Sept. 30, 1952 P. W. KLIPSCH 2,612,558

CROSSOVER FILTER NETWORK

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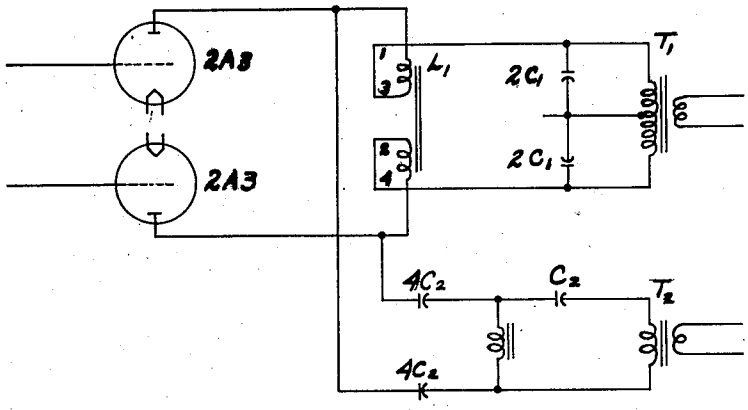

FIG. 1

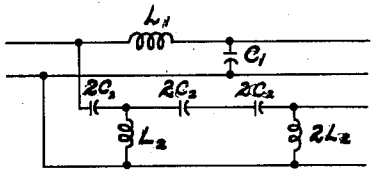

FIG. 2

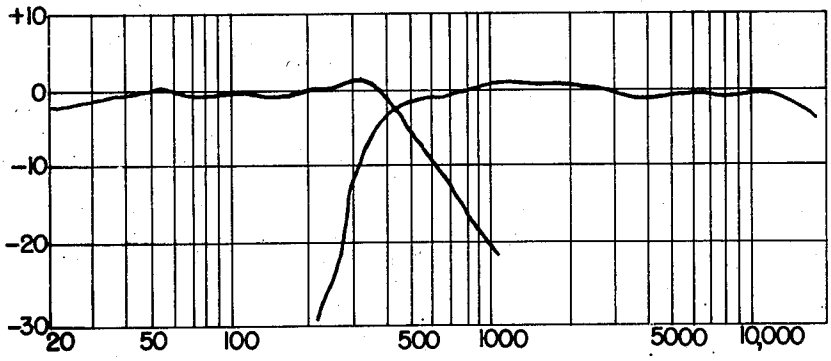

FIG. 3

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CROSSOVER FILTER NETWORK

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3 Claims. (Cl. 178—44)

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This invention relates to electric filters and particularly to crossover filter networks used to allocate power in different frequency ranges to different loads, especially to two or more loudspeakers.

An object is to provide a simplified crossover network of low cost but with frequency response performance comparable with one using costly "high-fidelity" components.

Another object is to provide a high-fidelity crossover network using fewer and less expensive parts than generally required for the same quality.

A further object is to provide a crossover network operable at a high impedance level where capacitor sizes are small and hence inexpensive.

A still further object is to limit overall distortion (both harmonic and intermodulation) of the combined network and transformers to even lower levels than have been possible with conventional designs.

Another object is to provide a network in which the transformers become a part, whereby each of said transformers can be designed for minimal distortion both from the standpoint of smoothness of frequency response as well as harmonic and intermodulation of the transmitted wave forms.

The foregoing together with other objects and advantages will become apparent from the following description considered in connection with the accompanying drawings in which:

Fig. 1 is a schematic circuit diagram showing one form of the invention which has proved highly satisfactory;

Fig. 2 shows another form of the invention adaptable to single ended amplifiers;

Fig. 3 is a frequency response curve, measured on a network designed for 400 cycle crossover.

Design of network

Fig. 2 shows the basis of the design, a ½-section low-pass portion to allocate the low-frequency output to the appropriate load, and a 1½-section high-pass portion to feed the high-frequency load. The choice of the number of sections is based on the fact that a steep cut-off for the low-pass portion is not necessary as most low-frequency loudspeakers will cut off somewhere near the crossover frequency. The high-frequency speakers, when fed power at frequencies below their acoustic cutoff, are very apt to radiate harmonics of the received power as indicated by applicant's paper, "A note on acoustic horns," Proc. I. R. E., vol. 33, No. 7, pp. 447, 449, July 1945.

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Since harmonics constitute one form of distortion, this is to be avoided. Thus a steeper cutoff is provided in the high-pass filter by the choice of 1½ sections. Ideally, the ½-section low-pass portion would produce 12 decibels' loss per octave above cutoff, and the 1½-section high-pass portion would give 27 decibels' loss per octave below cutoff; practically, the loss is somewhat less due to the fact that the reactive elements will not exhibit zero power factor.

Fig. 1 is a circuit, based on the design of Fig. 2, but adapted for push-pull output, that is to say, a balanced line. Inductor L_1 is split so half of it is on each side of the line. This is the circuit as actually used in applicant's developmental sound reproducing system.

Design equations

The numerical design follows conventional practice; see for example, "Radio Engineers Handbook," F. E. Terman (McGraw-Hill, 1943). The impedance level was chosen as 5000 ohms, the nominal load for a pair of type 2A3 tubes operated self-bias. The class A triode was chosen because of its inherently low distortion; with pentode or beam tubes it is imperative that feedback be employed to minimize distortion within the amplifier as well as to produce a low equivalent generator impedance to prevent distortion from being produced by the speakers. The cutoff frequency f was chosen as 500 cycles. The half-section low-pass portion requires elements

$$L_k=\frac{5000}{\pi f}=3.2 \text{ henrys}$$

$$C_k=\frac{1}{5000\pi f}=.13 \text{ microfarad}$$

The isolated half-section causes a nominal loss of 12 decibels per octave, with the intercept of the zero loss line and the asymptote of 12 decibels per octave slope occurring at 1.41 times the nominal cutoff frequency. This intercept may be moved down to the cutoff frequency by doubling L, C, or the LC product.

In the conventional half-section, the capacitance would be $\frac{1}{2}C_k$. This was doubled, to make $C_1=C_k$, so as to give 3 decibels' loss at the nominal cutoff frequency and the desired slope beyond cutoff. Hence, in the figures C_1 is twice the computed value, so that in Fig. 2, $L_1=\frac{1}{2}L_k=1.6$ henrys, $C_1=C_k=0.13$ microfarad, and in Fig. 1, $L_1=\frac{1}{2}L_k=1.6$ henrys reckoned with the 2 windings connected series aiding and $2C_1=0.25$, two of which in series gives the equivalent $C_1=0.13$. This arbitrary change does not appear to have

caused any undue variation in the input impedance.

The high-pass portion is calculated from

$$L_k=5000/4\pi f=0.8 \text{ henry}$$
$$C_k=1/(5000\times 4\pi f)=0.032\ \mu f.$$

The configuration calls for architrave capacitors of $2C_k$=.064 and pillar inductances of L_k=0.8 for the full-section and $2C_k$ and $2L_k$ for the following half-section. Thus $C_2=C_k$. The inductance $L_2=L_k=0.8$, and T_2 has a primary inductance $L_t=2L_2=1.6$.

Construction details

In Fig. 1, the inductance L_1 should be wound with the two windings closely coupled to enforce balance; it has been found satisfactory to put one winding on top of the other. Both L_1 and L_2 should have adequate iron and copper to permit opening up a large air gap to hold the inductances constant at all levels of A.-C. voltage under which they will be required to operate. The matching transformer T_2 can be any readily available transformer of proper ratio which will give the required 2 henrys' inductance when the air gap is opened up sufficiently to hold the inductance nearly constant over all levels of excitation.

Transformer T_1 should have a high primary inductance, of the order of at least 100 henrys to keep the exciting current low at the lowest frequencies to be transmitted, but its leakage inductance need not be held as low as demanded by expensive high-fidelity equipment; the leakage inductance as measured at the primary can be as high as one henry.

Transformer T_2 must have a low leakage, preferable less than 0.05 henry to hold the loss at 10 kilocycles to less than 3 decibels. It may be a low-cost unit, however, since its primary inductance need be only 1.6 henry after opening the air gap, say 100 henrys before adding said air gap.

Formulation of transformer requirements

Let

L_{p1}=primary inductance of low frequency transformer T_1

L_{p2}=primary inductance of high frequency transformer T_2

L_{x1}=leakage inductance of T_1

L_{x2}=leakage inductance of T_2

R_p=plate impedance of the driving amplifier

R_L=load impedance at the plate side of the transformers

$\omega=2\pi f$ where f_1=lowest frequency to be transmitted; f_2=highest frequency to be transmitted; f_x=crossover frequency

D=the per-unit distortion.

It can be shown that:

$$L_{p1}\geqq R_p/3\omega_1 D \qquad L_{x1}\geqq (R_p+R_L)/2\omega_x$$
$$L_{p2}\geqq R_p/3\omega_x D \qquad L_{x2}\geqq (R_p+R_L)/2\omega_2$$

If a single transformer were used,

$$L_p\geqq R_p/\omega_1 D \qquad L_x\geqq (R_p+R_L)/2\omega_2$$

If we put distortion at ½% or D=0.005, F_1=30 or ω_1=187, and f_2=14,000, ω_2=88,000, the values would be L_p=710 and L_x=0.04, a ratio of over 17,000 to one. That would be difficult if not impossible of achievement in the present state of the art. By using 2 transformers, and a crossover of say 500 cycles, ω_x=3140, then the ratios of primary to leakage inductance become of the order of 1000:1 which is readily attainable even in low cost designs. Actually the 2 transformers will not only result in less distortion, and at least as good a frequency response, but also cost less than a single transformer of the high fidelity class.

Transformers T_1 and T_2 should have the proper turns ratios to match the 5000-ohm filters to the respective loads.

It should be understood that all the inductive elements must be designed to carry their exciting currents without distortion, in the same way that any output transformers should be so designed. The air gaps in the inductors should be large so that harmonics of the exciting voltage contained in the exciting currents are small; the transformers should have high primary inductances to minimize the exciting current. If these precautions are not observed, intermodulation distortion may be so high as to render the 2 speaker system less than desirable than a single speaker.

Collecting and tabulating the constants for Fig. 1 gives

Impedance level	5000 ohms
Crossover frequency	500 cycles
Inductance L_1	1.6 henrys (series aiding)
Inductance T_1	100 henrys minimum
Inductance L_2	0.8 henry (with air gap)
Inductance T_2	1.6 henrys (with air gap)
Capacitance C_1	0.128 microfarad
or $2C_1$	0.256 microfarad
Capacitance C_2	0.064 microfarad
or $4C_2$	0.128 microfarad
Leakage ind. of T_1	Not over 0.8 henry
Leakage ind. of T_2	Not over 0.05 henry

Tolerances of ±5% or even 10% may be permitted.

Conversion of the constants given for other impedance levels and/or other crossover frequencies is well known and will not be discussed.

The fact that a pair of 2A3 or 6 A5G tubes exhibit a plate impedance of only 2000 ohms when the load impedance and the surge impedance of the filters are 5000 ohms raises the question of reflections. The mismatch is not great enough to cause serious reflection from a theoretical standpoint. A measured performance curve, Fig. 3, shows each output to be flat in its transmission range and to fall off smoothly in its attentuation range.

This curve was run on a network, using the actual speakers as connected loads rather than using fixed resistors for such loads. Variations in voice coil impedance of the speakers, account for the minor variations in output, in no case exceeding 4 decibels and this extreme only at the extreme frequency limits. This curve was for a 400 cycle crossover frequency.

The aforementioned mismatch is necessary for proper loading of the output tubes and if this is a necessary evil from the filter standpoint it is of sufficient unimportance in the present application to be disregarded. It should be remembered that speaker voice-coil impedance is itself highly variable.

Performance

This crossover network has been in experimental use with a speaker combination comprising a low-frequency horn as described in applicant's paper "A low frequency horn of small dimen-

sions," Jour. Acous. Soc. Am., vol. 13, No. 2, pp. 137–144, October 1941. See also U. S. Patents 2,310,243 and 2,373,692. This speaker was originally designed to operate between 40 and 400 cycles and a high-frequency horn operating from 400 cycles up. The high frequency horn was described in applicant's paper "A high quality loudspeaker of small dimensions," Jour. Acous. Soc. Am., vol. 17, No. 3, pp. 254–258 January 1946.

Two types of low-frequency horns and three types of high-frequency horns, all of different acoustic length, have been tried in various combinations. The output appears smooth throughout the transmission range including that in the immediate vicinity of crossover. Phase relations at this point seem to have little or no bearing on performance. Even complete 180 degree reversal of one speaker (in any of the several combinations) makes no difference in performance. Thus phase shifts in the described crossover network or in the speakers themselves may be neglected. Of course, if two identical speakers are used on one channel, they should be properly phased to prevent peculiar radiation patterns.

An obvious advantage of this crossover network is the fact that low-cost components may be used to give the same high-fidelity performance as more expensive equipment.

A less obvious but even more important advantage is that distortion produced by the necessary coupling transformer may be drastically reduced by the use of two transformers and the permissive design of each transformer to cover less than the entire audio spectrum.

The experimental model was constructed at less cost that that of a single high-fidelity matching transformer, and the performance is such that the combined power output is flat within 2 decibels from less than 30 cycles to cover 10,000 cycles.

The distortion is extremely low; intermodulation between 60 and 6,000 cycle waves simultaneously applied are within tolerable listening limits up to the load limit of the amplifier.

Alternately, in some cases, a high pass portion of one section instead of 1½ sections, may be employed. In this case, the capacitor values are altered; in Fig. 1, the capacitor marked C_2 becomes $2C_2$, and in Fig. 2 the third of the three capacitors marked $2C_2$ is omitted by bridging the diagram at that point; also the inductance of T_2 is not reduced to the indicated value in the table by opening up an air gap but is left as high as obtainable, preferably of the order of 50 henries or more to minimize the exciting current.

A network such as described herein, but designed for 400 cycles crossover was described in applicant's article "Woofer-Tweeter crossover network," Electronics, November 1945, and the performance curve for that same network was shown in "Design of two-horn loudspeakers," Electronics, February 1946.

The invention claimed is:

1. A crossover network comprising low-pass and high-pass portions each associated with a matching transformer, said low-pass portion comprising a balanced two-winding inductor with one of said windings in each side of a balanced line, a one-half section filter with said inductor as input element and with equal series capacitors as terminating element, said one-half section filter being connected to one of said matching transformers with high shunt inductance to minimize its exciting current, a connection between the midpoint of said series capacitors and the midpoint of the primary of said transformer, said high-pass portion comprising architrave capacitors and at least one pillar inductor and terminated by the other of said matching transformers with low leakage inductance.

2. In a crossover network for allocating power in different frequency ranges from a power source to different loads, the combination comprising, separate low-pass and high-pass circuits connected to the source of power to be allocated, said low-pass circuit including a pair of closely coupled windings and an output low frequency transformer connected thereto and having a high primary inductance, and said high-pass circuit comprising a high-pass filter and an output high frequency transformer connected thereto and having a low leakage inductance, said last mentioned transformer having a leakage inductance less than the ratio of impedance to the maximum angular velocity to be transmitted, and each of said transformers having a primary inductance equal to or greater than $R/3\omega D$, where R is the source impedance, ω is the lowest angular velocity to be transmitted by the transformer and D is the permissible per-unit fraction of harmonic distortion.

3. A crossover network to be operated between a power source and a load and comprising, low-pass and high-pass filters each terminated by a matching transformer, the two transformers exhibiting the parameter limits

$$L_{p1} \geqq R_p/3\omega_1 D$$
$$L_{x1} \geqq (R_p+R_L)/2\omega_x$$
$$L_{p2} \geqq R_p/3\omega_x D$$
$$L_{x2} \geqq (R_p+R_L)/2\omega$$

where L_{p1} and L_{x1} are respectively the primary and leakage inductances of the transformer terminating the low-pass filter, L_{p2} and L_{x2} are the primary and leakage inductances of the transformer terminating the high-pass filter, R_p is the source impedance, R_L is the reflected load impedance at the transformer primary, ω_1, ω_x and ω_2 are 2π times the lowest, crossover, and highest frequencies to be transmitted, and D is the per-unit permissible harmonic distortion.

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