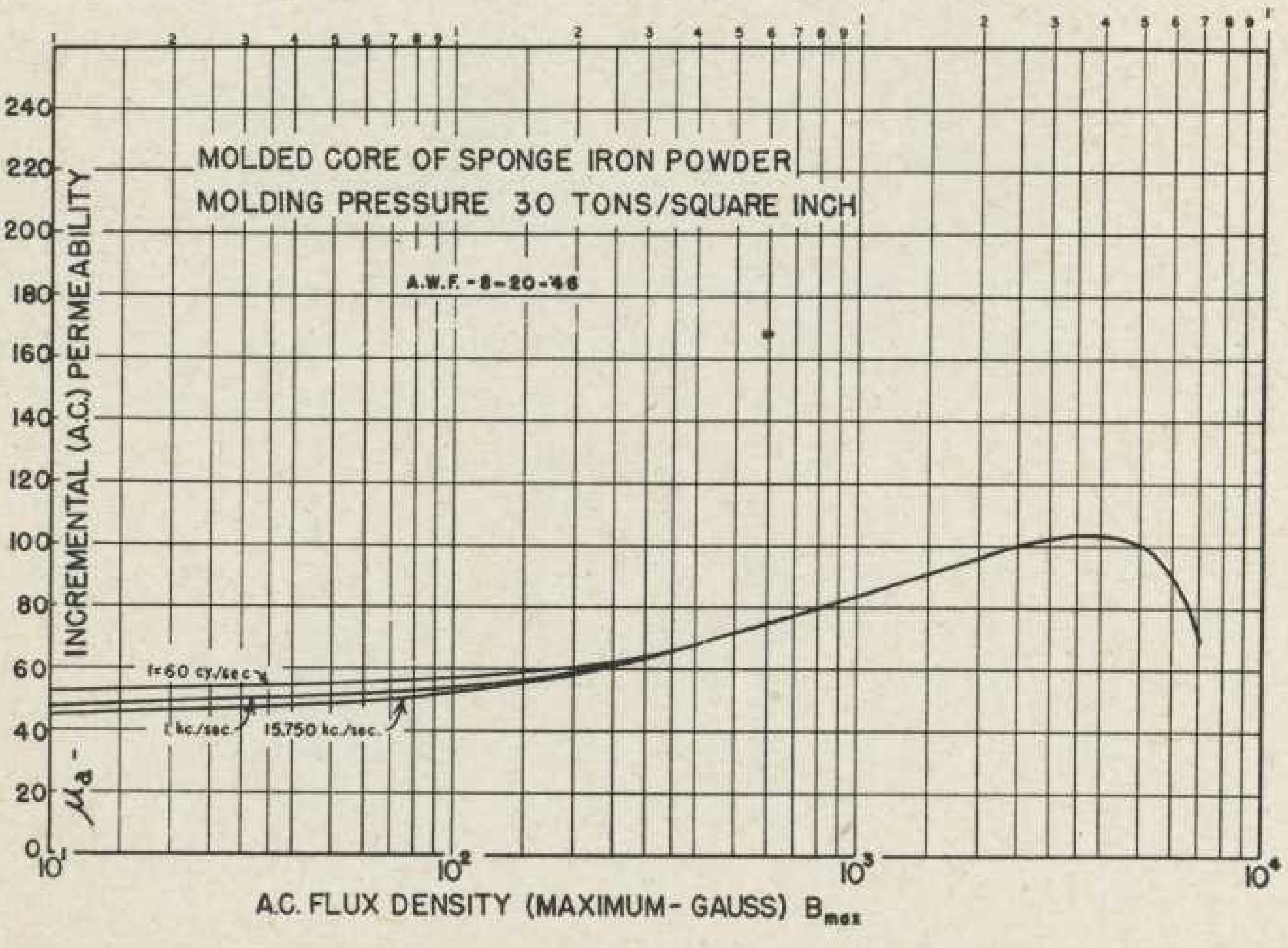


FIG. 6—Incremental a-c permeability μ_a of sponge-iron-powder core, molded at a pressure of 30 tons per square inch, as a function of peak flux density B_{max} in gauss, at frequencies of 60, 1,000 and 15,750 cycles per second.

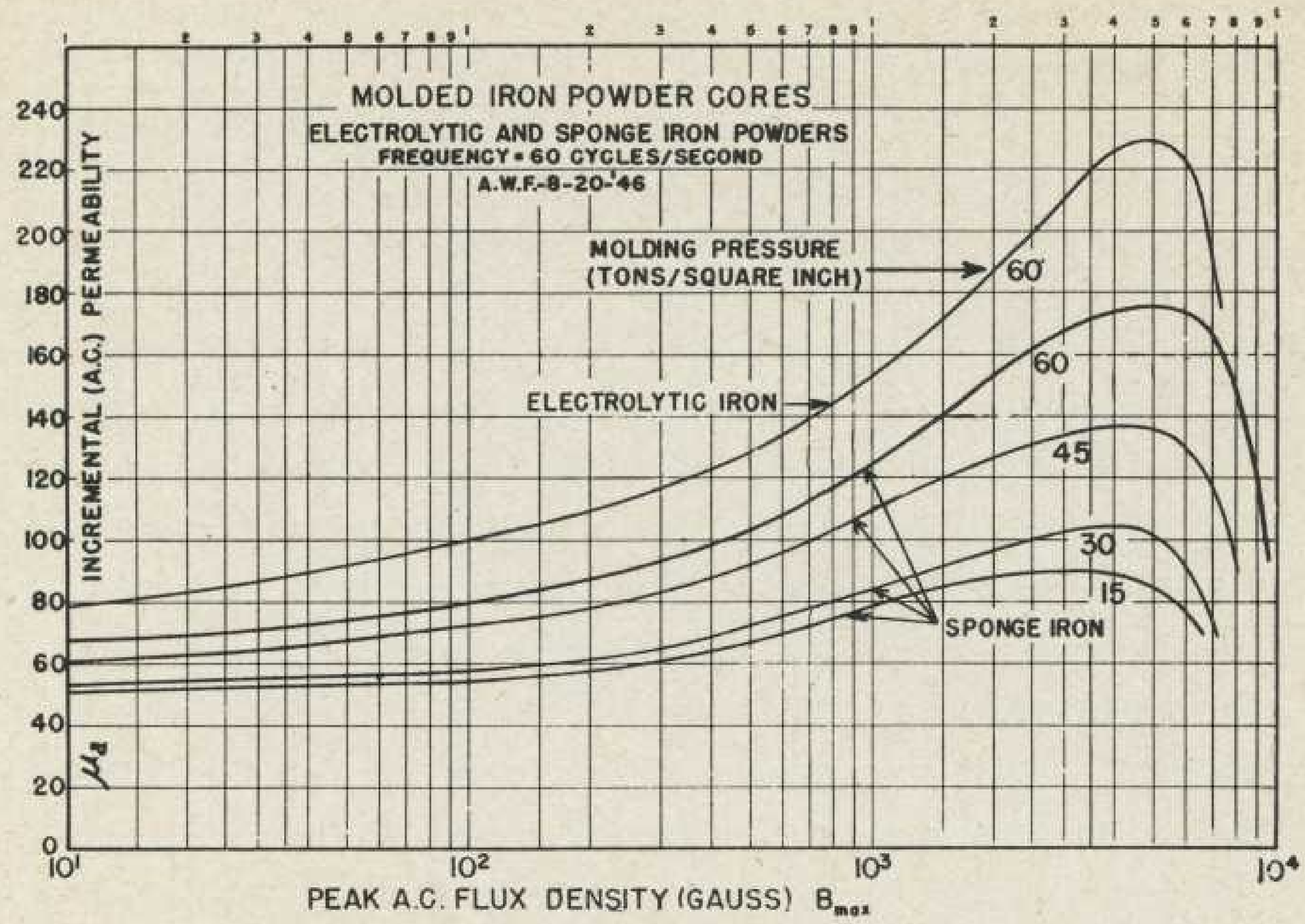
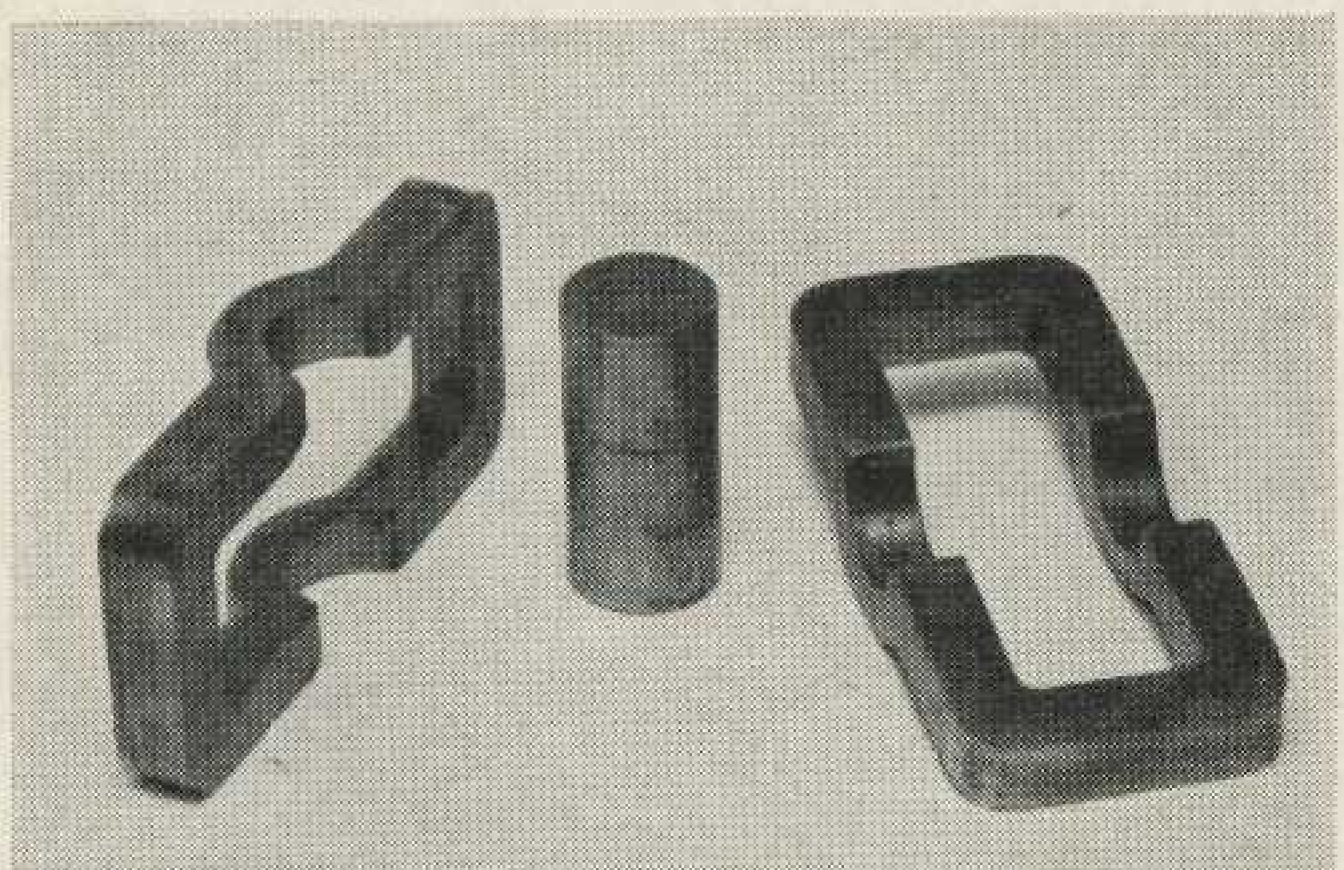


FIG. 7—Incremental a-c permeability μ_a of sponge and electrolytic-iron-powder cores, molded at various pressures, as a function of peak a-c flux density B_{max} in gauss, at 60 cycles per second.



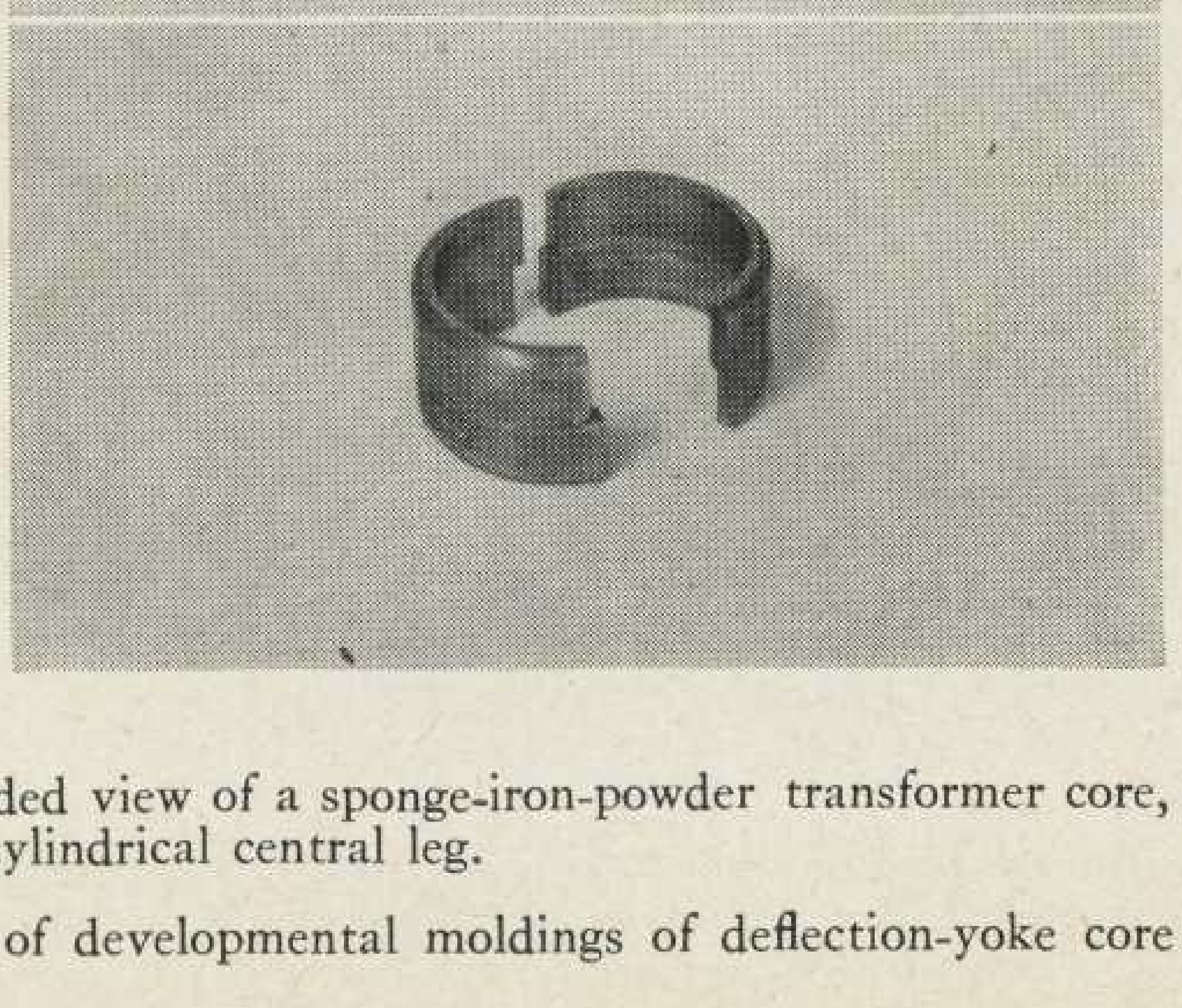


FIG. 8—An exploded view of a sponge-iron-powder transformer core, molded with a two piece frame and a cylindrical central leg.

FIG. 9—Samples of developmental moldings of deflection-yoke core sections of sponge-iron powder.

frame and a cylindrical central leg. Figure 9 shows samples of molded yoke core sections used in early tests.

In Fig. 10 is an unassembled view of a production type horizontal deflection transformer. The molded iron-dust core is to be clamped together by the two sheet-metal clips which are then held together by the long tie-strap. The coil and terminal board assembly are held in place by the core members. The two-turn coil of polyethylene insulated wire is supported between the two textolite discs. It is used as a filament winding for a type 1B3-GT/8016 pulse rectifier tube supplying the first and second anode voltages to the kinescope. Two views of the assembled transformer are shown in Figs. 11 and 12.

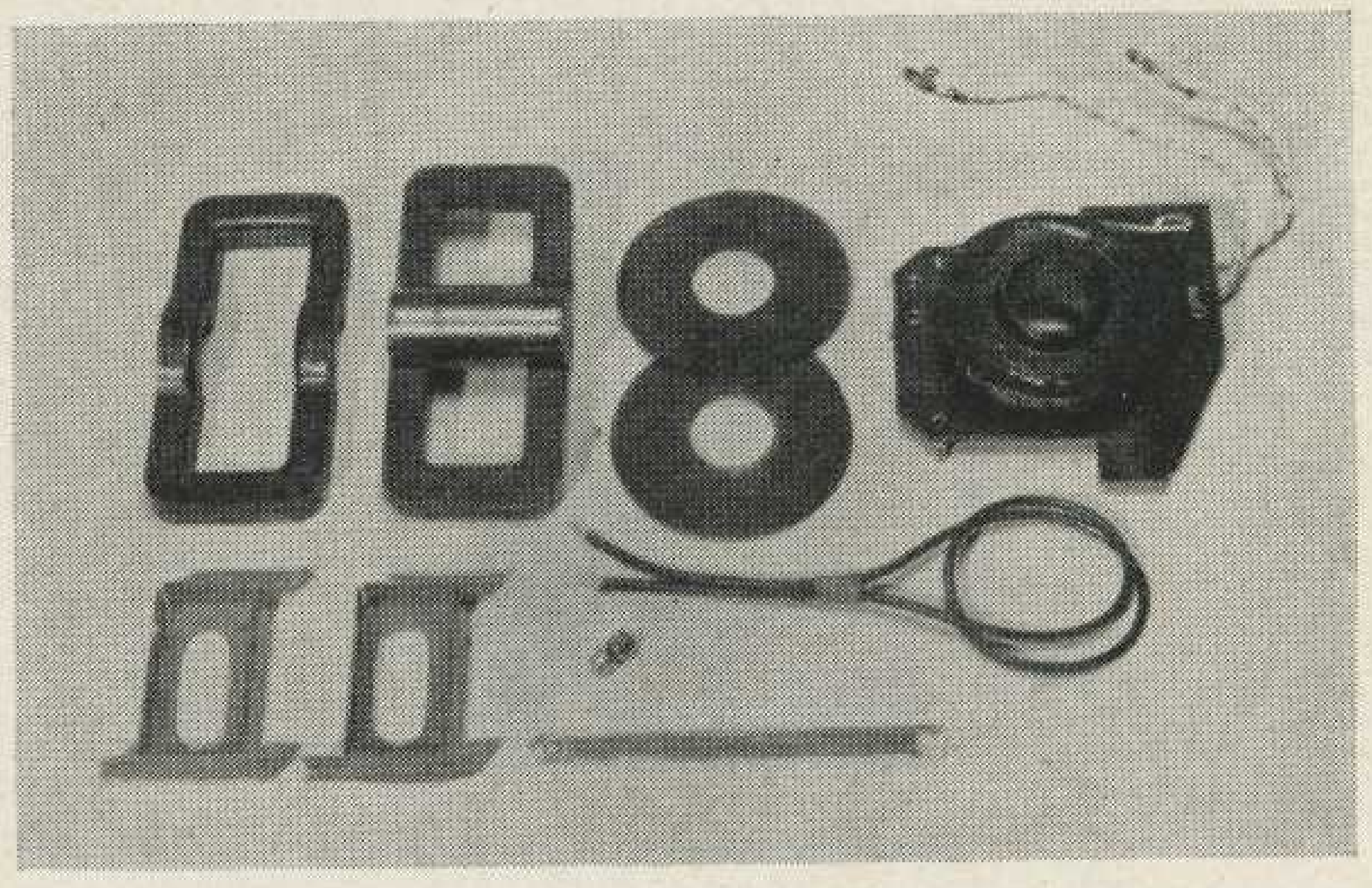


Fig. 10—An unassembled view of a complete production-type horizontal-deflection transformer with a high-voltage winding.

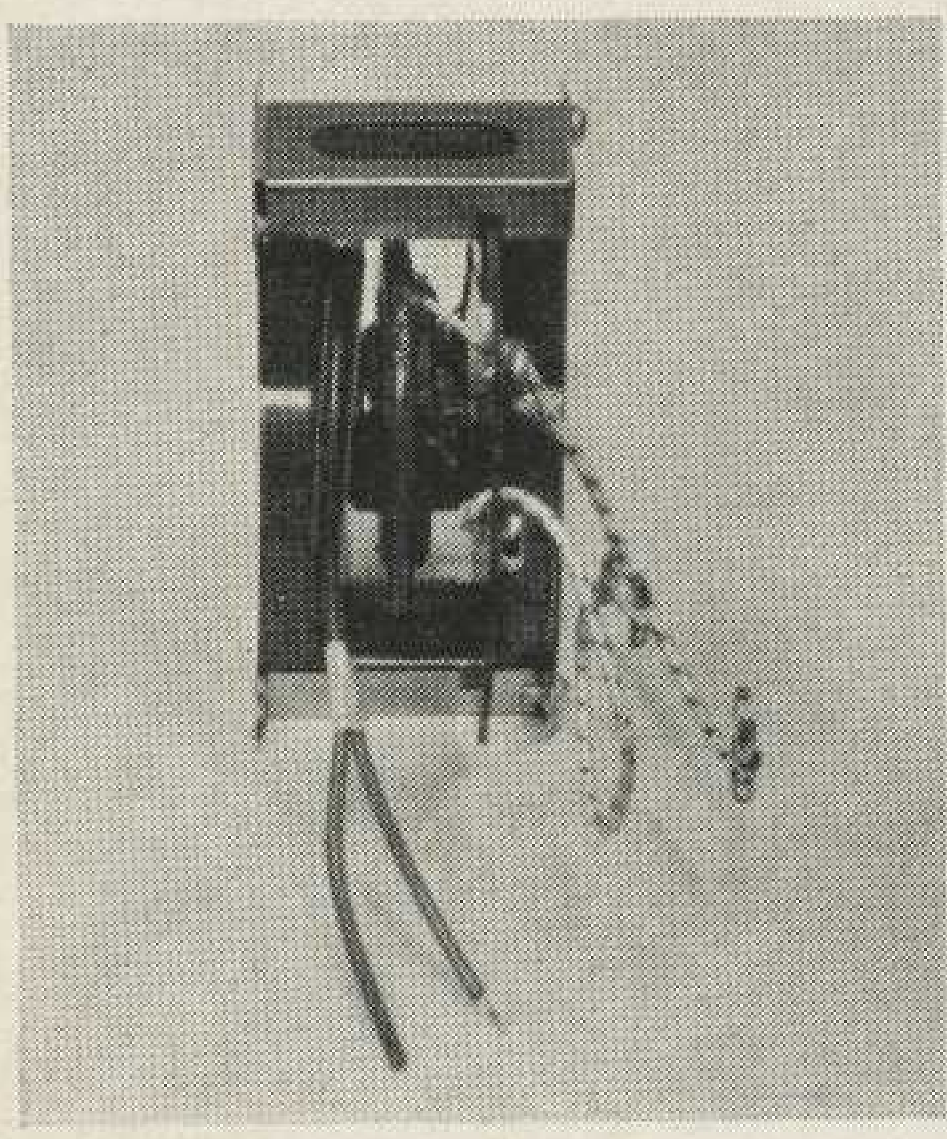


Fig. 11—Side view of a horizontal-deflection transformer with molded sponge-iron-powder core.

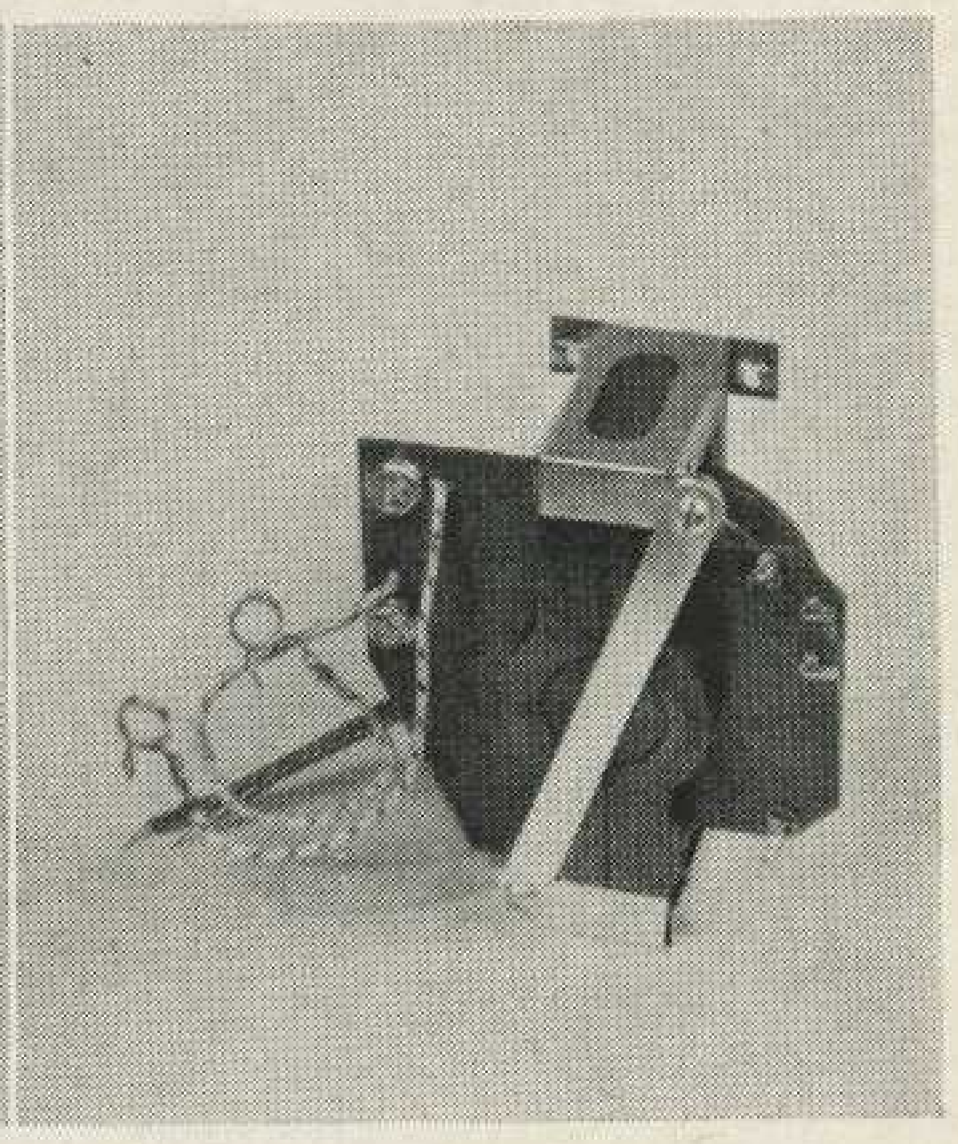


Fig. 12—Top view of horizontal-deflection transformer with molded sponge-iron-powder core.

Figures 13, 14, and 15 are three views of a production type deflection yoke. It will be noticed, in Fig. 13, that this particular yoke has an iron wire-wound magnetic core. The iron-dust-sector core may be substituted directly in this identical yoke arrangement in place of the iron-wire core. Fig. 16 shows a developmental sample yoke with an iron-dust core in place.

In present production the cores are being molded in the 15 to 30 tons per square inch pressure range. Depending upon economic factors, higher pressures may later be used.

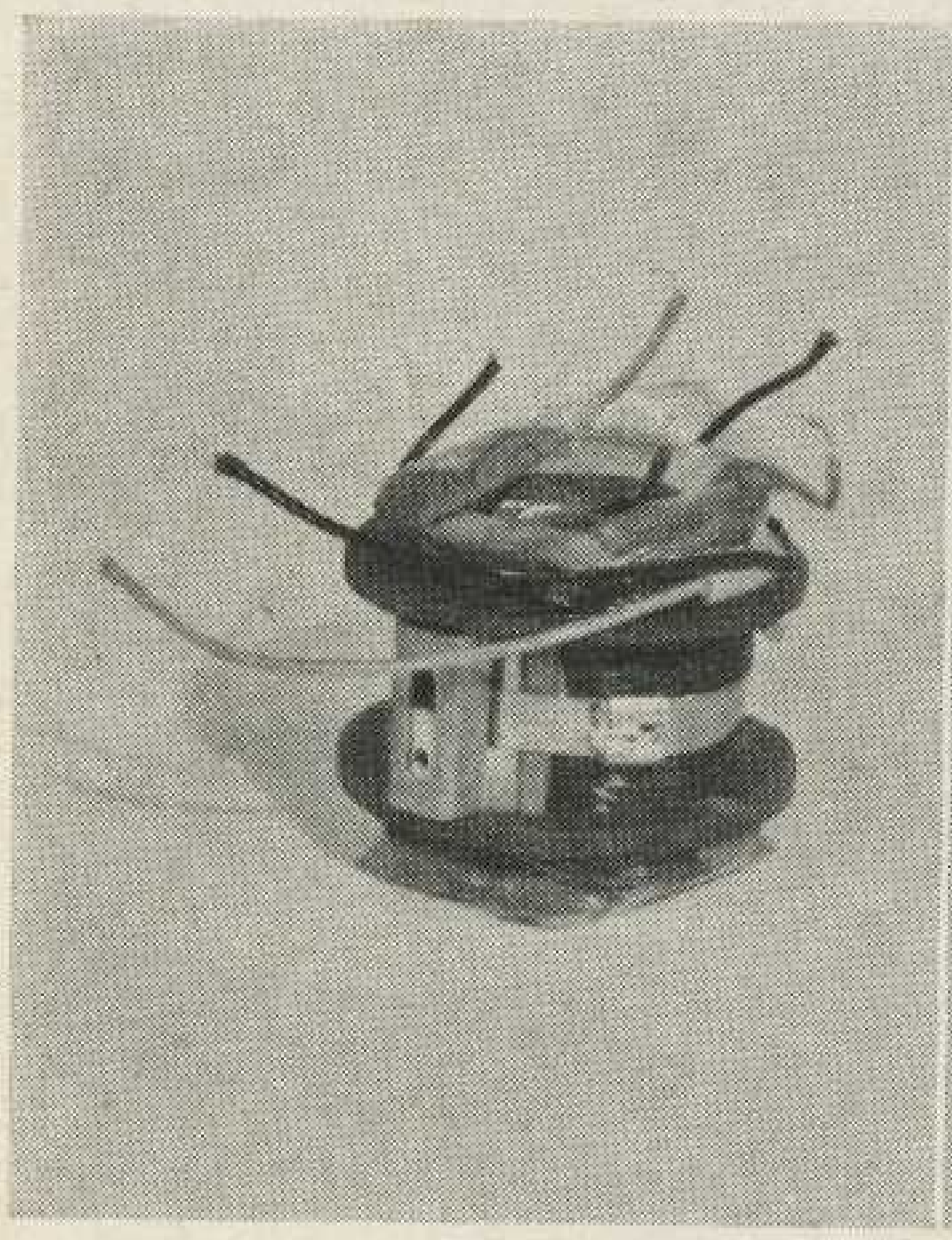
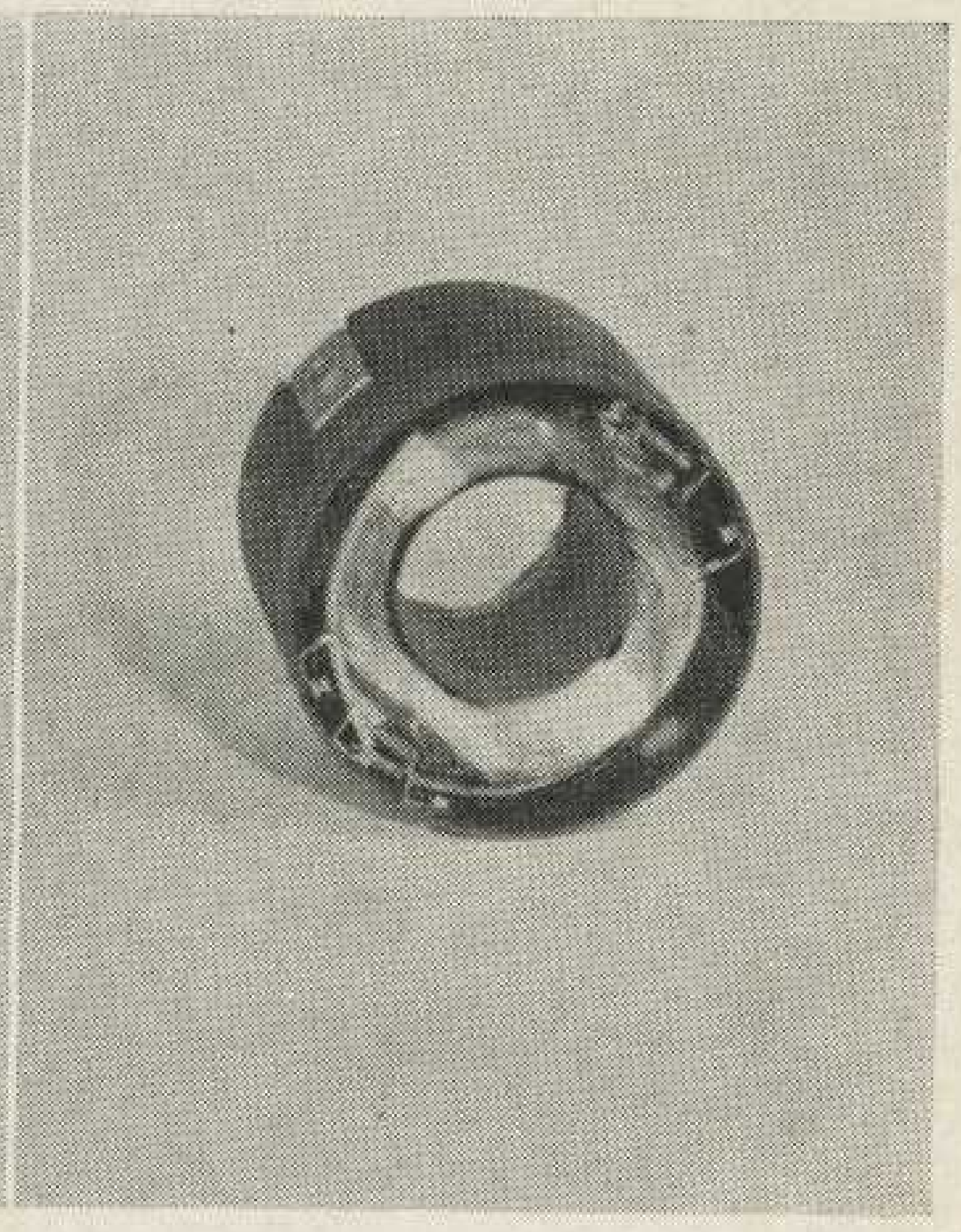



FIG. 13—Unmounted deflection yoke with iron wire-wound core. A molded iron-dust core may be used on this same yoke without other alterations.

FIG. 14—Deflection yoke of Fig. 13 with terminals and side cover in place.


FIG. 15—Complete deflection yoke with all covers in place.

FIG. 16—First developmental sample deflection yoke with molded iron-dust core in place.

The curves of Figs. 5, 6, and 7 indicate the apparent incremental or a-c permeability μ_a as a function of the maximum magnetic flux density B_{max} for samples of sponge-iron core materials molded under pressures of 15, 30, 45, and 60 tons per square inch, respectively. The maximum values of permeability occurred in each case at values of B_{max} about ten times that of the normal working range. At 15 tons per square inch, the maximum permeability is approximately one and one half times that of the normal working range. The 60-ton-per-square-inch sample had a peak permeability almost twice that attained at lower values of B_{max}.

At the horizontal scanning frequency (15,750 cycles per second) the apparent permeability μ_a of the 15 tons to the square inch sponge-iron material increases from 59 to 70 as B_{max}, increases from 200 to 600 gauss (peak). The 30 tons per square inch sample exhibits an apparent permeability μ_a varying from 58 to 78 in the same range of maximum flux density. Pressures in this range are employed in making the present production type iron-dust cores for television horizontal deflection transformers, the operating levels of which are within the above stated range of maximum flux density B_{max}. The operating incremental permeability is approximately 63. Molding pressures of 45 and 60 tons per square inch yield operating incremental permeabilities of approximately 90 and 110, respectively.

When these materials, molded at pressures up to about 30 tons per square inch, are used, it is noted that, within the normal working range of flux density (Figs. 5 and 6), the large equivalent series air gap is sufficient to mask a large part of the effect of the variation of the true permeability of the iron particles with changes in the magnetomotive force. Therefore, the apparent incremental (or a-c) permeability measured at quite low flux densities is of the same order of magnitude as the working value at somewhat higher levels. In fact, the d-c magnetizing flux may cancel a portion of this increase. Within the limits of the required engineering accuracy, the operating value of apparent a-c permeability may be corrected from the value determined by low level measurements to the operating value, by reference to the curves. Values determined by means of a standard laboratory type bridge or Q-meter may thus be utilized in calculations involving normal operating flux densities. Such corrections may involve considerable error if applied to powdered materials molded at higher pressures or when laminated iron cores of either the stacked or wound type are considered.

The higher apparent permeabilities attained at higher values of peak magnetic flux density B_{max} are *not* available for use in horizontal deflection transformers on account of inductance value and winding requirements. The higher losses incurred at the higher frequencies involved in this application also make it impossible to utilize more than a small fraction of the maximum indicated permeability of practical laminated core materials. For instance, in the case of one type of material of 0.003 inch lamination thickness, the apparent operating permeability range, for horizontal deflection transformer uses, is between 200 and 300, depending upon the precise operating conditions.

On account of the small variation of inductance with operating level and the small energy losses produced when molded iron-powder cores are used, it is permissible to make certain assumptions regarding the equivalent circuit of the horizontal deflection system. They may be checked by simple measurements, with considerably more assurance than is possible when other core materials are used.

III. THE DEFLECTING CYCLE

In the horizontal deflection of the electron beam in television picture tubes, there are two major periods, that of the trace and of the retrace.

During the trace period, the electron beam should trace a single horizontal line of the picture at a constant rate of movement. To produce this effect, the current flowing through the horizontal deflecting windings of the yoke must change its magnitude at a constant rate with respect to time.

During the retrace period, the beam is blanked out, and the current may return to the state existing at the start of the trace period by following any conceivable law of change. At present, it seems that the most practical function is that representing a half-cycle of a cosine wave at the natural resonant frequency of the deflecting system. There may also be waves of small amplitude and of higher frequency superimposed upon the main retrace wave. They are caused chiefly by the leakage reactances in combination with shunt capacitances, resonating at higher frequencies.

IV. THE DEFLECTING CIRCUIT

Assuming this simple picture to represent the major phenomena involved in the current variations during the deflecting cycle, it is possible to impose these currents upon a simplified equivalent circuit of the deflecting system.

Figure 17 illustrates a schematic diagram of a common deflecting system. It is to be noted that a saw-tooth voltage-wave is developed across the discharge capacitor C_1 by the discharge action of the tube V_1. There is in addition a negative voltage pulse, developed during the discharge period, added at the negative peak of the voltage saw-tooth wave.

In the active deflecting portion of this circuit, the driver tube V_2, is a beam tetrode controlled by the above described voltage wave. This tetrode has its screen-grid by-passed to the cathode. The plate load arm is essentially a low-loss inductance coupled to V_2 by means of an inductance changing device, the horizontal deflection transformer T.

Tube V_3 is a part of a damper system which is required by low-loss deflection systems to control the oscillation begun at the start of the retrace period, and to return as much as possible of the stored energy, in the resonant system, to active use in the next and succeeding trace periods. V_3 is usually a high perveance triode, or perhaps, for economic reasons, a diode.

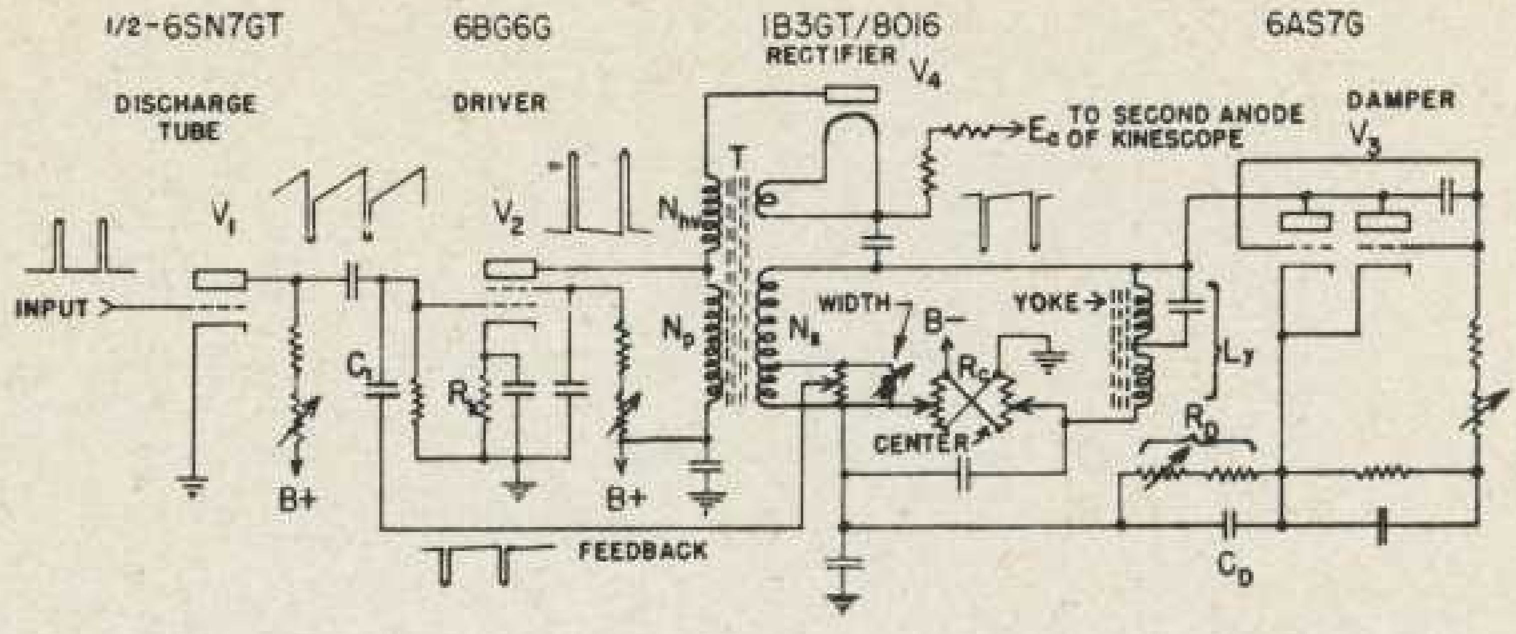


FIG. 17—Circuit diagram of a standard type horizontal deflection system utilizing a triode damper tube V_3 and a damper load resistor R_D. Voltage wave forms are indicated.

A simplified equivalent circuit is presented in Fig. 18. The deflection transformer T is represented as an equivalent "T" network with the deflection yoke inductance L_y converted to its equivalent transformed value $[(L_p / L_s) L_y]$. It will be noted that all dissipative elements have been neglected, thus presupposing the use of moderately low-dissipation circuit elements.

For successful performance in the present standard television system, the natural resonant frequency must be greater than 69 kc and may be designed to be about 77 kc, in order to insure a sufficiently short retrace period (one half cycle of this resonant frequency). The standard sweep recurrence rate (15,750 cycles per second) is roughly one fifth of the desired resonant frequency, so it is possible in the first approximation to neglect the distributed capacitances in considerations of the equivalent circuit during the sweep or trace period. These approximations have been applied to produce the simplified equivalent circuit of Fig. 18.

Upon admitting these assumptions, it may be shown that the equivalent plate-load inductance is given by

$$(1) \qquad L_b = L = L_p \left[1 - \frac{K^2}{1 + (L_y / L_s)} \right]$$

where: L_b = plate-load inductance of driver tube V_2

L = inductance between the input terminals of the deflection-system network

L_p = transformer primary winding inductance

L_s = transformer secondary winding inductance

$K = M / \sqrt{L_p L_s}$ = coupling coefficient of transformer

M = mutual inductance between primary and secondary of transformer

L_y = inductance of the horizontal deflection-windings yoke

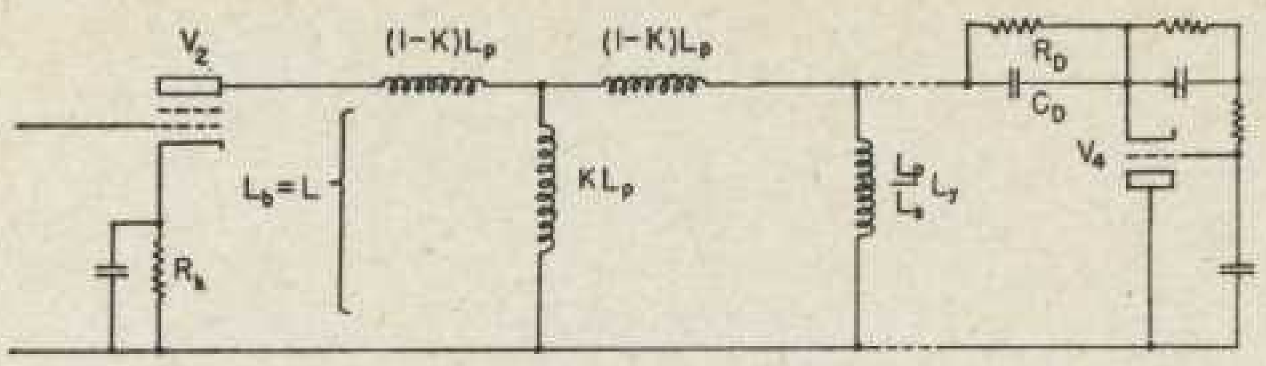


FIG. 18—Simplified equivalent schematic diagram of a low-loss horizontal-deflection system. The damper circuit components must be assumed to be converted to equivalent primary-circuit values.

It is obvious that the full input deflecting ability which might be obtained from a directly connected deflecting yoke of the correct inductance L is not available in a circuit utilizing a coupling transformer. The series leakage inductances $[(1\text{-}K)L_p]$ and the shunt magnetizing inductance $[KL_p]$ unite to form an attenuating network between the driver tube V_2 and the deflecting-yoke windings.

Suppose that a practical deflecting yoke is available and that a low-loss inductance-changing or "deflection" transformer is constructed to provide the desired inductance ($L_b = L$) to match the characteristics of the driver tube V_2. This combination of transformer and yoke windings with the driver tube V_2 possesses a certain electron beam deflecting ability measured in terms of the beam deflecting angle $2\,\theta_H$ for a beam accelerating voltage E_a. Consideration of only the deflecting yoke operation yields the equation

$$\theta_H = \sin^{-1} (\alpha s N_y \mathrm{i_y} / 2 \sqrt{E_a}) \tag{2}$$

where: θ_H = *half* the total deflection angle (degrees)
α = a constant depending upon the geometry of the yoke magnetic path
s = the equivalent yoke length (centimeters)
N_y = the total number of turns on the yoke horizontal windings
$\mathrm{i_y}$ = the *peak-to-peak* value of the yoke-winding current (amperes)
E_a = the electron beam accelerating potential (volts)

Suppose that a special type of yoke might be designed to provide the correct load inductance L_b *without* the use of a matching transformer, but with the same values of α, s, and E_a, so that $L_y = L_b = L$. Assuming that the same driver tube is employed, it is found that the deflection angle is increased due to removal of the transformer attenuating factors. From Eq. (2), it is apparent that this is equivalent to increasing the number of yoke peak-to-peak ampere-turns.

It can be shown, from the simplified equivalent circuit, of Fig. 18, that the ratio of the yoke peak to peak ampere-turns, with a practical transformer, to that value with no transformer (or with an *ideal* transformer) is equal to the deflection factor

(3)

$$F_D = \frac{(N_y \mathrm{i_y})\ \text{(practical transformer)}}{(N_y \mathrm{i_y})\ \text{(ideal transformer)}} = 1 \Big/ \sqrt{\frac{L_s}{L_y}\left[\frac{1}{K^2} - 1\right] + \left[\frac{2}{K^2} - 1\right] + \frac{1}{K^2 L_s / L_y}}$$

The maximum deflection factor F_D is generally desirable and may be determined as a function of K and the ratio (L_s/L_y) from the equation

(4a) $$(L_s/L_y) = 1/\sqrt{1 - K^2}$$

which yields

(4b) $$F_D = \sqrt{\frac{(L_s/L_y) - 1}{(L_s/L_y) + 1}}$$

A family of curves representing (3) together with a dashed plot of (4) is shown in Fig. 19. If the approximate value of the coupling coefficient for a particular type of transformer windings and core is known, then the optimum value of (L_s/L_y) and the corresponding value of F_D may be determined from the curves of Fig. 19.

As an example, it is known that, for a certain winding and core structure arrangement, $K = 0.94$. Hence, from the dashed curve for maximum F_D values it is seen that $(L_s/L_y) = 2.92$. From the deflection factor scale, at the left edge of the chart, $F_D = 0.702$. In other words, the best transformer which can be constructed with a coupling coefficient of 0.94 will yield 70.2 per cent of the deflection ampere-turns which could be obtained with a perfect transformer.

Regardless of cost, it is difficult to build a satisfactory horizontal deflection transformer having a deflection factor greater than about 0.85. When a kinescope second-anode voltage supply is to be operated from the same transformer, it is even more difficult.

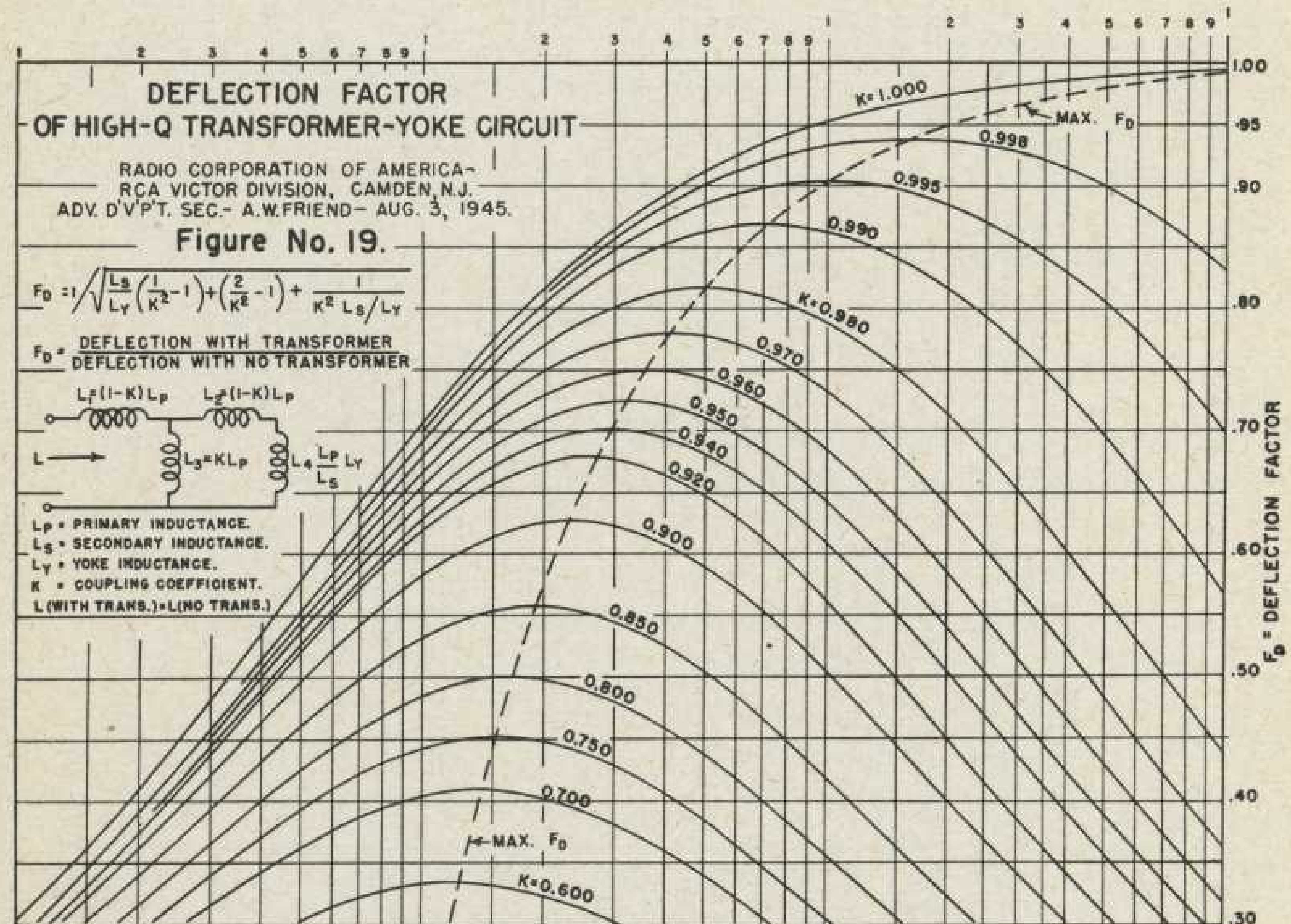


FIG. 19—Deflection factor F_D versus secondary to yoke inductance ratio (L_s/L_y) as a function of transformer coupling coefficient K.

V. DESIGNING THE HORIZONTAL DEFLECTION TRANSFORMER

There is a number of viewpoints from which one may begin in the design of a horizontal deflection transformer, depending upon the originally known factors. Generally, an intended driver tube (or tubes) and a damper tube are assumed.

The maximum inductance of the yoke may be limited by the damper tube peak-inverse voltage rating. From experience, or from calculations, it may be determined that a certain yoke inductance is approximately the maximum which may be used within the peak voltage rating of a given damper tube while deflecting an electron beam of known characteristics.

Suppose that a type 6AS7G damper triode is to be used. This tube has a peak inverse pulse rating of 1750 volts, according to its present design characteristics. The maximum yoke inductance across which this tube may be utilized in the present average application is approximately 10 millihenries. A production-type yoke is available with an inductance of 8.2 millihenries.

The deflection characteristics of a developmental sample yoke of this inductance have been determined by experiment and plotted. It is convenient to determine the product αs of Eq. (2) for a particular type yoke structure. With a given number of turns and a certain required value of

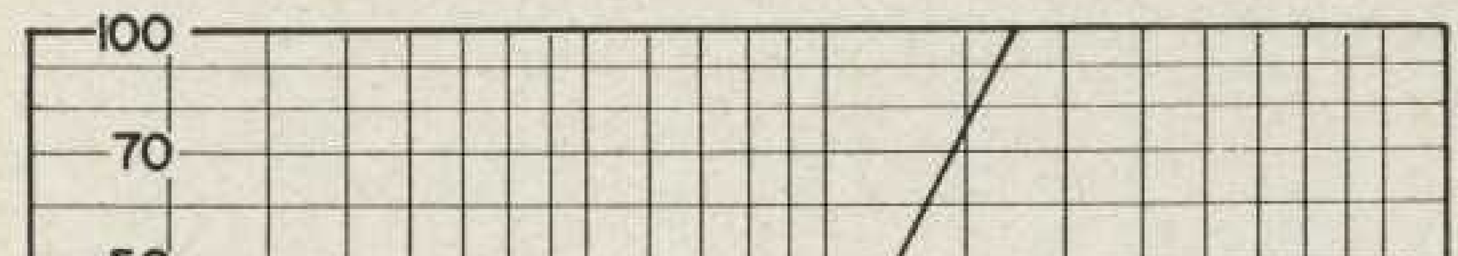


FIG. 20—Peak-to-peak horizontal deflection-yoke current versus kinescope second anode (accelerating) potential for a developmental sample yoke with a molded iron-powder core.

maximum horizontal deflection angle $2\theta_H$ it is possible to plot the peak-to-peak yoke current versus the electron beam accelerating voltage E_a. These values for the particular developmental sample yoke are plotted in Fig. 20. This curve represents the results obtained with a yoke employing a molded-iron-dust yoke core. Figure 21 is a plot of Q values of several deflection yokes as a function of frequency. The amazing increase in Q when the iron-dust core is used is notable.

In the transformer design sequence, the peak-to-peak yoke current is selected from the curve of Fig. 20 for the full deflection of a normal 4/3 aspect ratio picture of a kinescope utilizing 50 total deflection angle 2θ for the picture diagonal, or a horizontal deflection angle of $2\theta_H = 40.92°$.

The time of one horizontal sweep cycle τ_H is the reciprocal of the horizontal sweep frequency f_H.

$$\tau_H = 1/f_H \tag{5}$$

According to the present television standards, $\tau_H = 1/15{,}750 = 63.5 \times 10^{-6}$ seconds or 63.5 microseconds. Horizontal blanking is allowed for 16 to 18 per cent of this time. After allowing for tolerances, there is a remainder of not more than about 11.5 per cent or 7.3 microseconds maximum time for the retrace period. Conservatively, it is wise to allot about 6.5 to 7.0 microseconds for the return time, in making design calculations. The retrace time τ_r is

$$\tau_r = 1/2f_r \tag{6}$$

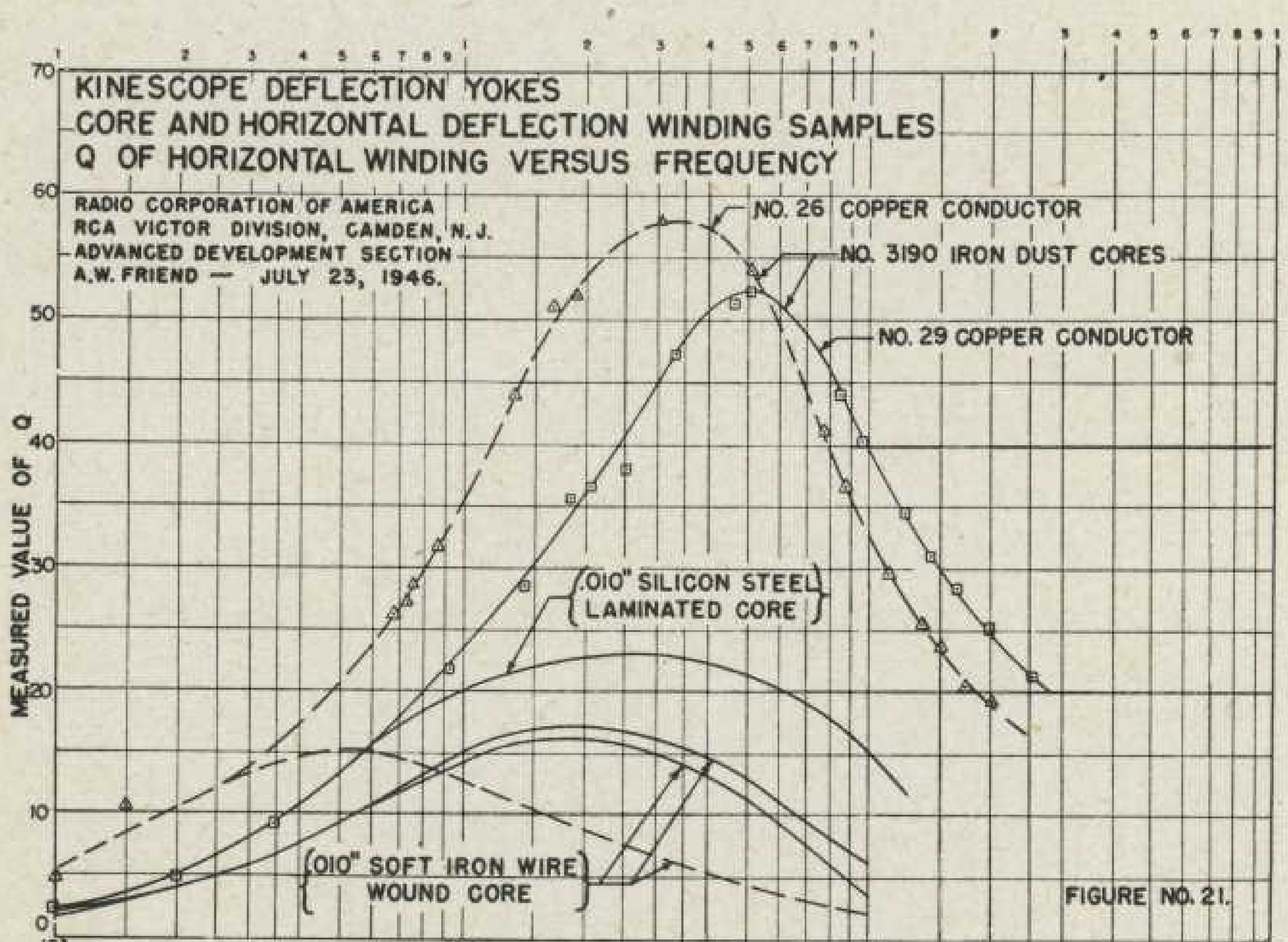


Fig. 21—Q of kinescope deflection yoke samples, with various cores and wire sizes, as a function of frequency.

where f_r is the free resonant frequency of the system. The trace time is

$$\tau_t = \tau_H - \tau_r = \frac{1}{f_H} - \frac{1}{2f_r} \tag{7}$$

Hence, the trace time τ_t should be about 56.5 to 57.0 microseconds.

To reproduce an acceptable picture, the rate of change of current through the deflection yoke windings during the trace period must be constant to within about five percent, when comparing any two portions of the trace. Within this limit, it is admissible to speak of the rate of change of yoke current during trace as a constant. The required peak to-peak yoke current $\mathbf{i}_y$ is taken from the deflection-current curve, or data, and the trace time τ_t is determined by (7). the rate of change of current during trace is

$$(di_y/dt)_t = \mathbf{i}_y / \tau_t \tag{8}$$

To allow for picture width-control operation and for variations in component parts and tubes, there should be an increase of about 10 per cent in $\mathbf{i}_y$ above the value indicated by the curve of Fig. 19.

The required d-c voltage E_{Lyt} applied across the inductive component of the yoke during the trace time may be determined from the equation

$$E_{Lyt} = L_y \, (di_y/dt)_t \tag{9}$$

When the saw-tooth current through the deflection yoke windings reaches its peak value it is being supplied by transformer action from the driver tube V_2. At this moment V_2 is suddenly and completely cut off, allowing the deflecting system to begin a period of free oscillation at its resonant frequency of about 70 kc. The rapid change of magnetic flux may induce perhaps 1500 peak volts or more across the yoke windings.

The oscillation begins with maximum current. At the end of the first half cycle, the yoke current reaches a new maximum value in the reverse direction. At this instant the induced yoke voltage reverses polarity and initiates damper-tube V_3 conduction, to begin the next sweep cycle. The current at this time is less than that at the start of oscillation due to circuit damping. It may be shown that the effects of damping yield a ratio of current defined by the current factor

$$F_I = i \,/\, i_{(Q_r = \infty)} = \varepsilon^{-n\pi/Q_r} = \varepsilon^{-\pi/2Q_r} \tag{10}$$

where: F_I = current factor

i = current with the circuit of measured value of Q_r

$i_{(Q_r = \infty)}$ = current with circuit of infinite Q_r

ε = 2.71828

π = 3.1416

n = number of cycles after start of damped wave = ½ cycle

Q_r = 2π (energy stored)/(energy lost per cycle at resonant frequency)

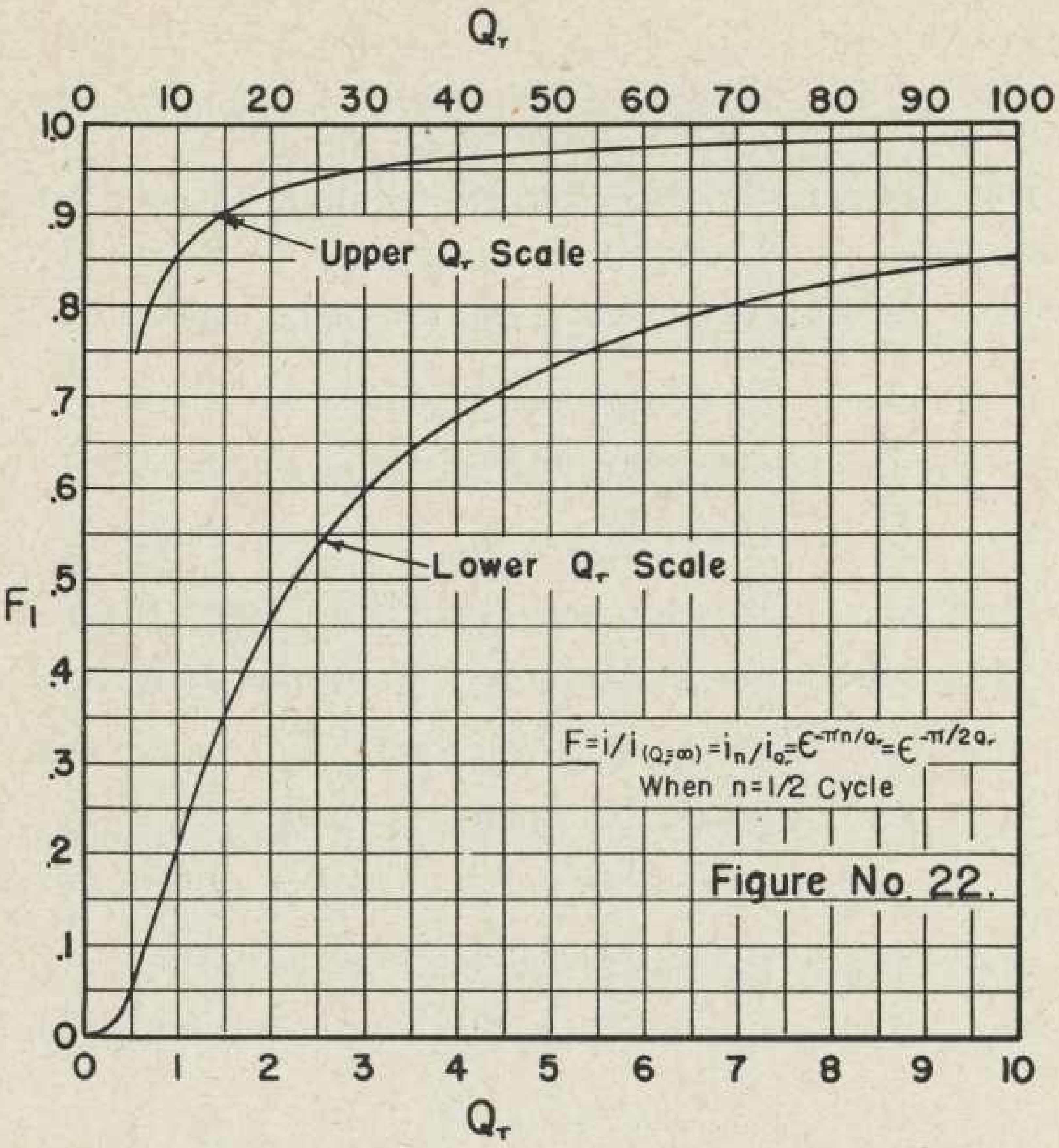


FIG. 22—Current factor F_I as a function of the Q of the system at its free resonant frequency f_r.

This function is plotted in Fig. 22 for values of Q from zero to 100. A value of $Q_r = 15$ is common in low-loss deflection systems. From Fig. 21, this yields $F_I = 0.902$.

During the first part of the trace period the yoke current is passed through the damper tube V_3 to an energy storage capacitor C_D. This damper load capacitor is made large enough to substantially stabilize the voltage into which the damper tube operates plus a uniform portion of the plate-voltage drop across the damper tube provides the fixed potential across which the inductive component of the yoke windings must work. The loss factors in the yoke require a saw-tooth voltage component which is supplied by the action of an *RC* network in the grid-cathode circuit of the damper triode reacting upon the triode plate-circuit characteristics.

As the energy stored in the magnetic circuit diminishes to the point of providing an inadequate rate of current change in the deflection-yoke windings, the driver tube V_2 begins to conduct. It carries its portion of the deflection burden to the end of the trace period. The damper tube should conduct only very slightly during this period, mostly during the initial or transitional phase.

It is important to the designer to know at what point in the trace period the burden of producing the change of yoke current is handed over from the damper system to the driver system. It is evident that the period of damper action dominance must be less than half the trace time, since the energy involved is only that stored in the magnetic fields at the end of trace minus the amount deducted to satisfy the system losses during the retrace period. The transition time may be defined as that time at which the projected linear rate of yoke current change passes through the zero-current axis.

In view of (3) the final yoke current may be designated as

(11) $$i_{yf} = \mathrm{i_p}\, F_D\, a_{ps}$$

where: $\mathrm{i_p}$ = peak driver plate current
F_D = deflection factor
a_{ps} = the primary to secondary transformer turns ratio = N_p/N_s

From (10) and (11), the initial yoke current is found to be

(12) $$i_{yi} = \mathrm{i_p}\, F_D\, F_I\, a_{ps}$$

By simple proportion, and in view of (11) and (12), it may be shown that the fraction of the trace time allotted to damper dominance is

(13) $$n_D = i_{yi}/(i_{yi} + i_{yf}) = F_I/(1 + F_I)$$

Alternately, the fraction allotted to the driver period is found to be

(14) $$n_d = i_{yf}/(i_{yi} + i_{yf}) = 1/(1 + F_I)$$

so that

(15) $$n_D + n_d = 1$$

From (14), if $F_I = 0.902$, the active driver portion of trace is $n_d = 1/1.902 = 0.526$, or 52.6 per cent.

The time of each conducting cycle of the driver tube V_2 is

(16) $$\tau_d = \tau_{tnd}$$

These data with the value of peak plate current allow the calculation of the rate of change of the primary current of the transformer. That current is also the driver-tube peak plate current.

(17) $$(di_p/dt)_t = \mathrm{i_p}/\tau_d$$

The probable maximum peak plate current for one type 807 or for one 6BG6G is 0.220 amperes.

From (3), (8), and (17), the required transformer turns-ratio may be shown to be

(18) $$a_{ps} = N_p/N_s = (di_y/dt)_t\,/\,(di_p/dt)_t F_D = \mathrm{i_y}\,\tau_d/\mathrm{i_p}\tau_t F_D$$

To determine the actual number of turns required from the primary and secondary windings one must know the geometry of the core structure and the permeability of the core material under the operating conditions. Alternately the operating permeance P of the proposed core structure may be determined by measurement. It is convenient to utilize the permeance in combination with the constants of the equation for the inductance

(19) $$L = 0.4\,\pi\, N^2 P \times 10^{-8} = N^2\,\beta$$

where: L = the inductance of the winding, in henries
β = $0.4\pi P \times 10^{-8}$
P = the permeance of the core structure

In the case of the production type transformer of Figs. 11 and 12, the value of β for the assembled core structure should be approximately equal to 2.70×10^{-7}.

From (4), or from Fig. 19, the value of (L_s/L_y) was determined in terms of the coupling coefficient K. Since the yoke inductance L_y was assumed to be known, it follows that the secondary inductance is

$$L_s = (L_s/L_y)\, L_y \tag{20}$$

From (18), the number of transformer secondary turns is

$$N_s = \sqrt{L_s/\beta} \tag{21}$$

and from (17), the number of transformer primary turns is

$$N_p = a_{ps}\, N_s \tag{22}$$

The primary inductance L_p, from (19) and (21), is

$$L_p = N_p^2\, \beta = a_{ps}^2\, N_s^2\, \beta = a_{ps}^2\, L_s \tag{23}$$

The voltage across the inductive component of the driver-tube load during trace may be determined if the value of this inductance is known. Equation (1) may be solved in terms the results of Eqs. (4) and (23).

From the plate load inductance and the rate of change of plate current $(di_p/dt)_t$ during the trace period, the inductive voltage drop across the plate load is

$$E_{Lb} = L_b\, (di_p/dt)_t \tag{24}$$

The value of E_{Lb} is the voltage drop across only the inductive component of the load. There is a saw-tooth voltage wave caused by the saw-tooth current wave passing through the equivalent resistance component. The energy loss is directly proportional to the amplitude of this voltage wave, which should therefore be minimized. A rigorous solution for the value of the equivalent resistance may involve the solution for an equivalent value for each harmonic frequency of the saw-tooth wave.

A short-cut method, generally, of sufficient accuracy, involves the assumption of a sine-wave sweep of the fundamental scanning frequency and of the saw-tooth amplitude together with a system having an energy storage factor Q_1 evaluated at the same frequency. From the known plate load inductance, frequency and Q_1 values, the equivalent resistance is

$$R_1 = 2\,\pi f_H\, L_b/Q_1 \tag{25}$$

From the above short-cut method, the dissipative peak voltage drop across the load is

$$e_{Rb} = i_p\, R_1 = 2\pi f_H L_b i_p/Q_1 \tag{26}$$

The minimum driver-tube plate voltage occurs when the voltage across the load is maximum. In view of allowable screen power dissipation, the minimum plate voltage e_{pmin} for either the 807 or 6BG6G should be *not less than* 60 volts. The total required plate power supply voltage E_b is the sum of these values plus the voltage drop E_k across the driver tube cathode resistor R_k and that E_c across the picture horizontal centering control R_c. Hence

$$E_B = E_{Lb} + e_{Rb} + E_k + E_c + e_p \tag{27}$$

This value may be used as a value of power supply voltage.

It should be noted that this total power supply voltage is not directly applied to the plate of the driver tube as d-c. The plate voltage wave should be a saw-tooth with a minimum of more than 60 volts and a maximum of perhaps 125 volts when a 6BG6G driver tube is used in a low-loss system. Superimposed on this wave is the half-sine-wave retrace pulse of several kilovolts peak positive amplitude.

VI. THE AMPLITUDES OF THE RETRACE VOLTAGE PULSES

The amplitude of the retrace voltage-pulse across the horizontal deflection windings of the yoke may be calculated from the theory of damped free oscillations. It may be shown that with a starting current i_o, the damped voltage across the LR branch of an LCR circuit is represented by

$$e = i_o \left(L\omega_r + \frac{R}{4Q_r}\right) \left(\sin\omega_r t\right) \varepsilon^{-\omega t/2Q_r} \tag{28}$$

Here, Q_r is the resonant Q of the system and

$$\omega_r = \pm \sqrt{\frac{1}{LC} - \frac{R^3}{4Q_r}} = 2\pi f_r \tag{29}$$

where f_r is the resonant frequency.

The maximum value of the first half cycle of this wave is the peak value of the major retrace wave across the yoke, when $i_o = i_y n_d$ and $L = L_y$. From (28) this maximum is

$$e_{yr} = i_y n_d \left(L_y\omega_r + \frac{R}{4Q_r}\right)\left[\sin\left(\tan^{-1} 2Q_r\right)\right] \varepsilon^{-(\tan^{-1} 2Q_r)/2Q_r} \tag{30}$$

A pulse factor F_p may be defined as the ratio of the peak voltage with any value of Q_r to that voltage when Q_r is infinite. When Q_r becomes infinite (30) becomes

$$e_{yr}\big]_{(Q=\infty)} = i_y\, n_d\, L\, \omega_r \tag{31}$$

So

(32)

$$F_p = \left[e_{yr}/e_{yr(Q=\infty)}\right] = (1 + 1/4Q_r^2)\left[\sin\left(\tan^{-1} 2Q_r\right)\right] \varepsilon^{-(\tan^{-1} 2Q_r)/2Q_r}$$

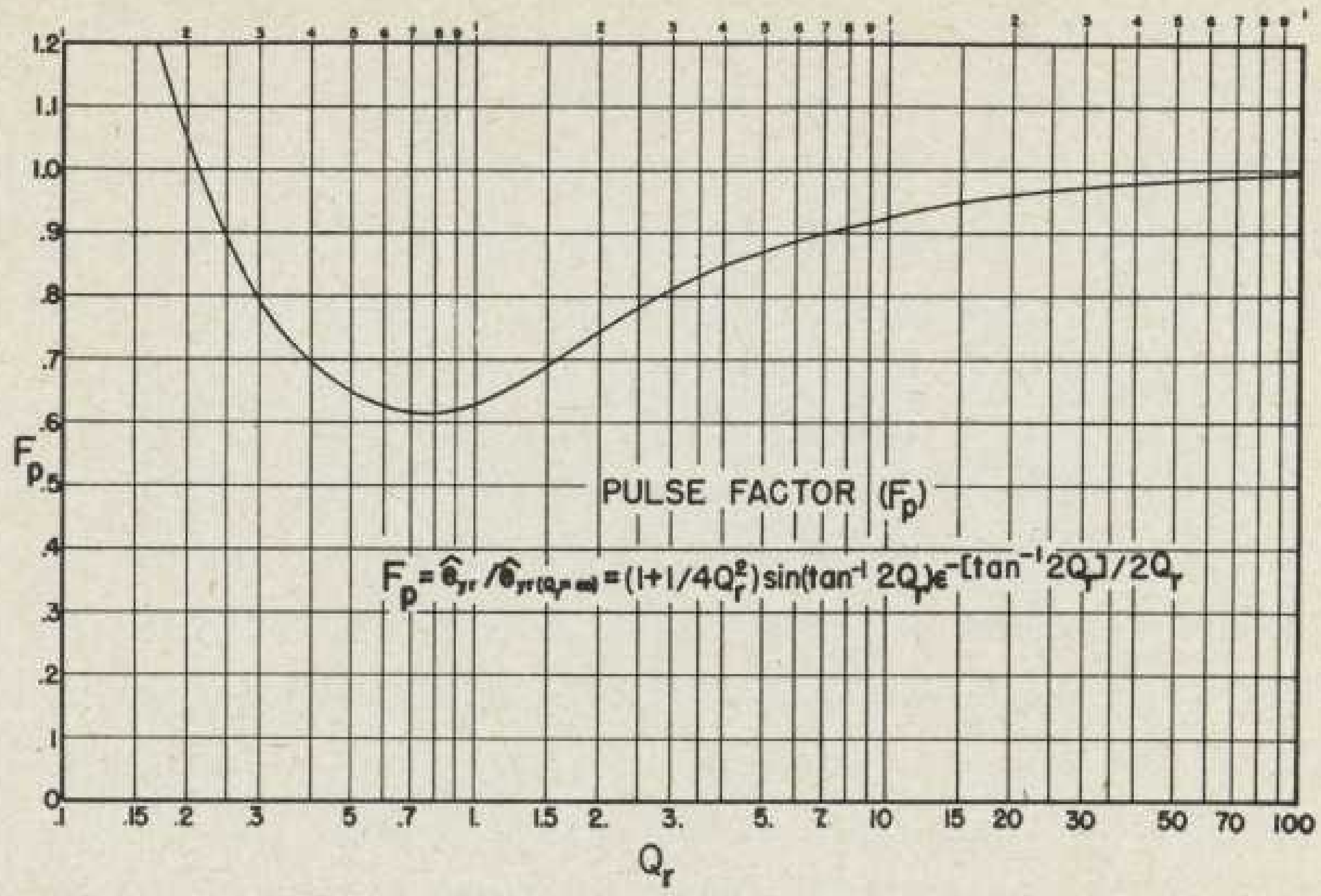


FIG. 23 —Pulse factor F_p as a function of the Q of the system at its free resonant frequency f_r.

The curve of F_p versus Q_r is plotted in Fig. 23 for values of Q_r from 0.1 to 100. For values of Q_r less than unity the value of F_p is chiefly of academic interest, since the resonant frequency f_r is depressed by the increased R/L ratio (see Eq. (29), requiring decreased inductance L to maintain the correct value of resonant frequency f_r and retrace time t_r. Hence, the rising F_p curve below $Q_r = 0.8$ does *not* imply an increase of retrace pulse voltage when the Q is reduced.

In view of (31) and (32), the yoke retrace pulse voltage Eq. (30))becomes

(33) $$e_{yr} = i_y\, n_d\, L_y\, \omega_r\, F_p = 2\pi f_r\, i_y\, n_d\, L_y\, F_p$$

The peak damper-tube retrace voltage e_{Dr} may be determined by adding the peak yoke retrace voltage e_{yr} and the damper load voltage E_{RD}. Since E_{RD} is equal to the yoke trace voltage minus the damper tube voltage drop during trace e_{Dt}, it follows that

(34) $$\overline{E}_{RD} = E_{Lyt} - e_{Dt}$$

In practice, e_{Dt} may be perhaps 10 to 40 volts in present systems. From this reasoning, in view of (34),

(35) $$e_{Dr} = e_{yr} + \overline{E}_{RD} = e_{yr} + E_{Lyt} - e_{Dt}$$

The peak of the major pulse of the driver tube plate voltage during retrace e_{dr} may be determined by the equation

(36) $$e_{dr} = e_{yr}\, a_{ps}\, K + \overline{E}_B - E_K - E_C$$

The leakage reactance resonance wave may add perhaps 5 to 10 per cent of this value.

VII. SECOND ANODE HIGH VOLTAGE POWER SUPPLY

The accelerating or second-anode voltage E_a for the kinescope may be derived from the retrace voltage pulse of the deflection system. A known value of E_a was assumed as a starting point in the transformer design data. With the presently available vacuum tubes and components, it seems impractical to attempt to obtain more than about 12 kilovolts peak pulse from a deflection system transformer while providing a linear rate of picture spot deflection, and while operating within the vacuum-tube ratings. A voltage-multiplier-rectifier system (Fig. 24) is required to produce higher voltages. Experience with the distributed capacitances encountered in such devices indicates that 18 kilovolts is near the limit which may be obtained with a voltage doubler system under these conditions. This limit is imposed by driver and damper tube peak-voltage ratings and by retrace time requirements.

An extra transformer winding is required to produce the additional voltage required for this purpose. This winding of N_{hv} turns is connected as an addition to the primary winding to produce an auto-transformer arrangement.

Connection of the rectifier system across the primary N_p, the deflection secondary N_s and the high voltage N_{hv}, all in "series-aiding" connection, yields a winding of total turns N_t, and maximum output voltage e_{tr} The coupling coefficient K_{ts} indicates the fractional coupling between the secondary N_s and the total N_t winding. The number of voltage multiplier stages may be designated as m and the total of secondary turns ratio as

$$(37) \qquad a_{ts} \equiv N_t/N_s = (N_p + N_s + N_{hv})/N_s$$

Assuming the free resonant frequency of the system to remain the same as previously assumed, the maximum value e_{tr} of the major retrace pulse across the entire winding may be found in a manner similar to that of (36). Adding to e_{tr}, times the multiplying-stage factor m, the proper d-c voltages yields the value of the no-load second-anode voltage

$$(38) \qquad \overline{E}_{ao} = e_{yr}\, a_{ts}\, K_{ts}\, m + \overline{E}_B - E_K - E_C$$

Combining (36) and (37) and solving for the number of turns on the high-voltage winding produces

$$(39) \quad N_{hv} = N_s \left\{ \left[(\overline{E}_{ao} - \overline{E}_B + E_K + E_C)/K_{ts}\, m e_{yr} \right] - 1 \right\} - N_p$$

On account of the added leakage-reactance resonance voltage, the calculated no-load voltage is very nearly that which should be attained with the average kinescope load during normal operation.

VIII. THE AVERAGE CURRENT AND POWER

The average plate current of the driver tube may be determined with reasonable accuracy, without the use of an elaborate integration method, by computation from the geometry of an assumed current-time plot. A triangular conduction current-time area may be assumed with a 15% current addition to account for the additional current due to lack of the ideal sharp current cut-off assumed by the theory. From these data it may be shown that the average plate current is

$$\bar{I}_p = 1.15\ \mathrm{i_p}\ \tau_d\ /\ 2\tau_H = 0.575\ \mathrm{i_p}\ \tau_d\ /\ \tau_H$$

$$(40) \qquad = 0.575\, \mathrm{i_p} f_H \,/ \left(\frac{1}{f_H} - \frac{1}{2f_r}\right) n_d$$

The energy dissipation on the plate of the driver tube is quite low on account of the low plate voltage during the conduction period. The plate power dissipation in watts P_p may be closely approximated by considering the peak saw-tooth voltage $\mathrm{e_{Rb}}$, the minimum plate voltage $e_{p\,\mathrm{min}}$, the maximum plate current $i_{p\,\mathrm{max}}$, the driver conduction time interval τ_d and the time of one sweep cycle τ_H. The power dissipation is derived from the basic form

$$(41) \qquad P_{p1} = \frac{1}{\tau_H} \int_o^T i_p\, e_p\, dt$$

from which it may be shown that, in the idealized case,

$$P_{p1} = \frac{1}{\tau_H} \int_o^{\tau_d} \left\{ (e_{p\mathrm{min}} + \mathrm{e_{Rb}}) \int_o^t \left(\frac{di_p}{dt}\right) dt - R_0 \left[\int_o^t \left(\frac{di_p}{dt}\right) dt \right]^2 \right\} dt$$

$$(42) \qquad = \mathrm{i_p}\ [((e_{p\,\mathrm{min}} + \mathrm{e_{Rb}})/6) + e_{p\,\mathrm{min}}/3]\ \tau_d\ /\ \tau_H$$

The addition of the 15 per cent remote cut-off current approximation allowance of Eq. (40) at plate voltage $(e_{p\,\mathrm{min}} + \mathrm{e_{Rb}})$ adds a term

$$(43) \qquad P_{p2} = 0.15\ \mathrm{i_p}\ (e_{p\,\mathrm{min}} + \mathrm{e_{Rb}})\, \tau_d\ /\ 2\tau_H$$

to yield the completed equation

$$(44) \qquad P_p = P_{p1} + P_{p2} = \frac{\mathrm{i_p}\, \tau_d}{\tau_H}\ [0.575\, e_{p\,\mathrm{min}} + 0.242\ \mathrm{e_{Rb}}]$$

In a typical example P_p = 5.12 watts driver-tube *plate* power dissipation. The rated value for the 6BG6G driver tube is 20 watts.

It is generally true that efficiently operated driver tubes are not limited by plate power dissipation, but rather by the energy dissipation on the screen grid. Limitation of the *minimum* plate voltage to values greater than 60 volts with no more than 300 volts applied to the screen grid usually assures operation of the 6BG6G well below its maximum screen dissipation rating of 3.2 watts.

IX. THE BOOSTER-DAMPER CIRCUIT

By use of the Booster-Damper circuit shown in Fig. 24, the stored energy retrieved from the magnetic circuit during the first part of each trace or sweep period, while the damper tube is functioning, is fed back into the power supply circuit of the driver tube. If the *average* damper tube current is made equal to the average driver plate current plus any other load current, then the voltage of the damper load circuit may be added directly to the voltage $\overline{E}_B$ of the power system. If rate of change of yoke current differs greatly from that of the driver plate current, the equalizing of the average plate currents requires some alterations of the circuit. A Booster-Damper circuit is shown in Fig. 25. The required voltage from the actual power supply is

$$\overline{E}_{BB} = \overline{E}_B - E_{RD} \tag{45}$$

The total input power to the plate circuit of the driver tube is

$$P_{BB} = \overline{E}_{BB} I_p \tag{46}$$

With perhaps 30 watts system input, there may be about 5 watts dissipated in the 6BG6G driver tube, approximately 6 watts in the deflection transformer and yoke system, 3 watts in the kinescope second anode power supply system and load, 7 watts in the auxiliary damper load resistor, about 8 watts in the damper tubes, and the balance of about one watt in the damper linearity control resistors.

The best available laminated core transformers with laminated or wire-wound yoke cores produce a combined loss about 50 per cent greater than that of the powdered iron core system. The use of laminations of more than 0.005 inch thickness produces a very inefficient system. One previously used system, with 0.014-inch transformer lamination thickness, was responsible for a 20 watt power loss in the transformer and yoke while deflecting a 7-kilovolt beam only sufficiently for a 40-degree kinescope. The same system with iron-dust cores adequately deflects a 12-kilovolt beam for a 50-degree kinescope.

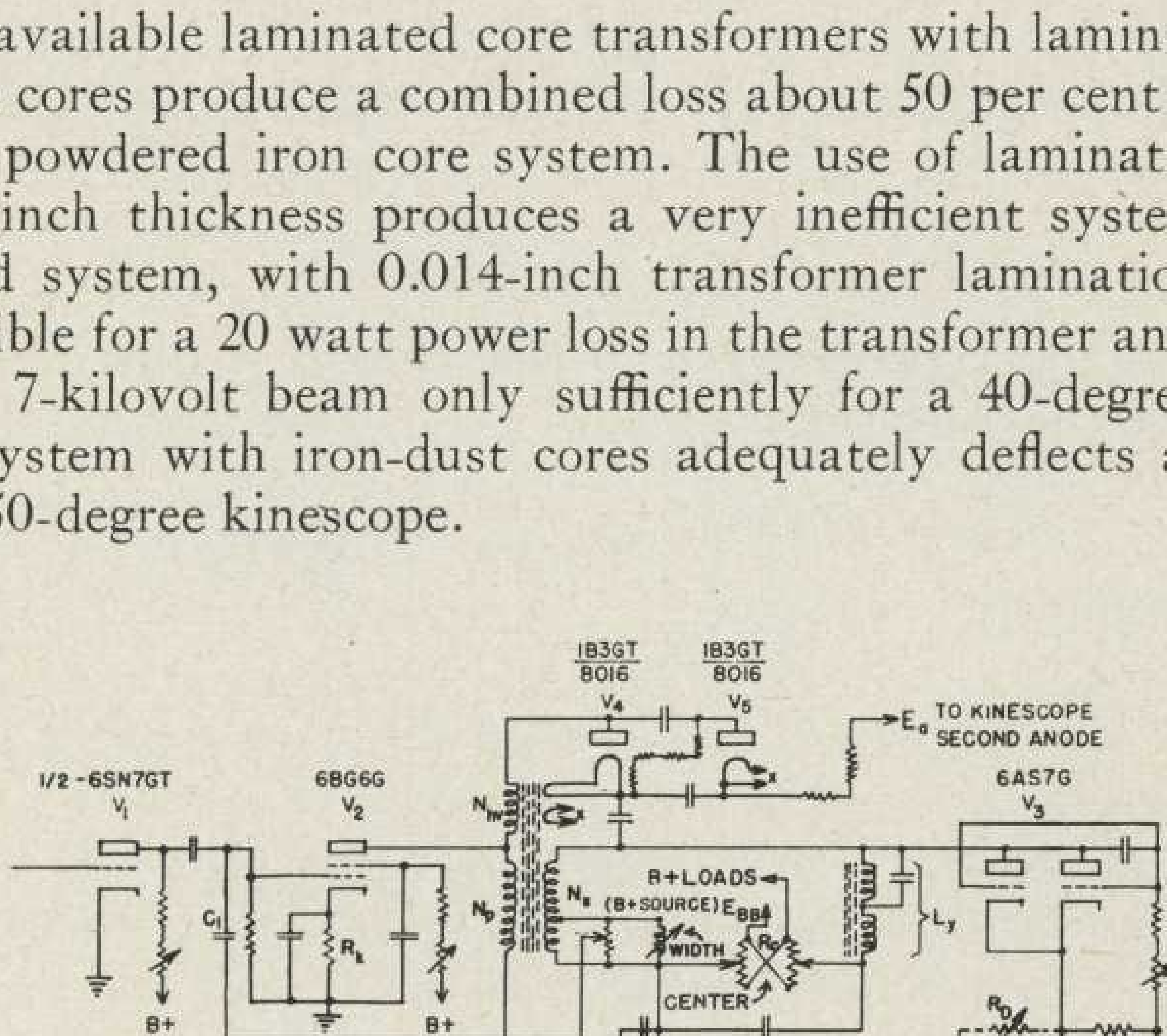


FIG. 24—The basic Booster-Damper circuit for recovering energy to be salvaged from low-loss horizontal deflection system. A voltage-doubler second-anode power-supply circuit is shown.

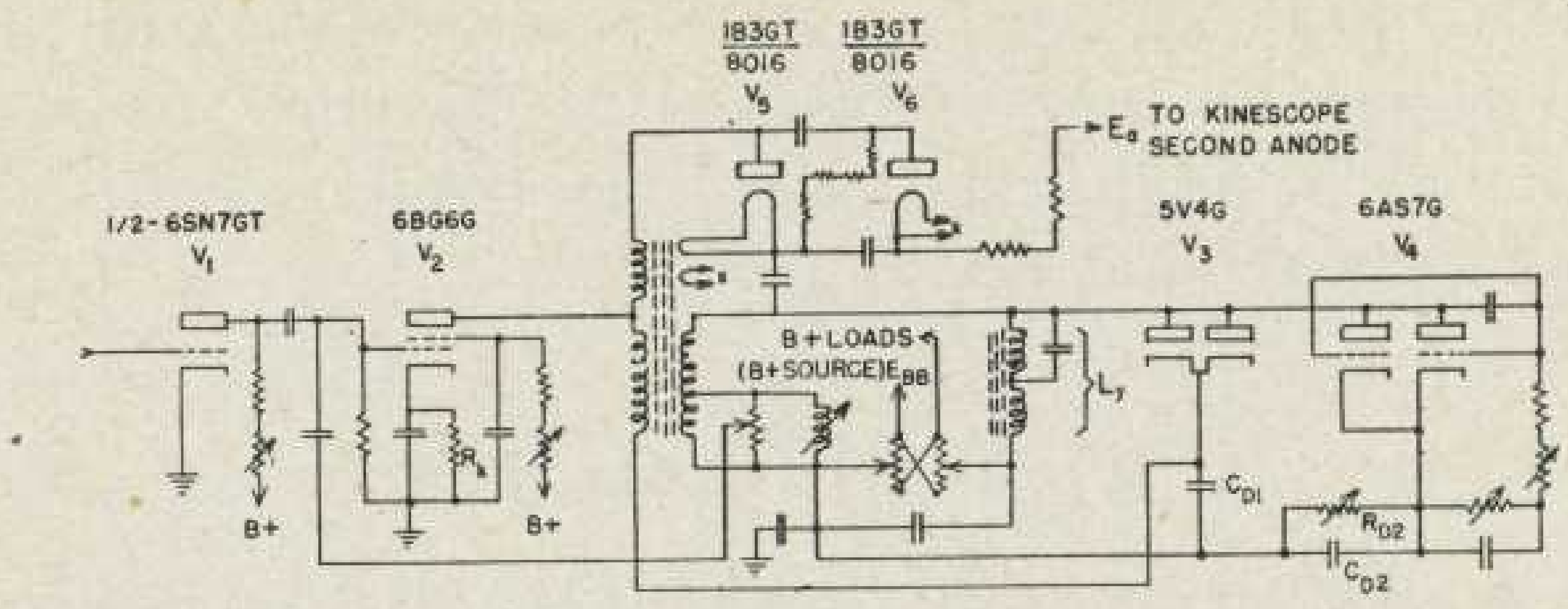


FIG. 25—The Booster-Damper circuit with a double damper arrangement for use with low-inductance horizontal deflection windings and a single 6BG6G or 807 driver tube.

With certain inductance, voltage and tube combinations, the d-c booster voltage approaches 200 volts. In this case no auxiliary damper load resistor is required, so its energy dissipation is eliminated.

The higher boost voltage may be attained without exceeding vacuum tube peak pulse ratings, and with the 8.2 millihenry yoke inductance, if *two* 6BG6G driver tubes are used in a highly efficient system with no auxiliary damper load resistor.

High energy recovery systems are possible only when low-loss transformer and yoke cores are used. At present the iron-dust core is the only low-loss, inexpensive core available for this application. Present estimates of powdered-iron cores indicate cost figures approximating one-fifth to one-tenth that of the nearest equivalent laminated-iron transformer core types.

Tube ratings at this time allow the development of up to 17 kilovolts with full deflection of a 50-degree kinescope by means of a single 6BG6G driver tube with an 8.2-millihenry horizontal deflection-yoke winding. With two 6BG6G beam tetrode driver tubes in parallel it is not difficult to produce at least 27-kilovolts second-anode potential when employing a pulse-voltage-tripler rectifier system. Under these circumstances more than adequate horizontal deflection of the electron beam is available for a standard 50-degree kinescope. It is necessary that a type 6AS7G duo-triode damper tube be used to produce the desired deflection linearity under these operating conditions.

X. CONCLUSIONS

Molded powdered-iron cores for television horizontal deflection transformers and yokes offer a simple solution to the problem of providing a high performance system at a *very* low cost. The great reduction in acoustic output from this core is fortunate. Higher permeabilities are desirable to increase the coupling coefficient of the deflection transformer. Higher molding pressures may provide high permeabilities if costs can be maintained at a low level.

A single driver tube, with a single damper tube and two small rectifier tubes, can provide full deflection and second-anode high-voltage for a 50-degree kineoscope, at accelerating voltages up to 17 kilovolts under present tube ratings. Two 6BG6G driver tubes can easily provide full deflection for a standard 50-degree kinescope operated at 27 kilovolts second anode potential which may be produced by a voltage-tripler rectifier arrangement from the deflection system retrace pulse.

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