

Jan. 15, 1935.

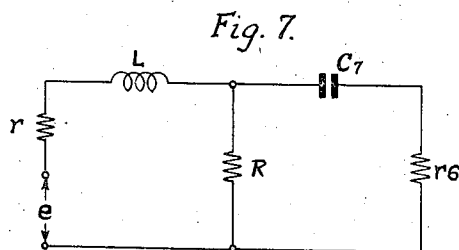
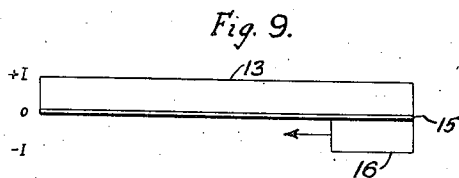
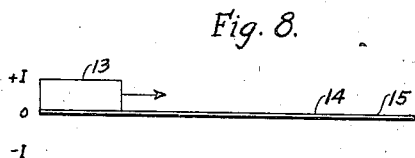
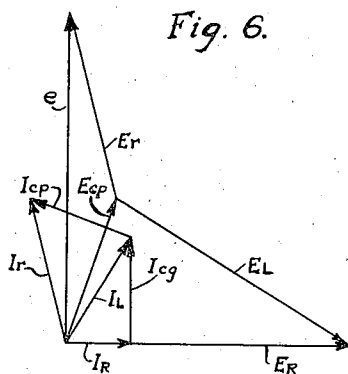
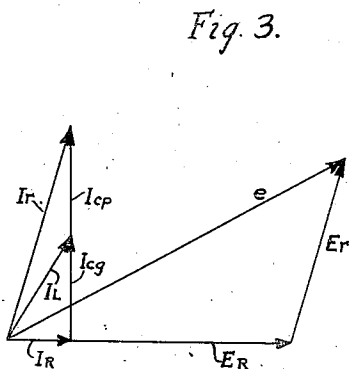
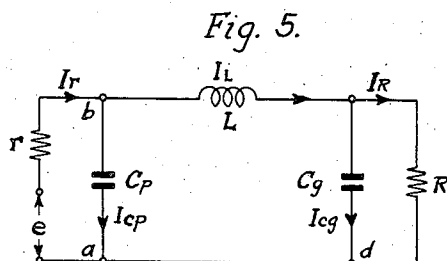
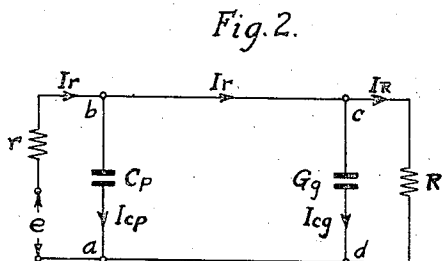
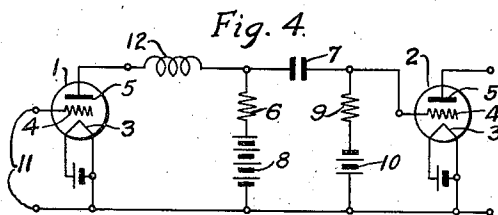
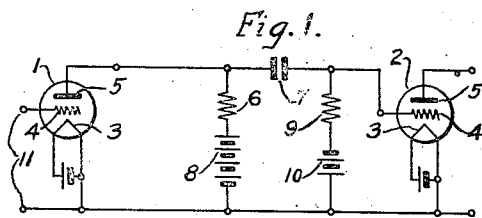
V. D. LANDON ET AL

1,987,687

AMPLIFYING SYSTEM

Filed March 20, 1930

2 Sheets-Sheet 1



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1,987,687

AMPLIFYING SYSTEM

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2 Sheets-Sheet 2

Fig. 10.

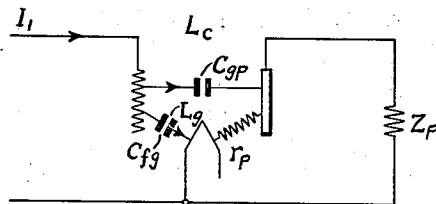


Fig. 11

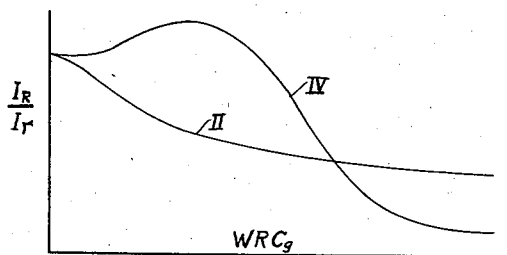


Fig. 12

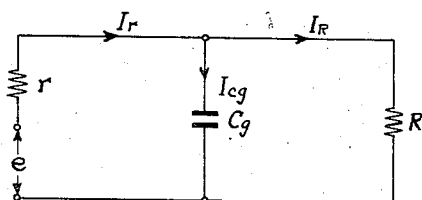


Fig. 13.

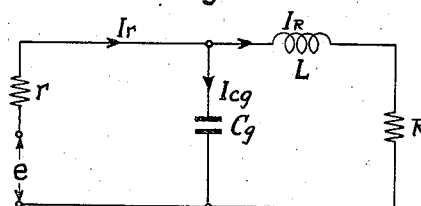
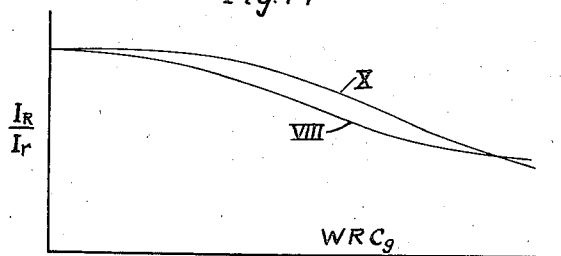


Fig. 14



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1,987,687

AMPLIFYING SYSTEM

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3 Claims. (Cl. 179—171)

Our invention relates to improvements in electrical amplifiers and it has particular relation to filtering systems.

One of the major problems in amplifier design is to provide uniform amplification of all signal frequencies. This problem has been solved in a fairly successful manner for amplifiers designed to operate in the audio-frequency range of the sound spectrum and particularly for frequencies between 100 and 5000 cycles per second. Very often, however, there is a demand for an amplifier that will operate effectively over a range extending well into the radio frequencies. For example, television signals, wherein the allowable distortion is extremely small, extend over a broad range of frequencies.

In the amplifiers constructed according to the teaching of the prior art, the interelectrode impedances of the amplifying tubes practically short-circuit the coupling impedances at high frequencies, thereby decreasing the amplitude of the signal transmitted from the plate circuit of one amplifier tube to the grid circuit of the succeeding tube and causing distortion.

It is, accordingly, an object of our invention to reduce distortion in electrical amplifiers.

Another object of our invention is to provide an amplifier that shall cut off sharply at the frequency beyond which no response is desirable.

Still another object of our invention is to increase the high-frequency response in electrical amplifiers.

An additional object of our invention is to provide an amplifier that shall have a uniform frequency-response characteristic.

More specifically stated, it is an object of our invention to eliminate reflection of electrical waves in the coupling circuits of electrical amplifiers.

According to our invention, we provide additional impedances for amplifier-coupling circuits of such value and nature that the characteristic impedance of the coupling circuit is made substantially equivalent to the coupling impedance.

The novel features that we consider characteristic of our invention are set forth with particularity in the appended claims. The invention itself, however, both as to its organization and its method of operation, together with additional objects and advantages thereof, will best be understood from the following description of a specific embodiment, when read in connection with the accompanying drawings, in which

Figure 1 is a schematic representation of a section of a resistance-coupled amplifier, not provided with our improvement.

Fig. 2 is a schematic representation of an electrical network equivalent to that of Fig. 1 at high frequencies.

Fig. 3 is a vector diagram showing the relation existing between the currents and electromotive forces indicated in Fig. 2.

Fig. 4 is a schematic representation of the section of the amplifier shown in Fig. 1, with a preferred embodiment of our invention included therein.

Fig. 5 is a schematic representation of the electrical network equivalent to that of Fig. 4 at high frequencies.

Fig. 6 is a vector diagram showing the relations existing between the currents and the electromotive forces indicated in Fig. 5.

Fig. 7 is a schematic representation of an electrical network equivalent to that indicated in Fig. 4 at low frequencies.

Figs. 8 and 9 are drawings inserted to aid in describing the effects of electrical discontinuities on the propagation of an electrical disturbance along a wire.

Fig. 10 is a schematic representation of a tube showing the agents that effect certain of its apparent properties.

Figs. 11 and 14 are graphs showing certain properties of the circuits discussed herein.

Figs. 12 and 13 are schematic drawings showing the apparent electrical form assumed by certain modifications of what is shown in Figs. 1 and 4.

Referring to the drawings in detail, the amplifier section depicted in Fig. 1 comprises two thermionic tubes 1 and 2, each containing a cathode 3, a control electrode or grid 4 and a plate electrode 5.

The two tubes are coupled electrically through a resistance 6. A blocking condenser 7 is customarily inserted in the grid circuit of the second tube 2 in order to block direct current from the plate battery 8, which would otherwise bias the grid of the second tube. A high resistance 9 is employed as a grid leak, and the usual grid-biasing battery 10 may be desirable.

The signal to be amplified is impressed upon the input terminals 11, and the amplifier response is impressed upon the grid-cathode circuit of the second tube.

The equivalent circuit of Fig. 2 is derived from the amplifier of Fig. 1. At high frequencies, the large blocking condenser 7 offers a negligible impedance to the signal current and may be considered a short circuit. The grid-leak resistor 9 is large, as compared with the coupling resistor 6 and also may be neglected.

The coupling circuit, at high frequencies, thus reduces the equivalent network of Fig. 2, in which the capacitor C_p is equivalent to the cathode-plate capacitance of the first tube 1, and the capacitor C_g is equivalent to the apparent cathode-grid capacitance of the second tube 2. r represents the plate resistance of the first tube in the

network, and e represents the effective signal voltage induced in the plate circuit of the first tube 1. The portion of the network of Figs. 2 and 5, designated by the characters $a b c d$, is ordinarily regarded as a coupling circuit.

The apparent cathode-grid capacity C_g is not precisely the actual capacity between grid and cathode, since the grid is influenced in its electrical action by the plate. The value of C_g is a function of the constants of the thermionic tube and of the circuit in which it is used. The functional relation between C_g and these constants may be obtained by analysis of the circuit represented in Fig. 10, in which

- 15 C_{gp} =grid-plate capacity
- C_{gc} =actual grid-cathode capacity
- r_p =apparent resistance between cathode and plate
- I =total current entering grid
- 20 I_c =current from grid to plate
- i_{cg} =current from filament to grid
- μ =positive value of the amplification coefficient of the tube
- Z_p =impedance of the external circuit into which the tube delivers its current

On setting up and solving Kirchhoff's equations for the electrical network comprised in the tubes, we find that the apparent grid-cathode capacity may be defined by the equation

$$(1) \quad C_g = \left(1 + \frac{\mu Z_p}{r_p + Z_p}\right) C_{gp} + C_{gc}$$

In certain tubes, particularly those of the three-electrode type, the capacities C_g and C_p are approximately equivalent. On the other hand, for tubes including a screening grid, and also, in rare cases, in three-electrode tubes, C_p may be neglected in comparison with C_g . Our invention may be applied to both types of tubes as will appear hereinafter. We shall first, however, apply it to the type wherein $C_p = C_g$.

As shown in Fig. 4, our invention comprises the disposition of an inductance 12 between the plate of one tube and the grid of the succeeding tube. In Fig. 5, which shows schematically the electrical equivalent of Fig. 4 at high frequencies, the inductance 12 is represented by the inductance L .

The function of the inductance 12 may be best demonstrated by analyzing Fig. 2, with the aid of Fig. 3 and Fig. 5, with the aid of Fig. 6, and by comparing the ratio of the output to the input currents which each analysis yields. From the analysis, the specific value of the inductance, as defined by the constants of the network, will also appear.

In addition to the characters in Figs. 2, 3, 5 and 6 defined hereinabove, other characters are present which represent the physical properties of the circuit as follows:

- 60 I_{cp} =current through C_p
- I_{cg} =current through C_g
- I_R =current through R
- I_r =current through r
- I_L =current through L
- 65 E_R =potential drop across R
- E_r =potential drop across r
- E_L =potential drop across L
- E_{cp} =potential drop across C

70 $\frac{W}{2\pi}$ =frequency of the alternating electromotive force impressed.

In Fig. 3, the vectors representing the currents and electromotive forces, for a circuit of the type illustrated in Fig. 2, are shown in their proper

phase and magnitude relationship. The vectors are labelled in accordance with the table hereinabove set forth.

In Fig. 6, the corresponding vector diagram is shown for Fig. 5.

By applying Kirchhoff's law to Fig. 2, as shown in Fig. 3, and adding the condition $C_g = C_p$, we obtain the expression

$$(2) \quad \frac{I_R}{I_r} = \frac{1}{\sqrt{1 + 4W^2 R^2 C_g^2}}$$

and the corresponding equation

$$(3) \quad \frac{I_R}{I_r} = \frac{1}{\sqrt{(1 - W^2 L C_g)^2 + W^2 C_g^2 R^2 (2 - W^2 L C_g)^2}}$$

is obtained by applying Kirchhoff's law as shown in Fig. 5, to Fig. 4, and adding the condition $C_g = C_p$.

Furthermore, by applying the condition

$$L = 2R^2 C_g$$

we obtain from (3) the relation

$$(4) \quad \frac{I_R}{I_r} = \frac{1}{\sqrt{1 - 4W^4 R^4 C_g^4 + 4W^2 R^2 C_g^2}}$$

Fig. 11 shows the curves obtained by plotting

$$\frac{I_R}{I_r}$$

against WRC_g . Curve II is obtained from Equation 2, and curve IV is obtained from Equation 4. Curves II and IV give the frequency characteristic of the output current of the two respective circuits for an input current having a flat frequency characteristic.

It is obvious from the two curves, that by adding the inductance in the circuit, we have raised the response at the high-frequency end of the spectrum. In addition, in an amplifier constructed according to our invention, the cut-off at the higher frequency is considerably sharper than in an amplifier constructed according to the teachings of the prior art.

It should be further noted that we may obtain a considerable variety of frequency-response characteristics for the output currents by merely putting into the circuit the requisite value of L . Of particular interest in this connection is the application of our invention to the case wherein the frequency-response characteristic of the input current is not flat.

We do not wish, therefore, to be restricted to the one value of L defined by the relation

$$L = 2R^2 C_g.$$

The particular application of our invention demonstrated hereinbefore may also be considered with reference to the well-established principle that, in order to prevent reflections of electrical waves in an electrical network, the network should terminate in a load impedance equal to the characteristic impedance of the network. This principle is reduced to our particular network, as will appear from the following consideration.

The characteristic impedance of a distributed network is represented by the well known relation

$$Z_0 = \left(\frac{\text{network impedance}}{\text{network admittance}} \right)^{1/2} = \sqrt{\frac{Z}{Y}}$$

A simple case illustrating the meaning of this statement is presented in Fig. 8, which shows a single electrical signal wave of current 13 traveling along a conductor 14, in the direction of the arrow, toward an open circuit (infinite imped-

ance) terminal 15. The wave, upon reaching this terminal, is reflected and reversed in sign, because of the high terminal impedance. The reflected wave 16, indicated in Fig. 9, since it is of opposite sign, wipes out the transmitted wave 13, the result being a zero transmission of current.

As the terminal impedance decreases, the value of the reflected wave decreases until the terminal impedance is equal, in value, to the characteristic impedance of the conductor. Such a terminal impedance, with reference to the current wave, cannot be distinguished from the conductor, and hence no reflection occurs.

Similarly, it may be shown that an electrical wave of voltage, upon meeting an open-circuit terminal, is reflected without reversal of sign, and, therefore, the reflected wave adds to the transmitted wave. As the terminal impedance decreases, the reflected voltage wave decreases until, when the characteristics of the line and terminal are equal, no reflection occurs. Further decrease of terminal impedance below the value of the conductor characteristic impedance causes the reflected wave to increase with reversal of sign with respect to the original reflections, until, for a short-circuit, (zero impedance) terminal, the current wave is fully reflected in the same sign, and the voltage wave is fully reflected in the opposite sign with respect to the transmitted waves. Thus, for a circuit in which the load or terminal impedance and the line characteristic are equalized and reflection is prevented, the transmitted power is theoretically equivalent to the input power.

For networks, such as the filter section of Figs. 2, 5 in which the impedance consists, practically, of inductance, and the admittance consists of capacitance only, the characteristic impedance becomes:

$$Z_0 = \left(\frac{\text{network inductance}}{\text{network capacitance}} \right)^{1/2} = \sqrt{\frac{L}{C}}$$

This relationship is approximate since in the present case the impedances L and C are lumped rather than distributed. However, the approximation is sufficient for practical purposes. A more accurate relationship for lumped impedances is found on page 125 of Johnson's "Transmission Circuits for Telephonic Communication" and has the following form when reduced to the symbols used here.

$$Z_0 = \sqrt{\frac{L}{2Cg(1 + \frac{1}{2}LCg)}}$$

Since the resistance-coupled amplifier of Fig. 1 has practically no inductance, the characteristic impedance of the filter section represented in Fig. 2 is:

$$Z_0 = \sqrt{\frac{L}{C}} = 0$$

which indicates that a balanced circuit (Z_0 =load impedance= R) requires a zero load or coupling resistance R . However, the value of the coupling resistance is determined by the electron-tube characteristics, and is a definite, finite value. It is essential, therefore, to provide an inductance having the value

$$L = R^2 C = 2R^2 C_g \text{ (assuming } C_p = C_g \text{)}$$

in order to have a characteristic impedance equal to the load impedance, thus insuring distortionless signal transmission.

If the more accurate relationship is used, the desired condition is

$$R = \sqrt{\frac{L}{2Cg(1 + \frac{1}{2}LCg)}}$$

but

$$W^2 = \frac{8}{LCg}$$

and hence

$$R = \sqrt{\frac{4}{W^2 C^2 g \left(1 + \frac{4}{W^2} \right)}}$$

which for large W becomes

$$R = \sqrt{\frac{4}{W^2 C^2 g}} = \sqrt{\frac{L}{2Cg}}$$

whence

$$L = 2R^2 Cg$$

The improved operation resulting from the particular application of our invention may be shown, in still another light, by referring to frequency cut-off values of filters. Low-pass filters, such that shown in Fig. 5, are designed to pass, with negligible attenuation, the currents of all frequencies below the cut-off frequency. When the terminal impedance R is equal to the characteristic impedance of the filter, all frequencies are passed, with low, uniform attenuation, below a cut-off frequency determined by the well-known formula:

$$(5) \quad \frac{W}{2\pi} = \frac{1}{\pi\sqrt{LC}} = \frac{1}{\pi\sqrt{LCg/2}} \text{ (assuming } C_p = Cg \text{)}$$

or since $L = 2R^2 Cg$ in this case,

$$(6) \quad \frac{W}{2\pi} = \frac{1}{\pi RCg}$$

It should be noted that the capacities C_g and C_p , being in series with respect to the resonant circuit, result in a net capacity lower in value than either of the component parts. The lower the net capacity, the higher is the high-frequency cut-off value as indicated by the Formula 5.

In the amplifying circuit, as heretofore used, illustrated in Figs. 1 and 2, the above Formula 5 would give a theoretical cut-off frequency of infinity, since $L=0$. The attenuation of current increases, however, with the frequency, since the capacitors are compensated by no inductance. Assuming an attenuation of approximately 30% that is, referring to Fig. 2.

$$1.00 - 0.30 = \frac{I_R}{I_r} = \frac{I_R}{\sqrt{I_C^2 + I_R^2}} \text{ (where } I_C = I_{Cp} + I_{Cg} \text{)}$$

$$= \frac{E_R/R}{\sqrt{W^2 E_R^2 C^2 + E_R^2/R^2}}$$

or solving for $\frac{W}{2\pi}$ the effective cut-off frequency,

$$(7) \quad \frac{W}{2\pi} = \frac{1}{2\pi RC} = \frac{1}{4\pi RCg} \text{ (assuming } C_p = Cg \text{)}$$

A comparison of Formulae (6) and (7) indicates that, by the insertion of a suitable inductance into the amplifier circuit, the effective cut-off frequency or maximum working frequency is practically doubled.

The device thus opens a new field to amplifiers of electrical energy, in which there has been a failure to obtain suitable response.

At low frequencies, the equivalent coupling network for the amplifier of Fig. 2 reduces to the form shown in Fig. 7. The electron-tube inter-electrode capacitances are very small and may be regarded as open circuits. The circuit of Fig. 7 resembles a high-pass filter, feeding the load

resistance r_s . The inductance offers a negligible impedance at the lower frequencies, and, therefore, does not materially alter the low-frequency transmission of signals by the coupling circuit.

- 5 The criterion of cut-off frequency is the relation of grid-leak resistance R to the impedance of the blocking condenser 7. Below the frequency at which these two quantities become equal, amplifier performances may be considered unsatisfactory.

10 We shall now demonstrate how our invention may be applied to a system wherein C_p may be regarded as small in comparison to C_g . In Figs. 12 and 13, the electrical form that the system, with and without our improvement, assumes at high frequencies is shown schematically.

Attention is called to the position of the inductance. As shown, it is now disposed between the coupling resistance and the grid.

- 20 For the circuit not including our improvement we obtain, by applying Kirchhoff's laws,

$$(8) \quad \frac{I_R}{I_r} = \frac{1}{\sqrt{1 + W^2 R^2 C_g^2}}$$

- 25 For the circuit including our improvement, Kirchhoff's laws yield

$$(9) \quad \frac{I_R}{I_r} = \frac{1}{\sqrt{(1 - W^2 L C_g)^2 + W^2 R^2 C_g^2}}$$

- 30 and if we put

$$L = \frac{R^2 C_g}{2}$$

(9) becomes

$$(10) \quad \frac{I_R}{I_r} = \frac{1}{\sqrt{1 + \frac{1}{4} W^4 R^4 C_g^4}}$$

- In Fig. 14, curve VIII shows graphically the frequency-response relation given by Equation 8, and curve X is the corresponding curve for Equation (10). The advantage of the inductance is obvious from the curves.

Analysis of the ratio

$$45 \quad \frac{I_R}{I_r}$$

shows that the inductor L may with considerable advantage be disposed between resistance r and the grid.

- 50 Our invention is not to be confused with the use of filters in combination with electron tubes for excluding certain undesirable signal or interference frequencies. We have appreciated the injurious influence of electron-tube interelectrode capacitances, and have coordinated these capacitances with the external amplifier circuits and appropriate compensating elements, to produce a resultant amplifier, not for the purpose of excluding certain frequencies, nor for narrowing the frequency range, but for passing all possible frequencies up to a certain point with substantially no discrimination.

Our invention, therefore, is directed to a balancing of the coupling-circuit characteristic im-

pedance to the coupling impedance and effects the following improvements in amplifier performance.

- (1) The effective maximum frequency of an electrical amplifier is increased.
- (2) The response of an electrical amplifier at high frequencies is increased.
- (3) Distortion is decreased.
- (4) Signals above the frequency range desired are diminished.

Each of these results, in itself, attests the importance of our invention, but the combination of these three benefits is unquestionably a marked advance in coupling-circuit design and permits vastly improved performance in the exacting field of broad frequency-range amplification.

Although we have shown and described certain specific embodiments of our invention, we are fully aware that many modifications thereof are possible. Our invention, therefore, is not to be restricted except insofar as is necessitated by the prior art and by the spirit of the appended claims.

We claim as our invention:

1. An electrical network comprising a plurality of electric discharge devices, each of said devices containing a plurality of capacitively related electrodes, means for electrically coupling certain of the capacitively-related electrodes of one of said devices to capacitively-related electrodes of another of said devices, impedance means in circuit with said coupling means, and means for decreasing the discrepancy, between the value of the characteristic impedance of the circuit comprising said coupling means and said coupled capacitively associated electrodes and the value of the impedance in circuit with said coupling means.

2. An electrical network comprising an electric discharge device containing output elements, said output elements being capacitively related, a second electric discharge device containing input elements, said input elements being capacitively related, coupling means for electrically connecting said output elements to said input elements, a coupling impedance in circuit with said coupling means, and means for making the characteristic impedance of the circuit, comprising said coupling means and said capacitively related elements, substantially equal to the impedance in circuit with said coupling means.

3. A resistance-coupled amplifier comprising thermionic tubes and an inductance disposed between the plate of one tube and grid of the succeeding tube, the value of said inductance being substantially defined by the relation $L = R^2 C$ where L is the value of said inductance, R is the value of said coupling resistance, C is the value of the sum of the apparent capacities of the condensers formed by the filament and the plate of said first tube, and by the filament and the grid of said second tube.

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