Load Circuit Stabilizing Networks for Audio Amplifiers*

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The constant-resistance filter, often used to cross over between loudspeakers in a two-way or multiway system, also forms a sound basis for designing a network to inhibit high-frequency instability in an amplifier when the load impedance varies. It further confers immunity against interference fed into an amplifier via the output lead.

INTRODUCTION: Designers of audio amplifiers often include, in series with the output lead, a network consisting of an inductance in parallel with a resistance. This serves to keep the amplifier stable in spite of changes in the output load which, in an audio amplifier, are completely out of the hands of the designer, can be extreme, and often contribute considerably to poor performance of an otherwise good amplifier.

Such a simple network can be most effective, its only limitations being 1) the incidental resistance introduced by the inductor, which limits the damping factor of the amplifier, and 2) the irrational fear of many designers of any inductive component.

Alternatively, a series resistance/capacitance network is sometimes shunted across the output terminals. When the open-loop output impedance is high, as from a common-emitter stage, the network prevents excessive loop gain at high frequencies, due either to an open-circuit load or to the inductance of a loudspeaker load. Also, when the amplifier output impedance is low, the network shunts and lowers the Q of any capacitance shunted across the load, thus damping its series resonance with the inductance produced by falling h_{fe} in the output, emitter-follower transistors.

THE PREFERRED NETWORK

In seeking a rational basis for designing such networks, it occurred to the author that if an additional capacitor is shunted across the output load, as in Fig. 1, the parallel LR network then becomes a simple first-order junction or branching filter, "crossing over" the output above the

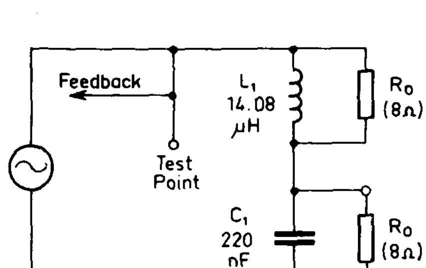

Fig. 1. First-order network. The lower R_0 represents the loudspeaker system by its nominal resistance; the upper R_0 is a resistor incorporated in the network.

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useful audio band from the lower resistance R_0, comprising the loudspeaker system, to the upper resistance R_0, a simple resistor for absorbing high-frequency energy. This network combines the advantages of the earlier networks and additionally filters out RF interference.

To find component values, a loss, say ¼ dB, is allowed at 15 kHz and the capacitance then taken at the nearest "preferred" value. With an R_0 of 8 ohms, this gives a capacitance C_1 of 220 nF, hence a characteristic time constant T_0 (C_1R_0) of 1.76 μs, which in a first-order filter with an inductance of 14.08 μH produces a loss of 3 dB at 90 kHz and 0.1 dB at 15 kHz. If 330-nF capacitors are readily available, they give, with an inductance of 21.12 μH, a time constant of 2.64 μs and a loss of 3 dB at 60 kHz and 0.26 dB at 15 kHz.

These additional losses of high-frequency response may be thought undesirable by connoisseurs of specifications and may seem to destroy features carefully sought by designers of amplifiers. However, the purpose of extended high-frequency response and of good transient response in an audio amplifier is to ensure good performance against overload. It does nothing, of itself, for the audible response. The low-pass filter in no way detracts from the performance of the amplifier so long as it is outside the feedback loop. For example, there could be no possible objection if such a filter, with a loss of ¼ dB at 15 kHz, were incorporated in the loudspeaker instead. In fact, with the filter included in the amplifier as described and thus presenting a much more predictable load impedance to the output stage, the feedback loop can be made more effective at high frequencies. Nevertheless, the true transient performance of the amplifier without masking by the filter is important also, and a test point should be available for checking it as shown in Fig. 1.

In Fig. 1 the numerical value of the inductance is specified in quite unnecessary precision to remind the reader that it can be calculated from the expression

$$L_1 = C_1 R_0^2 \tag{1}$$

as an alternative to the expressions

$$L_1 = R_0 T_0 \tag{2}$$

and

$$C_1 = T_0 / R_0 \tag{3}$$

where

$$T_0 = 1/\omega_0 \tag{4}$$

ω_0 being the angular cutoff (−3 dB) frequency.

When the output terminals are truly terminated with R_0, the network presents R_0 to its input terminals. However, what happens when the output load varies?

Shunt capacitance is usually the most troublesome kind of change in the output load impedance, but the network has already shunted across its output terminals a capacitance whose value is on the high side of what is usually considered the dangerous region for output loading. Certainly this shunt capacitance is so large that it will be changed very little by any likely capacitance in the output lead. Yet this capacitance is an integral part of the network, which ensures that a constant resistance R_0 is presented to the amplifier output so long as the network is terminated by R_0. Probably the most serious mistermination occurs now when the network output is open-circuited.

This open-circuit impedance may be calculated from the expression

$$Z_{OC} = R_0[\omega^2 T_0^2 - (j/\omega T_0)] \,/\, [1 + \omega^2 T_0^2] \tag{5}$$

and plotted as in Fig. 2.

As might be expected, it is reactive and high at low frequencies and resistive, approaching R_0, at high frequencies but, more importantly, its modulus goes through a broad minimum of $0.71R_0$ in the first half-octave above cutoff.

EFFECT OF CHANGING LOAD RESISTANCE

When the load resistance varies, the response varies as shown in Fig. 3. With the nominal load resistance R_0 (8 ohms in Fig. 1) the response falls off as a simple first-order network, with 0.3-dB loss when the normalized frequency ωT_0 or f/f_0 is 0.25, indicated by the arrow in Fig. 3.

When the actual load resistance is doubled to $2R_0$, i.e., with a 16-ohms output load in Fig. 1, the response rises to a low peak 0.5 dB high when ωT_0 is 0.57 and passes through +0.2 dB when ωT_0 is 0.25. When the load is completely open circuit, the response peaks at a +3.3-dB high at 0.86 f_0 and passes through +0.5 dB at 0.25 f_0.

When the load resistance is ½R_0, i.e., with a 4-ohm output load in Fig. 1, the high-frequency response drops off earlier and more severely, with 1.3 dB-loss at $0.25f_0$. Such

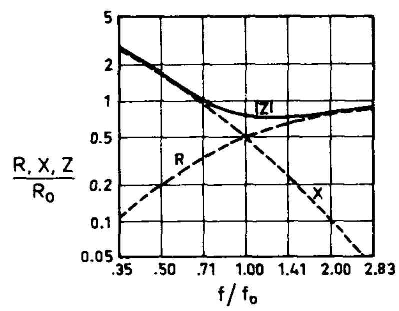

Fig. 2. Impedance presented to amplifier when output of first-order network is open circuit.

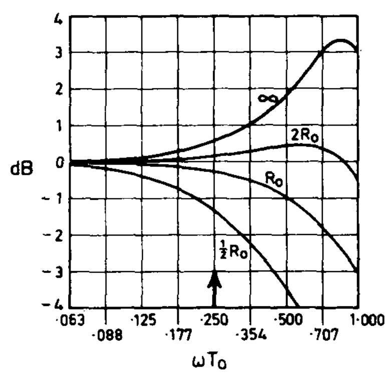

Fig. 3. Variation of first-order network response with output load.

a loss occurs only when the load is both low and noninductive, but it may still be too great in an amplifier intended for a wide range of nominal load impedances.

To meet this condition the crossover frequency f_0 can be moved upward. When f_0 is 120 kHz, for example, ωT_0 is 0.125 at 15 kHz and the loss with $\frac{1}{2}R_0$ load is 0.4 dB. Results may easily be interpolated between these low figures of loss or gain since dB loss or gain varies approximately as frequency squared.

INTERFERENCE SUPPRESSION

The new network also provides a shunt path to earth for signals fed *into* the amplifier from the output lead. Amplifiers are sensitive to signals outside the audio band which are easily rectified by the first stage and thence passed on as audio. Using valves, with a "grid" base of at least 1 volt, the problem was bad enough. In transistors, with a "base" base around 25 mV, the problem is much more serious. Interference must be excluded from the first base-emitter junction, for once it arrives there and is rectified, it cannot be removed later.

Audio amplifiers are the more sensitive to interference induced in their long, usually unprotected, output leads when they use small "phase-advance" or "lead-compensation" capacitors shunting the series feedback resistor. These effectively couple radio frequency directly from the output terminals to the base-emitter junction of the input transistor. They can thus nullify otherwise careful shielding of the input stage and produce the rather ridiculous result that out-of-band interference picked up on the output leads is detected by the amplifier and returned to the loudspeaker as an audible signal.

From the author's experience, most clicks and pops from the mains are induced in this way rather than on an "audio-to-audio" basis. Certainly when the present network was installed in an amplifier, it removed both the interference from a 5-kW medium-frequency transmitter one mile away and clicks from a refrigerator nearby.

Interfering signals are demodulated when a nonlinear relationship exists between input and output, and this must be at least a power law, if not a threshold effect. Thus when an interfering input signal is reduced, e.g., by the filter, the output interference is reduced even more.

HIGHER ORDER NETWORKS

The first-order network of Fig. 1 will provide stability and sufficient interference protection for many applications. However, when greater interference protection is needed, the second-order filter of Fig. 4 can be used. In it the component values are

$$C_2 = 1.414 T_0 / R_0 \tag{6}$$

$$L_2 = 0.707 T_0 R_0 \tag{7}$$

and the output response is a second-order Butterworth which, with the numerical values of Fig. 3, has a loss of 3 dB at 28 kHz and 0.3 dB at 15 kHz. On the other hand, if C_2 is 680 nF and L_2, here calculated from

$$L_2 = \tfrac{1}{2} C_2 R_0^2 \tag{8}$$

is 22 μH, the loss would be 3 dB at 41 kHz and 0.07 dB at 15 kHz. When its output is open-circuited, the network presents an input impedance that goes through a minimum of 0.37 R_0 when the frequency is 0.81 f_0.

Fig. 5 illustrates a further form of the network with a third-order Butterworth response. L_{3A}, L_{3B}, and L_{3C} are, respectively, 2/3, 4/3, and 2 times $T_0 R_0$, while C_{3A}, C_{3B}, and C_{3C} are, respectively, 1½, ¾, and ½ times T_0/R_0. Once more T_0 is $1/\omega_0$ and ω_0 is the angular frequency where the loss is 3 dB. The numerical values, which can be safely rounded to preferred values of capacitance, produce 3 dB loss at 31 kHz and 0.06 dB at 15 kHz. The greater RF filtering of signals from the output lead make this network suitable for the more serious cases of interference.

PRACTICAL CONSIDERATIONS

In constructing any filter and making it work, care must be taken of the "invisible" components not shown in the diagram, such as series resistance and shunt capacitance in inductors and series inductance in capacitors. In the present filter, series resistance in the inductors is likely to be the most troublesome. In the discussion below, a maximum of 6% total series resistance is assumed, thus allowing a deviation of ½ dB in response when the loudspeaker impedance goes from nominal to infinity. A fair allocation to the various sources is 2% to the amplifier output impedance, i.e., a damping factor of 50, 1% to the present network, and 3% to the usual crossover network which operates at much lower frequencies and thus with much larger values of inductance. The dc resistances of the coils must therefore be kept to $0.01R_0$, i.e., 80 mΩ in the first-order network or $0.005R_0$, i.e., 40 mΩ each in the second- and third-order networks. More important from the point of view of coil design, the time constants L/R of the coils must be of the order of 400 μs for the first-order filter and 1600 μs for

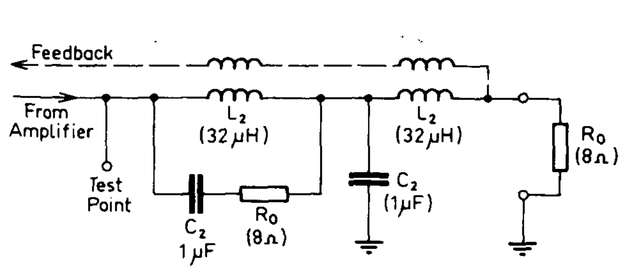

Fig. 4. Second-order network.

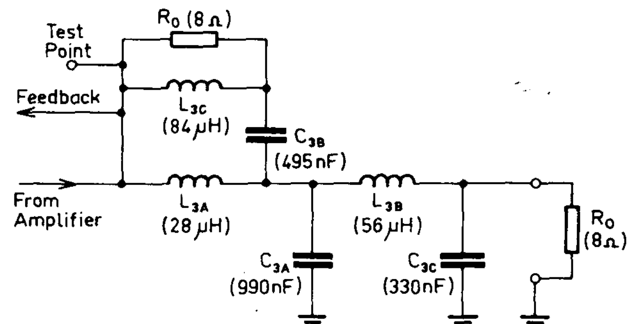

Fig. 5. Third-order network.

the second- and third-order filters. These only require small amounts of copper [1], 15 and 120 g, respectively.

In any case, the time constants may be reduced, if need be, by using somewhat smaller capacitor values and hence smaller circuit time constants but at the price of less protection for the amplifier.

An intriguing, but so far untried, possibility for greatly reducing the effect on the amplifier's damping factor of series resistance in the coils is illustrated by the dashed part of Fig. 4. Extra windings, bifilar with the filter coils, are connected in series to conduct the feedback voltage from the output terminals. The coupling between bifilar windings is extremely tight. Thus the feedback voltage is an accurate version of the amplifier output voltage, except for the IR drops in the coils. In that respect only, the feedback reads the output voltage of the network. Since the feedback circuit is of much higher impedance that the load, the extra windings can be of much finer wire and need contribute little to the size of the coil.

With such large values of circuit capacitance, the self-capacitance of the coils is unimportant. On the other hand, the RF filtering can easily be limited by the self-inductance of the capacitors. Hence the type of capacitors and the method of connecting them in circuit, e.g., length of lead, must be chosen with care.

REFERENCE

[1] A. N. Thiele, "Air Cored Inductors for Audio," *Proc. IREE*, vol. 36, no. 10, pp. 329-333 (Oct. 1975).

THE AUTHOR

Neville Thiele was educated at the University of Queensland and the University of Sydney, graduating as Bachelor of Engineering in 1952. He joined the staff of E.M.I. Australia Ltd. in 1952 as a development engineer in the Special Products Division. During 1955 he spent six months in England, Europe and the United States and on return was responsible for the development of E.M.I. Australia's television receiver. He was appointed advance development engineer in 1957.

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