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TRIODES VERSUS PENTODES IN HIGH-FIDELITY OUTPUT STAGES

By F. Langford-Smith.

Pentodes and beam power pentodes, single-ended and push-pull, are almost universally used in all types of amplifiers other than high-fidelity. Why are push-pull triodes so often used in place of pentodes in high-fidelity amplifiers?

The answer to this question is not a simple one. Push-pull triodes have lower total harmonic distortion than push-pull pentodes when operated into a constant resistive load — from 1% to 2% for triodes under low distortion conditions, and about 3.5 to 5% for pentodes, but about 2% for beam power tetrodes operated under the best possible conditions. So that on these figures there is not very much to choose between triodes and beam power tetrodes, allowing for use of, say, 20 db of feedback.

In practice, a power amplifier never works into either a constant load or a purely resistive load, since a loudspeaker has a highly variable impedance, and a highly variable phase angle. The magnitude of the impedance of a typical direct-radiator loudspeaker, either on a flat baffle or in a totally-enclosed cabinet, may vary from 80% to 1000% of its rating impedance, and the phase angle from + 60° to — 50°.

It so happens that push-pull triodes are very little affected by changes in load impedance — the increase up to 10 times the rating impedance actually causes a decrease in distortion, while the decrease to 0.8 times the rating impedance merely results in a very slight increase in distortion — and there is no danger of running into grid current, provided that the input voltage is kept constant. On the other hand pentodes or beam power tetrodes are very much affected by changes in load impedance, and the distortion increases rapidly with low or high values of load impedance, as illustrated in Fig. 1.

It will be seen that the distortion increases from 2% to over 16% when the load resistance is increased from 7,000 to 14,000 ohms. Negative voltage feedback acts so as to reduce the rate of rise of distortion with load resistance, because the increased load resistance gives higher stage gain and hence greater reduction in distortion. In this particular case the 2:1 increase in load resistance gives roughly 2:1 increase in gain, and 20 db of feedback would reduce the 2% distortion at the nominal load resistance to about one tenth of this value (i.e., 0.2%), whereas the 16% distortion at twice nominal load resistance would be reduced to about one twentieth of this value (i.e. 0.8%).

It is true that if pentodes or beam power tetrodes are operated at reduced signal input voltages both the initial distortion and the rate of rise of distortion at higher load resistances will be decreased. By this means pentodes may be shown to appear in a less unfavourable light when compared with triodes operating on a varying load resistance, but the maximum power output for low distortion would also be reduced.

Danger of running into Grid Current

If feedback is not used, neither triodes nor pentodes are in danger of running into grid current, with any value of purely resistive load, provided that the signal input voltage is maintained constant at the recommended value for normal load resistance. When negative voltage feedback is used with push-pull triodes or pentodes, with constant signal input voltage, no harm occurs with any value of resistive load from the nominal value upwards, but grid current is encountered when the load resistance is decreased below normal. Two possible methods of coping with this problem are:

1. To reduce the input signal voltage sufficiently, pentodes require considerably greater reduction than triodes.

2. To base the design on the minimum loadspeaker impedance, as shown by measurements, in the range from 100 to 400 c/s, instead of its rating impedance value.

Effects of Reactive Loads

The discussion up to this point has been limited to the rather academic case of purely resistive loads. This applies to the peak of the impedance characteristic at the bass resonance frequency, and at one or two other individual frequencies, but does not apply generally. The effect of a partially reactive load impedance is to produce an elliptical loadline which can readily be seen on a C.R.O. when the output from a beat frequency oscillator is amplified and used to feed a loudspeaker. One result of an elliptical loadline is the tendency to run into grid current, even though the input voltage is maintained constant at a value which does not result in grid current, with a purely resistive load. The grid current may be avoided by a reduction in signal input voltage, resulting in a reduction of power output. This effect is much more marked with pentodes than with triodes. It seems that push-pull triodes are as little affected by reactive loads as any conceivable "ideal" output system.

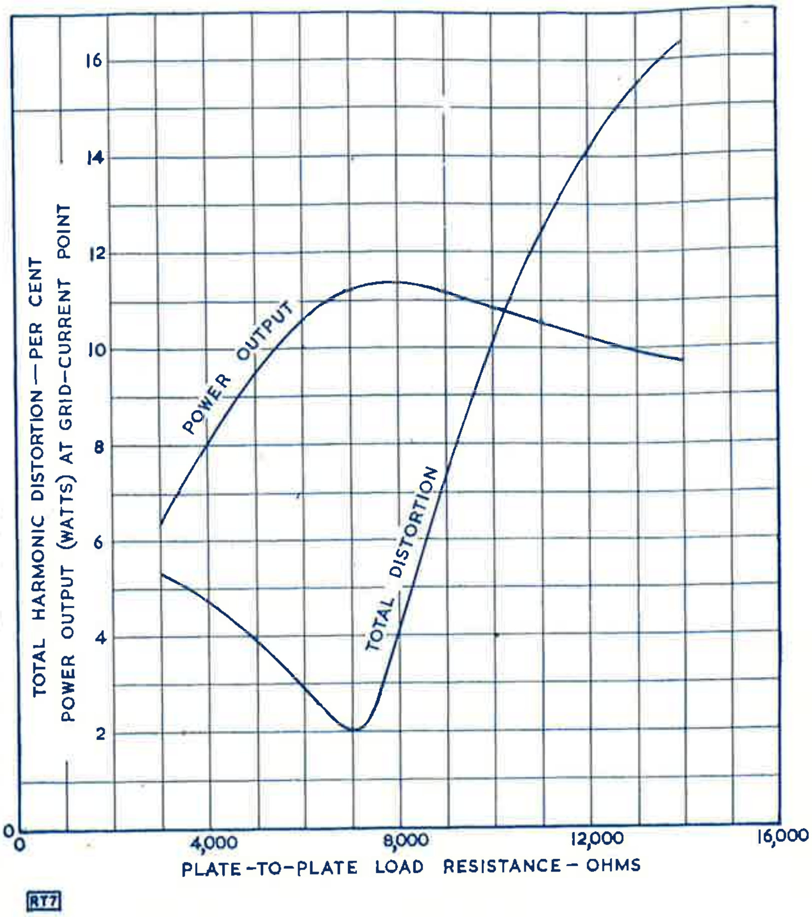

Fig. 1. Total harmonic distortion and power output of type 6L6 beam power amplifier in Class A push-pull, plotted against plate-to-plate load resistance. Supply voltage 266.5 volts, cathode bias resistor 257 ohms (by-passed); zero signal cathode current 81 mA; internal resistance of B supply 1000 ohms

Conclusions

For these reasons, push-pull triodes with negative voltage feedback are almost universally used with the highest quality class of high-fidelity amplifiers. This usage has been adopted in spite of the known fact that, by the use of pentodes with somewhat increased feedback, an equally small percentage of total harmonic distortion may be achieved under the normal test conditions with a purely resistive load of optimum value. The increased feedback required by the pentodes usually involves complications to maintain the stability margin, thus adding to the cost.

However, the whole subject has not yet been thoroughly investigated, and it is possible that methods may be discovered for minimizing the deficiencies of pentodes. One possible method is the so-called "Ultra-Linear" Circuit in which the screen is taken from a tapping on the primary of the output transformer, giving any desired compromise between triode operation (with the tapping at the plate end) and pentode operation (with the tapping at the cold end). It is hoped, in the not distant future, to carry out extensive tests with this circuit.

An alternative approach is to use pentodes with a considerable margin of power output—possibly a nominal power output of twice that for the equivalent triode amplifier feeding the same loudspeaker.

A still further approach is to reduce the rise of impedance in the loudspeaker. This may be done, to a limited extent, in the design of the loudspeaker. Alternatively the rise of impedance at the bass end may be reduced by the careful use of an acoustical damping cloth over the whole rear area of an open-back loudspeaker enclosure (Ref. 1). It is hoped to include an article in Radiotronics on this subject. But no really satisfactory method has yet been developed to reduce the rise in impedance of an existing loudspeaker at the high frequency end. Shunting by a capacitance results in resonance, and capactive reactance at frequencies higher than the resonant frequency with loss of high frequency response. Shunting by a network consisting of a resistance and capacitance in a series has a somewhat similar effect, although less drastic.

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Revision of Article in Feb '55. — P.25.

LOUDSPEAKER DIVIDER NETWORKS

by J. Stewart and F. Langford-Smith.

When two or more loudspeakers are used in a two or three-way system, the optimum cross-over frequencies and attenuation characteristics of the divider network are functions of the loudspeakers employed.

Dual and triple loudspeaker systems are covered in the Radiotron Designer's Handbook, pages 860-861, while frequency divider networks are covered on pages 184-185 and 887-889. The present article is a more detailed presentation of this information, with the inclusion of some more recent additions to the literature on the subject.

Attenuation 6db/octave (quarter section)

This is the simplest and cheapest form, and is quite satisfactory for two-way systems, provided that the high frequency speaker is robust and capable of handling the full power output of the system down to a frequency one-half octave below the cross-over frequency, and appreciable power down to three octaves below the cross-over frequency. A good quality 5 inch or 6 inch speaker, with good high frequency performance, would be satisfactory for domestic use. This rate of attenuation is generally unsatisfactory with horn type tweeters, since a horn is not capable of handling even small amounts of power below its cut off frequency. The cross-over frequency would normally be between 800 and 1200 c/s, and the low frequency speaker should have reasonably good response up to at least 2000 c/s. Cross-over frequencies lower than 800 c/s may be used if the high frequency speaker is a direct radiator type capable of handling the maximum power down to a frequency one-half octave below the cross-over frequency (i.e., frequency ratio 1.4:1). The attenuation characteristics are the same as those for a single stage resistance-coupled amplifier e.g., see R.D.H. page 495, Fig. 12.9 (A).

Attenuation 12db/octave (half section)

This rate of attenuation does not make such drastic demands on the loudspeakers, permits wider choice of high frequency units, and makes possible the use of three-way systems. With the more rapid rate of attenuation than the simpler 6db/octave network, there is less danger of overloading the high frequency speaker in two-way systems, and even horn tweeters may be used provided that the cross-over frequency is two octaves above the horn cut-off frequency. Thus the choice of cross-over frequency is largely controlled by the characteristics of the high-frequency speaker, and may vary from 400-5000 c/s. With three-way systems it is possible to secure better results, since each speaker is not called upon to handle such a wide frequency range. The woofer may handle an upper limit between 300 and 600 c/s, the middle unit an upper limit between 2000 and 5000 cycles, and the tweeter will then look after the higher frequencies.

Attenuation 18db/octave

The use of such a high rate of attenuation is generally avoided owing to the higher cost and the greater amount of transient distortion. Its use is normally limited to two-way horn loudspeakers or horn tweeters, where it minimises the peaks and valleys in the overlap regions caused by phase differences, and also rapidly reduces the power input to the horn tweeter below the cross-over frequency.

Network types

Loudspeaker electrical dividing networks are of two principal types, the constant resistance filter type and the conventional filter type. Both of these are designed on the assumption that the load is purely resistive and constant in value. Since loudspeakers are neither purely resistive nor constant in impedance, care must be taken that the impedances of the two loudspeakers at the cross-over frequency are equal; they would then be reasonably well matched throughout the overlap region. It is highly desirable to make use of the loudspeaker impedance characteristic, which is sometimes available from the manufacturer, and in other cases can be measured (e.g., Refs. 16 and 17). The impedance should preferably be measured with the loudspeaker in the enclosure in which it will be used, although the impedance at frequencies above 400 c/s is not normally affected by the enclosure. The foregoing treatment of matching is based on the impedance of the loads at the cross-over frequency and gives results approximating to the theoretical values for the filter network. However, looked at from the amplifier point of view, better matching is sometimes achieved by the use of loudspeakers of different voice coil impedances—for example, 15 ohms for the low frequency speaker and 10 ohms for the high frequency speaker—owing to the rise in impedance which usually occurs above 500 c/s.; a lower nominal voice coil impedance would thus provide a better match at the higher frequencies with a pentode output stage. This would have an effect on the characteristics of the filter network in the cross-over region, and consideration should be given to both points. The constant resistance filter type of network (Refs. 10, 13, 24, 26, 27) has certain advantages over the conventional

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Figures 1, 2 and 3 are redrawn with permission of the I.R.E. from a paper by A. Meyer which appeared in the Trans. I.R.E. P.G.A. AU-1.5 Sept.-Oct., '53, p. 5.

filter type and the writers prefer it. The advantages of the constant resistance type are as follows:—

(a) The presentation of a constant resistive load on the amplifier throughout the frequency range, provided, of course, that the network is correctly terminated into constant resistive loads.

(b) For quarter and half section networks the components used in the network are of the same value, i.e., the inductors have the same value and the capacitors have the same value.

(c) The responses of the two filters are complementary so that the sum of the energies delivered to the loads is constant at all frequencies.

(d) The phase response of each filter is symmetrical about the cut-off frequency.

(e) The difference in phase between the signal delivered to the outputs is constant.

The conventional filter type has the advantage of greater attenuation outside the pass-band (Ref. 27). It has the slight disadvantages of giving a 1 db dip at the cross-over frequency when fed from a zero impedance source (Ref. 10), also of giving slightly greater insertion loss in the pass-band (Ref. 27).

A further class of filter is the mechanico-acoustical.

Some designs of tweeters make use of high-pass mechanico-acoustical filters to give the desired attenuation characteristics with the use of a series capacitor as the only electrical filter. Similarly, a low-pass mechanico-acoustical filter may be used for the low frequency unit without any electrical filter.

Practical aspects

(a) *The slope of the attenuation characteristic.* This choice is controlled by two factors, the speakers and the cost. The sharpness of the attenuation slope ensures that each unit handles only those frequencies for which it was intended, and that frequencies beyond its range are attenuated as much as possible. In the case of high pass sections this is important because frequencies below the acoustical cut-off of a loudspeaker suffer distortion if permitted to pass, also damage can be caused to the voice coil assembly. From the cost point of view, the Quarter section offers simplicity and a reduction in the capacitance and inductance required. Since the capacitors must be of high quality, preferably oil filled, the reduction of approximately 30% in capacitance with this network offers a saving to the home constructor.

(b) *Phasing.* It is important that the sound emitted from the two speakers be in phase. With a constant resistance type filter the phase difference is 90° for quarter section, 180° for half section and 270° for full section. No complications occur with the 180° difference for the half section, since this can easily be checked by reversing the leads to one of the units; if one of the speakers is out of phase there will be considerable loss of middle. For the other cases it is usually desirable to place the diaphragms one-quarter wavelength apart at the cross-over frequency, which will give in-phase output by correct connection. Phasing is not important above 3000 c/s. as the wavelength of the sound is so short; it may be found, owing to the binaural hearing effect, that phasing can be neglected down to even lower frequencies.

(c) *Sound level compensation.* The loudspeakers used in two and three-way systems usually have different sensitivities and power-handling capabilities, and some method of adjusting the overall level, either to one's own taste or by measurement, is desirable. This can be done by inserting a simple voltage divider between the dividing network and the speaker. Horn and direct radiator types of tweeters frequently have high sensitivity and need this type of attenuator. If desired, the attenuator may be made variable to act as a tone control—if this is done it is preferable to use a constant impedance method.

(d) *Insertion loss.* Owing to the fact that the coils used in dividing networks have resistance, insertion loss occurs. It is important therefore that the coils be wound with heavy gauge wire. For example, in a 100 watt system, 1 db loss corresponds to a power loss of about 20 watts. A loss of 0.5 db or 11 watts in a 100-watt system is as small as can normally be attained.

Use of dividing networks other than at loudspeaker terminals

Dividing networks may be used ahead of amplifiers to split the signal into two or more frequency bands; each band of frequencies being amplified in a separate channel and fed to the appropriate loudspeaker. This method offers advantages in that the networks can work correctly into purely resistive loads, which is the condition usually assumed but not attained in practice when the network feeds the loudspeaker directly. It is well known that loudspeakers cannot be considered as pure resistances, due to motional impedance and voice coil inductance. When the network is ahead of the amplifier, a volume control may be incorporated into one or both channels to adjust the overall tonal balance. Also with this system the amplifiers and components are only called upon to transmit a limited range of frequencies, this applies especially to the output transformers which may be of cheaper construction. However, the fact of having two or more amplifier channels, with their associated power supply demands, adds considerably to the cost over more conventional types. One feature of the multichannel method is that a certain amount of tone control is possible by controlling the volume of the bass and treble amplifiers. For this to be satisfactory the cross-over frequencies should be: bass not higher than 200 c/s, treble not lower than 3000 c/s. Another method claimed to have certain advantages is to place the dividing network between the output stage and the output transformers, using separate transformers for high and low channels. It is claimed that output transformer construction is simplified and less distortion is evident with this method.

Capacitors for dividing networks

The tolerance in capacitance may be ± 5% for ordinary use, but the manufacturers will supply closer tolerances to special order.

The use of electrolytic capacitors is undesirable for the following reasons:—

1. The capacitance varies with frequency.
2. The capacitance varies with applied voltage.
3. The tolerances in capacitance are very large.
4. They are normally manufactured for use in circuits with direct polarizing potential greater than the a.c. component.

Inductors for dividing networks

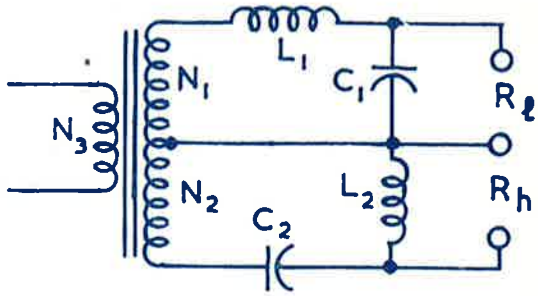
The requirements in respect of inductors used in networks between the output transformer and loudspeakers are as follows:—Low resistance, constant inductance and freedom from distortion (these usually exclude iron-core inductors), high "Q" and suitable orientation to prevent coupling. If air cored coils are used, it is important to space them as far apart as practicable, at right-angles to one another and as far as possible from metallic objects. In the case of inductors used ahead of amplifiers, the insertion loss is not important. Graphs are presented for the design of inductors giving maximum "Q" at 1000 cycles for use between the output transformer and speakers. The graphs give the physical sizes, number of turns, and d.c. resistance directly. Although these graphs are based on a frequency of 1000 c/s., no appreciable error will result from their use for frequencies up to 4000 c/s. At higher frequencies the self-capacitance of the winding will need to be taken into consideration, and it is desirable that the assembled filter have its characteristcs measured.

The curves determining the number of turns (Fig. 2) are based on perfect layer winding; as this cannot be obtained in practice the indicated number of turns may be increased by a factor of, say, 10%. The inductance should be checked on a bridge and any surplus turns removed.

Unequal voice coil impedances

Sometimes it is desirable to utilize, because of their performance, speakers which have different voice coil impedances. The usual method when this is desired is to design the output transformer with taps to feed each speaker, modifying the values of inductance and capacitance of each section of the network to suit the respective impedances. A circuit of this kind is shown below. The introduction of the transformer into the circuit gives complete control over the values of the circuit elements by virtue of the turns ratio of the transformer. The following relationships may be obtained:—

$$\left[\frac{N_3}{N_1}\right]^2 L_1 = \left[\frac{N_3}{N_2}\right]^2 L_2 \quad \ldots\ldots\ldots (1)$$



$$\left[\frac{N_1}{N_3}\right]^2 C_1 = \left[\frac{N_2}{N_3}\right]^2 C_2 \quad \ldots\ldots\ldots (2)$$

$$\left[\frac{N_3}{N_1}\right]^2 R_l = \left[\frac{N_3}{N_2}\right]^2 R_h \quad \ldots\ldots\ldots (3)$$

and the following design equations are applicable:—

$$L_1 = \frac{0.23 R_l}{f_c}$$

$$L_2 = \frac{0.23 R_h}{f_c}$$

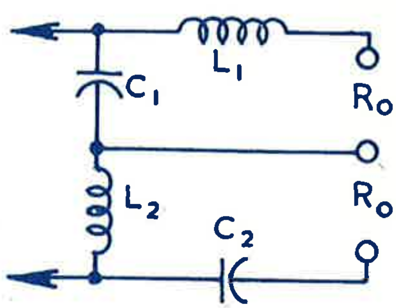
$$C_1 = \frac{0.11}{R_l f_c}$$

$$C_2 = \frac{0.11}{R_h f_c}$$

The above will be seen to be equivalent to the parallel-connected constant resistance type of network.

Design examples

Example (1). Owing to its popularity, it is proposed to discuss the design of a series-connected constant-resistance type of network for a cross-over frequency of 1000 c/s. and voice coil impedances of 10 ohms at 1000 c/s.



Since

$$L_1 = \frac{0.11\ R_o}{f_c}$$

and

$$C_1 = \frac{0.25}{f_c R_o}$$

then

$$L_1 = \frac{.11 \times 10 \times 1000}{1000} \text{ mH} = 1.1 \text{ mH}$$

and

$$C_1 = \frac{.23 \times 10^6}{1000 \times 10} \mu F = 23 \mu F.$$

To obtain the dimensions of the inductors, use can be made of the graphs. In the graphs a factor δ has been introduced as a constant having the following relationship to the coil dimensions:

$$\delta = 3A = 9B = 10C = 64.23 \left(\frac{L}{\phi^2}\right)^{1/5}$$

Where A, B and C are coil dimensions as given in the sketch, L is the coil inductance and ϕ is the number of turns possible per square inch. To use the graphs it is necessary to select a "Q" factor that is considered satisfactory, and then when the design has been completed check to see whether this figure has been attained. From Fig. 1 we get the wire size, and if we assume a "Q" of 25, 14.7 A.W.G. is given as the correct wire gauge, and since 15 A.W.G. is the next smaller gauge available, this is selected.

Reading now from the lower set of curves gives the value of δ which is 6.5. A, B and C can now be found from the relationship above, as also can the inside and outside diameters from the following:

$$ID = \frac{0.7}{3} \delta \qquad OD = \frac{1.3}{3} \delta$$

$$A = \frac{6.5}{3} = 2.16 \text{ in., } B = \frac{6.5}{9} = 0.72 \text{ in}$$

$$C = \frac{6.5}{10} = 0.65 \text{ in.}$$

$$ID = \frac{0.7}{3} \times 6.5 \approx 1.5 \text{ in.,}$$

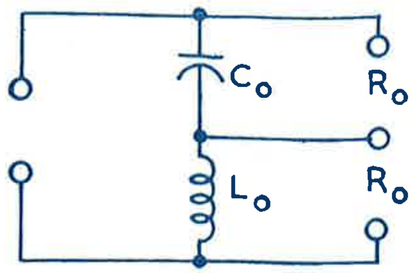
$$OD = \frac{1.3}{3} \times 6.5 \approx 2.8 \text{ in.}$$

From Fig. 2 the number of turns can be determined as 140, and from Fig. 3 the resistance as 0.27 ohm. From Fig. 1 it can now be seen that the "Q" will be at least 25.

Example (2). Because of its simplicity the quarter section series network may have appeal to the home constructor. Suppose speakers with 8 ohm voice coils are available and a cross-over of 1200 cycles is desired, it follows that:

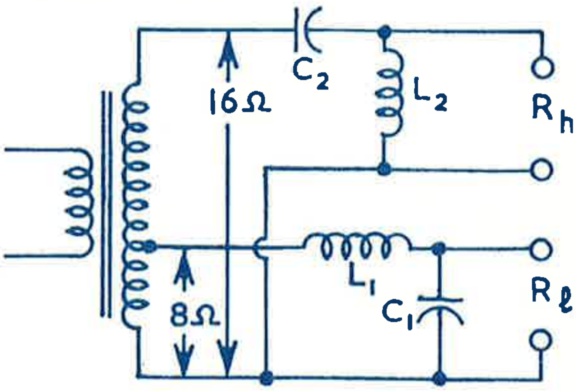
$$L_o = \frac{0.16 R_o}{f_c} \text{ and } C_o = \frac{0.16}{R_o f_c}$$

$$L_o = \frac{0.16 \times 8 \times 1000}{1200} \text{ mH} = 1.06\text{mH}$$



$$C_o = \frac{0.16 \times 10^6}{8 \times 1200} \mu F = 16.6 \mu F.$$

Example (3). Sometimes it is desired, as stated above, to use speakers of differing voice coil impedances; design in this case can proceed as follows:—Assuming the speaker transformer provides taps suitable for the speakers to be used, which have 8 and 16 ohm voice coils, and we choose a cross-over frequency of 2000 c/s.



$$L_1 = \frac{0.23 R_l}{f_c}$$

$$C_1 = \frac{0.11}{R_l f_c}$$

$$L_2 = \frac{0.23 R_h}{f_c}$$

$$C_2 = \frac{0.11}{R_h f_c}$$

$$L_1 = \frac{0.23 \times 8 \times 1000}{2000} \text{ mH} = 0.92\text{mH}$$

$$C_1 = \frac{0.11 \times 10^6}{8 \times 2000} \mu F = 6.88 \mu F$$

$$L_2 = \frac{0.23 \times 16 \times 1000}{2000} \text{ mH} = 1.84\text{mH}$$

$$C_2 = \frac{0.11 \times 10^6}{16 \times 2000} \mu F = 3.44 \mu F.$$

It is often more convenient to select stock sizes of capacitors such as 4 and 8 μF and readjust the values of inductance and f_c to suit, as odd sizes of capacitors are not available. Usually the slight consequential shift in cross-over frequency is of no importance.

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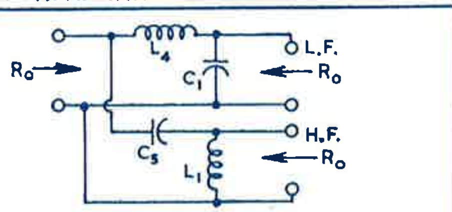
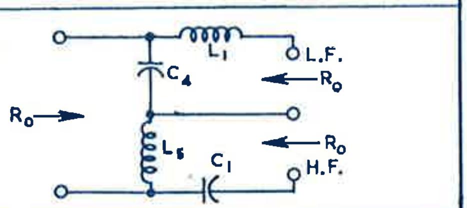
CONVENTIONAL FILTER TYPES

HALF SECTION: ATTENUATION 12 db PER OCTAVE

Parallel

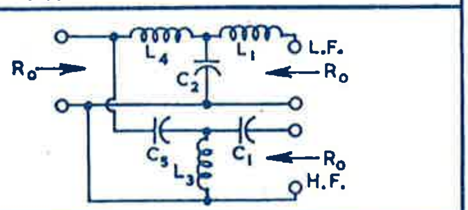


Series

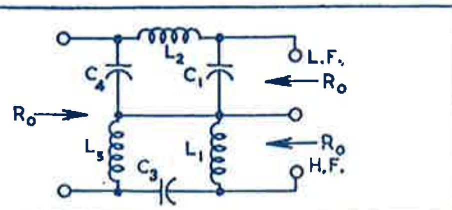


FULL SECTION: ATTENUATION 18 db PER OCTAVE

Parallel



Series



FORMULAE

$$L_1 = \frac{R_o}{2\pi f_c} \approx \frac{0{\cdot}16 R_o}{f_c}$$

$$L_2 = \frac{R_o}{\pi f_c} \approx \frac{0{\cdot}32 R_o}{f_c}$$

{m = 0·6 Design constant}

$$L_3 = \frac{R_o}{2(2\pi f_c)} \approx \frac{0{\cdot}08 R_o}{f_c}$$

$$L_4 = (1+m)\frac{R_o}{2\pi f_c} \approx (1+m)\frac{0{\cdot}16 R_o}{f_c}$$

$$L_5 = \frac{1}{(1+m)}\,\frac{R_o}{2\pi f_c} \approx \frac{1}{(1+m)}\,\frac{0{\cdot}16 R_o}{f_c}$$

$$C_1 = \frac{1}{2\pi f_c R_o} \approx \frac{0{\cdot}16}{f_c R_o}$$

$$C_2 = \frac{1}{\pi f_c R_o} \approx \frac{0{\cdot}32}{f_c R_o}$$

$$C_3 = \frac{1}{2(2\pi f_c R_o)} \approx \frac{0{\cdot}08}{f_c R_o}$$

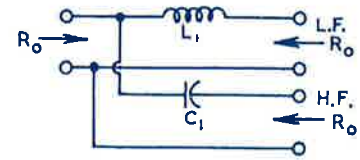
$$C_4 = (1+m)\frac{1}{2\pi f_c R_o} \approx (1+m)\frac{0{\cdot}16}{f_c R_o}$$

$$C_5 = \frac{1}{(1+m)}\,\frac{1}{2\pi f_c R_o} \approx \frac{1}{(1+m)}\,\frac{0{\cdot}16}{f_c R_o}$$

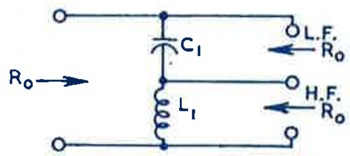
CONSTANT RESISTANCE FILTER TYPES

QUARTER SECTION: ATTENUATION 6 db PER OCTAVE

Parallel

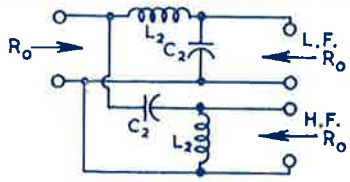


Series

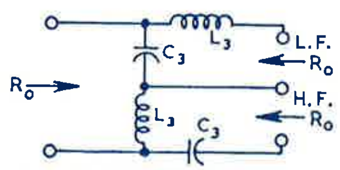


HALF SECTION: ATTENUATION 12 db PER OCTAVE

Parallel

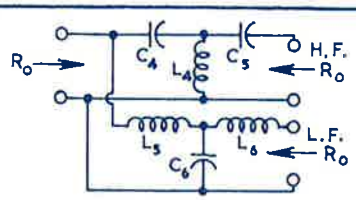


Series

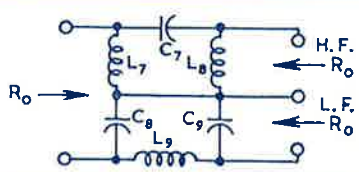


FULL SECTION: ATTENUATION 18 db PER OCTAVE

Parallel



Series



FORMULAE

$L_1 = \frac{R_o}{2\pi f_c} \approx \frac{0.16 R_o}{f_c}$	$C_1 = \frac{1}{2\pi f_c R_o} \approx \frac{0.16}{f_c R_o}$
$L_2 = \frac{\sqrt{2} R_o}{2\pi f_c} \approx \frac{0.23 R_o}{f_c}$	$C_2 = \frac{1}{2\sqrt{2}\pi f_c R_o} \approx \frac{0.11}{f_c R_o}$
$L_3 = \frac{R_o}{2\sqrt{2}\pi f_c} \approx \frac{0.11 R_o}{f_c}$	$C_3 = \frac{\sqrt{2}}{2\pi f_c R_o} \approx \frac{0.23}{f_c R_o}$
$L_4 = \frac{0.75 R_o}{2\pi f_c} \approx \frac{0.12 R_o}{f_c}$	$C_4 = \frac{1}{3\pi f_c R_o} \approx \frac{0.11}{f_c R_o}$
$L_5 = \frac{1.5 R_o}{2\pi f_c} \approx \frac{0.24 R_o}{f_c}$	$C_5 = \frac{1}{\pi f_c R_o} \approx \frac{0.32}{f_c R_o}$
$L_6 = \frac{0.5 R_o}{2\pi f_c} \approx \frac{0.08 R_o}{f_c}$	$C_6 = \frac{1}{1.5\pi f_c R_o} \approx \frac{0.21}{f_c R_o}$
$L_7 = \frac{1.33 R_o}{2\pi f_c} \approx \frac{0.21 R_o}{f_c}$	$C_7 = \frac{1}{2.7\pi f_c R_o} \approx \frac{0.12}{f_c R_o}$
$L_8 = \frac{R_o}{\pi f_c} \approx \frac{0.32 R_o}{f_c}$	$C_8 = \frac{1}{1.3\pi f_c R_o} \approx \frac{0.24}{f_c R_o}$
$L_9 = \frac{1.33 R_o}{2\pi f_c} \approx \frac{0.21 R_o}{f_c}$	$C_9 = \frac{1}{4\pi f_c R_o} \approx \frac{0.08}{f_c R_o}$

Fig. 1.

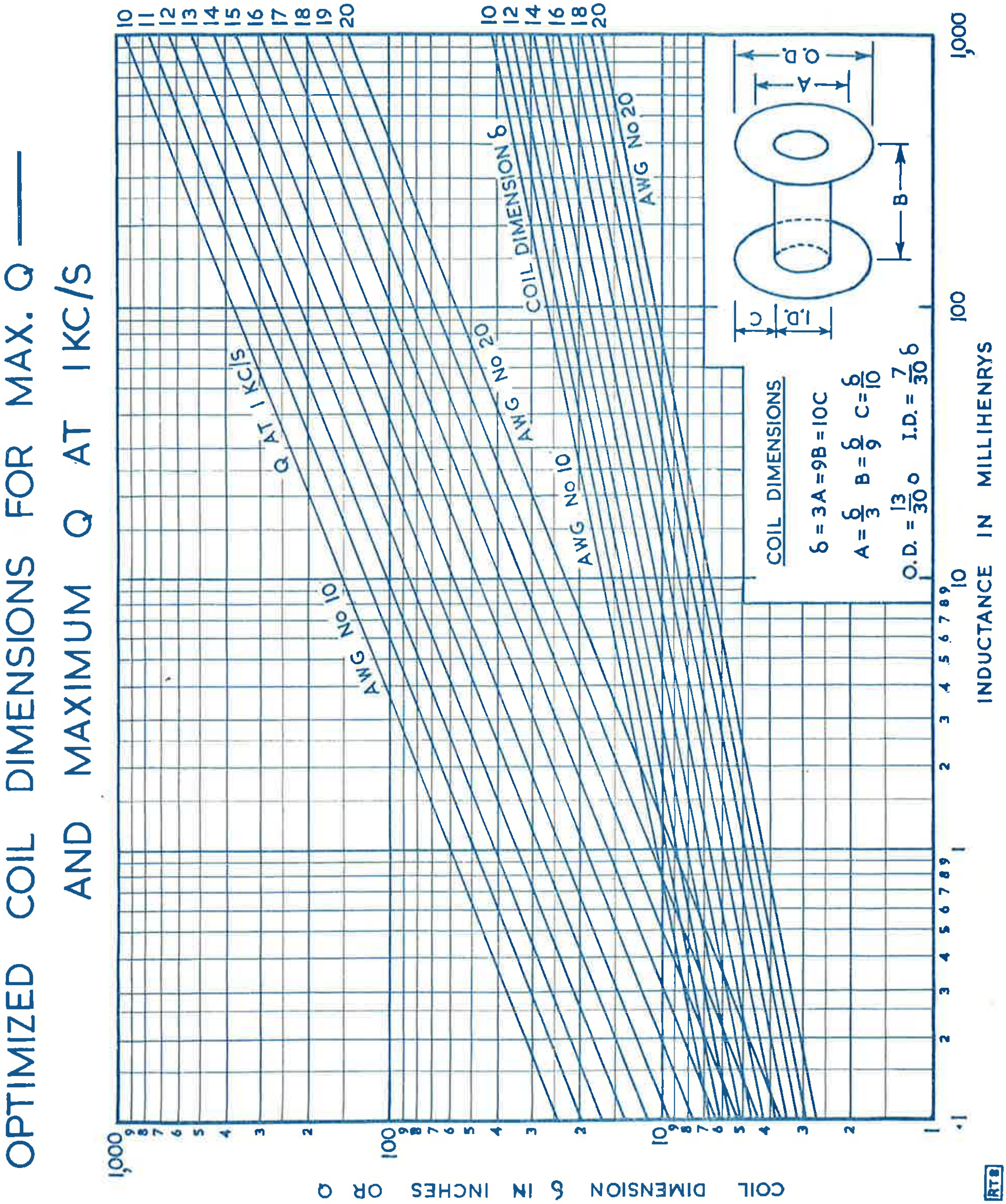

OPTIMIZED COIL DIMENSIONS FOR MAX. Q ——
AND MAXIMUM Q AT 1 KC/S
Q AT 1 KC/S
AWG No 10
AWG No 20
COIL DIMENSION δ
AWG No 10
AWG No 20
COIL DIMENSIONS
δ = 3A = 9B = 10C
A = δ/3 B = δ/9 C = δ/10
O.D. = 13/30 δ I.D. = 7/30 δ
O.D.
A
B
C
I.D.
COIL DIMENSION δ IN INCHES OR Q
INDUCTANCE IN MILLIHENRYS
RTB

Fig. 2.

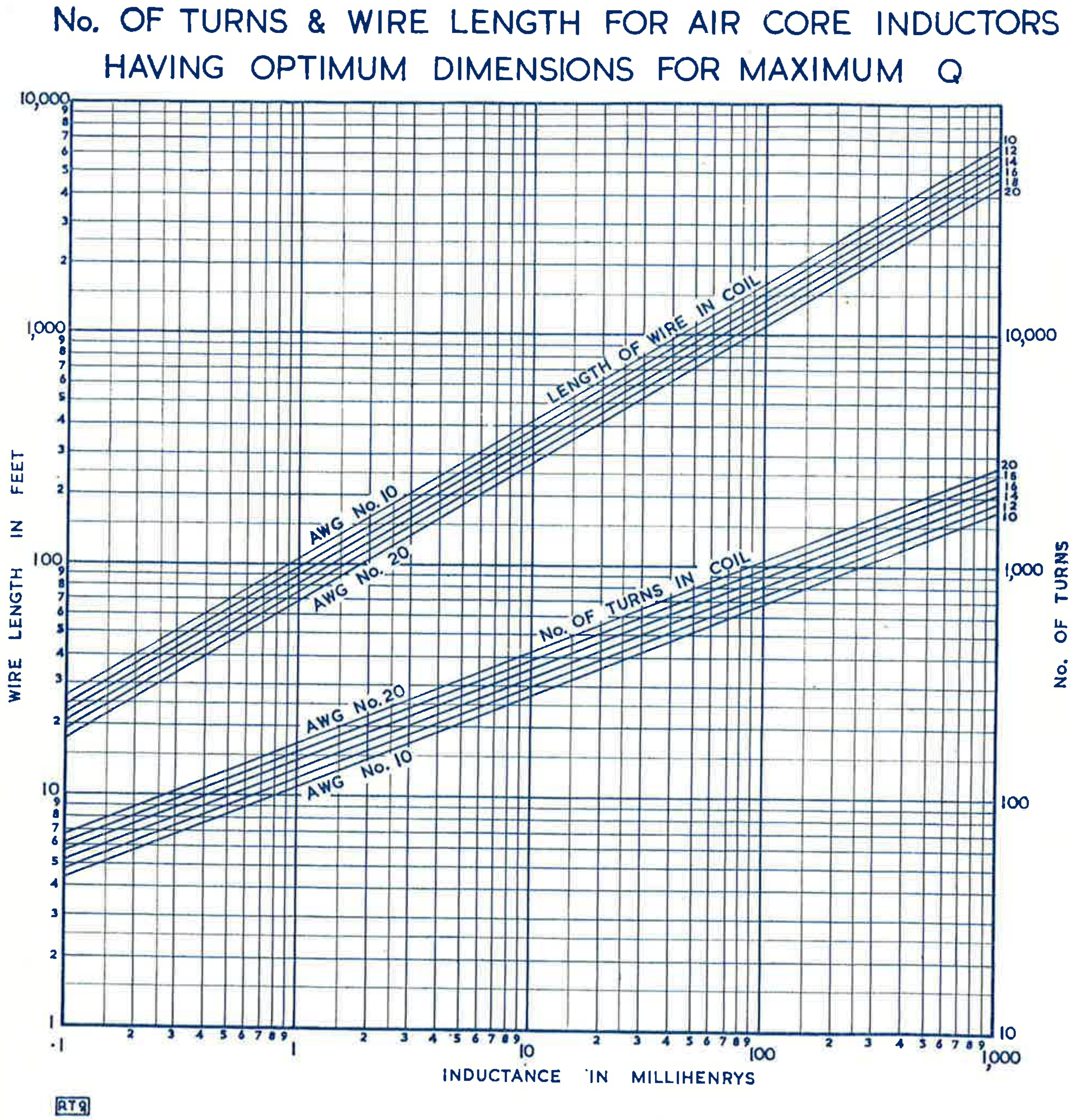
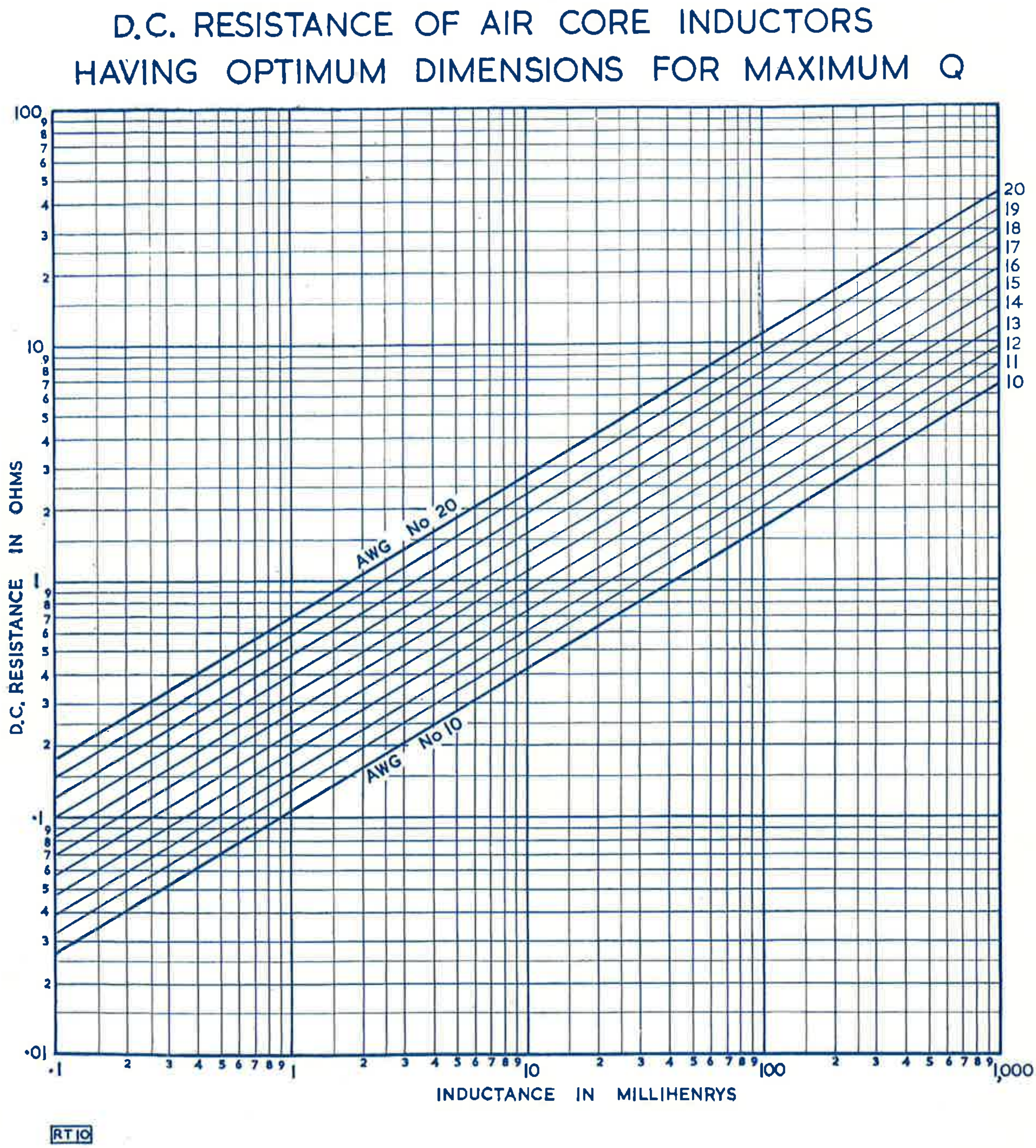

No. OF TURNS & WIRE LENGTH FOR AIR CORE INDUCTORS
HAVING OPTIMUM DIMENSIONS FOR MAXIMUM Q
LENGTH OF WIRE IN COIL
AWG No. 10
AWG No. 20
No. OF TURNS IN COIL
AWG No. 20
AWG No. 10
WIRE LENGTH IN FEET
No. OF TURNS
INDUCTANCE IN MILLIHENRYS
RT9

Fig. 3.


D.C. RESISTANCE OF AIR CORE INDUCTORS
HAVING OPTIMUM DIMENSIONS FOR MAXIMUM Q
D.C. RESISTANCE IN OHMS
INDUCTANCE IN MILLIHENRYS
AWG No 20
AWG No 10
20
19
18
17
16
15
14
13
12
11
10
·01
·1
1
10
100
1,000
RT10

THE USE OF CRACKED CARBON RESISTORS IN AMPLIFIERS

By F. Langford-Smith.

Cracked carbon* resistors have the special characteristics of high stability in resistance, linearity in resistance, close tolerances, low noise, low voltage coefficient, low temperature coefficient, and low inductance (except for the higher resistance values). The question arises, in connection with an amplifier, which resistors should be cracked carbon, and for which positions are ordinary composition resistors satisfactory.

The most important position requiring a cracked carbon resistor is the plate load resistor of a pre-amplifier, particuarly if the input voltage is less than 100 millivolts. Ordinary composition resistors suffer from current noise (resistance fluctuation noise) because there is direct current flowing through the resistor.

A second position for the use of a cracked carbon resistor is the cathode resistor of a pre-amplifier stage, if the resistor is not by-passed.

A possible further position for the use of a cracked carbon resistor is in the grid circuit of a low-level pre-amplifier stage, although our experience indicates that a good quality composition resistor in this position gives equally good results. Since this resistor is not carrying any appreciable direct current, nor is any high voltage applied to it, the use of a cracked carbon resistor is not necessary, but may be regarded as optional.

The final position for the use of cracked carbon resistors is the feedback resistor, when this is taken from the one of the plates in the output stage. The reason for this choice is that a composition resistor suffers from a high voltage coefficient—the resistance decreases as the applied voltage is increased. This variation results in a non-linear characteristic in the feed-back loop, causing increased amplifier distortion. A cracked carbon resistor is not normally necessary when the feedback is derived from the transformer secondary, since the voltages are much lower than from the primary.

* Also known as High Stability Cracked Carbon resistors; Deposited Carbon resistors; Resistors, fixed, composition, Grade 1 (R.I.C.).

Editor .. .. Ian C. Hansen

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