LETTERS TO THE EDITOR

COMMENTS ON "ON RIAA EQUALIZATION NETWORKS"

One cannot help being impressed by the thoroughness with which mathematician Stanley P. Lipshitz has investigated and presented the theory of disk replay equalization circuitry.[1] He has obviously much enjoyed doing this, just as I have enjoyed the several days required to fully absorb the contents of his paper and to compare some of his results with my own unpublished analyses.

However, while I share with him a general liking for designing circuits correctly on a properly understood theoretical basis, I think nevertheless that there is much virtue in avoiding analytical complexities whenever results of the required precision can be readily obtained in simpler ways. In the present context, the following considerations seem to me to be important:

a) Modern wide-band low-noise integrated operational amplifiers, such as the Signetics NE5534AN or the Mullard TDA1034NB, have such excellent performance that there is little need to allow for the effects of finite gain in the analysis, even when quite high accuracy of response is required.

b) There are other circuit arrangements than those considered by Mr. Lipshitz, which are both economical and much more straightforward to analyze accurately.

c) I do not agree that trimming is "extremely difficult to carry out successfully," provided the right technique is used.

I looked rather carefully into the optimum design of RIAA replay circuits in 1967 on behalf of a British audio firm, and recommended the circuit shown in basic form in Fig. 1. In the event, a more conventional circuit, of the type analyzed by Mr. Lipshitz, was adopted, largely on the grounds that, with the two-transistor amplifier then used, it gave somewhat less distortion at high frequencies because of the increased feedback. With modern operational amplifiers, however, the distortion with the Fig. 1 scheme may be made extremely low, and the design procedure for obtaining an accurate RIAA response is delightfully simple:

1) Decide on a reasonable practical value for R_3.
2) Calculate C_1 from $C_1R_3 = 3180\ \mu s$.
3) Calculate R_1 to give the required zero-frequency (ZF) overall gain, from:

$$R_1 = \frac{10R_3}{9 \times (\text{ZF gain})} .$$

(The zero-frequency gain is nearly enough 10 times the 1000-Hz gain—the precise figure is 9.898.)

4) Calculate R_2 from $R_2 = (R_3/9) - R_1$.
5) Decide on a reasonable practical value for R_4.
6) Calculate C_2 from $C_2R_4 = 75\ \mu s$.

There are no approximations in the above, assuming infinite operational amplifier gain, perfect RIAA response being theoretically obtainable.

In practice a capacitor C_3 will usually be added in series with R_1, for without it the high zero-frequency gain of the circuit, in association with the input bias current and offset voltage of the amplifier, is liable to lead to an output offset that may well be over 1 V. When the 7950-μs bass rolloff of the IEC-amended RIAA characteristic is required, C_3 may be given an appropriately small value, together with slight alteration, if necessary, to other component values, as discussed by Mr. Lipshitz. However, before plunging headlong into the mathematics associated with the component interactions involved, it is worth considering the problem on a simpler basis, as follows.

Of vital importance is the question whether, for a given value of amplifier output voltage V' in Fig. 1, the insertion of C_3 in series with R_1 does, or does not, cause a significant change in the current flowing in R_1 at any audio frequency. If this change turns out to be insignificant, say less than 0.1 dB, then all need for complex mathematical analysis disappears, for the feedback voltage V_{fb}, for constant V', is simply increased above its value without C_3 in accordance with the impedance of the series combination of C_3 and R_1, which should therefore be given a time constant C_3R_1 of 7950 μs.

Consider a typical practical case where a circuit as shown in Fig. 1 has been designed to give an output of 100 mV rms at 1000 Hz for an input of 2 mV rms. The impedance of the complete feedback network will then

[1] S. P. Lipshitz, *J. Audio Eng. Soc.*, vol. 27, pp. 458–481 (1979 June).

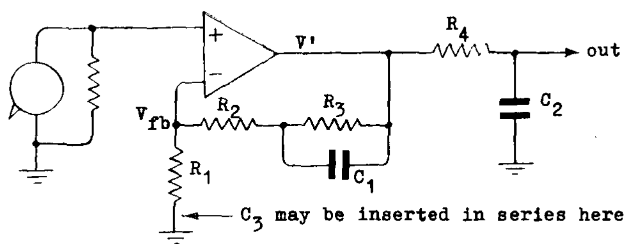

Fig. 1. Basic RIAA equalizer circuit discussed.

vary from about $50R_1$ to about $500R_1$ as the frequency falls from very high values down to zero, in the absence of C_3. Contemplation of this situation then makes it obvious that the insertion of C_3, having a reactance equal to R_1 at about 20 Hz, will cause a change in the impedance of the whole feedback network of much less than 1 part in 50 at any audio frequency; for it is only at quite low audio frequencies that C_3 is of any significance, and at such frequencies the impedance of the complete network is nearer to $500R_1$ than it is to $50R_1$. Thus the effect of inserting C_3 on the current in R_1 will be much less than 1 part in 50 (0.17 dB) and will be more nearly 1 part in 500 (0.017 dB). So we can say, without having to do any detailed mathematics, and without having to consider the phase characteristic of the feedback network, that the departure of the response from the ideal IEC-amended RIAA one, if C_3R_1 is made 7950 μs and the other values are determined as already explained, will be well under 0.1 dB at any audio frequency.

For most practical purposes, therefore, the design of a precision circuit of the Fig. 1 type, including C_3 for the IEC response, can proceed on the basis of items 1) to 6) enumerated above, followed by:

7) Calculate C_3 from $C_3R_1 = 7950$ μs.

A similar simplification of procedure can also often be adopted, of course, in relation to the Fig. 3 circuit in Mr. Lipshitz's article.

If it is desired to analyze the Fig. 1 circuit (with C_3) on a more rigorous basis—which might be considered worthwhile when the overall gain required is a good deal less than in the above example—it may be done as follows.

The feedback voltage V_{fb} in Fig. 1 may be shown to be related to the amplifier output voltage V' by:

$$\frac{V'}{V_{fb}} = \frac{1 + p[C_3(R_2+R_1) + C_1R_3 + C_3R_3] + p^2C_3(R_2+R_1)C_1R_3}{(1+pC_3R_1)(1+pC_1R_3)} . \quad (1)$$

(s may be written in place of the Heaviside operator $p = j\omega$, if preferred; p is strongly defended in Head and Mayo.[2])

The most obvious thing to do next is to factorize the numerator, which involves solving a quadratic equation, so that the relation may be got into the form

$$\frac{V'}{V_{fb}} = \frac{(1+pT_1)(1+pT_4)}{(1+pT_2)(1+pT_3)} \quad (2)$$

where the time constants T_1 to T_4 have the same significance as in Mr. Lipshitz's analysis.

The frequency response corresponding to transfer function (2) may be represented by the straight-line asymptotes shown in Fig. 2.

T_2 and T_3 have simple and direct relationships to the component values, as is seen on comparing Eqs. (1) and (2), but the expressions for T_1 and T_4 in terms of component values turn out to be cumbersome, as also do the corresponding expressions for the component values in terms of T_1 to T_4—not quite so cumbersome as the expressions in Mr. Lipshitz's Table 3, however, for his more elaborate circuit involves solving a cubic equation rather than a quadratic one.[3]

Now it seems to me that when a piece of circuit analysis threatens to yield inconveniently complex results like this, the best thing to do is to sit back and consider whether the analysis can be simplified at the beginning, without significant loss of accuracy even for precision design purposes. In the present instance, it is indeed possible to simplify matters a great deal by taking into account the following two facts:

a) The ω_1 corner frequency (at which the overall gain has fallen to unity) is extremely low, much less than 1 Hz in most practical designs.

b) The precise value of ω_1 is of no interest whatsoever to a design engineer.

Thus referring again to Eq. (1), the 1 in the numerator may be omitted for all practical purposes, since its presence mainly ensures that the response levels off below ω_1 instead of continuing downward in accordance with the broken line in Fig. 2.[4] Thus we may use Eq. (3) in place of Eq. (1), still retaining very high accuracy at audio frequencies (AF):

$$\left[\frac{V'}{V_{fb}}\right]_{\text{at AF}} \approx \frac{p[C_3(R_2+R_1) + C_1R_3 + C_3R_3] + p^2C_3(R_2+R_1)C_1R_3}{(1+pC_3R_1)(1+pC_1R_3)}$$

$$C_3R_1 = T_2 = 1/\omega_2 \qquad C_1R_3 = T_3 = 1/\omega_3 \quad (3)$$

Above ω_3 the two denominator factors contribute a 40-dB/decade falloff in response with rising frequency,

[2] J. W. Head and C. G. Mayo, *Unified Circuit Theory in Electronics and Engineering Analysis* (Iliffe, 1965).

[3] I was caused much puzzlement by the appearance in his Table 1(b), (c), and (d) of relationships such as $R_1/R_2 = R_1C_1/R_2C_1$, which I thought at first must be misprints, since the C_1 appeared to cancel, yielding the unhelpful information that $R_1/R_2 = R_1/R_2$! Later it dawned on me that one could get R_1C_1 and R_2C_1 into the form of functions of T_3 to T_6, from the information in the table, thus producing an expression for R_1/R_2 in terms of T_3 to T_6, which turns out to be so large, however, that it could not be fitted into the available space in the table. Mr. Lipshitz was evidently using expressions such as R_1C_1/R_2C_1 to avoid this difficulty, but I do feel that some explanation was called for.

[4] There is also a very small effect on the ω_4 (318 μs) corner frequency, which is when the sum of the first and last terms in the numerator of Eq. (1) becomes equal in magnitude to the middle or p term. At this frequency, even in a design giving a 1000-Hz gain as low as 10, the magnitude of the p^2 term at ω_4 is approximately 2500, and it is even higher in the higher gain designs. Thus the effect of neglecting the 1 in the numerator is unlikely in practice to exceed 1 in 2500, or 0.0035 dB.

countered, between ω_3 and ω_4, by the 20-dB/decade rise due to the p term in the numerator. This accounts for the 20-dB/decade falloff in the overall response between ω_3 and ω_4 shown in Fig. 2. At ω_4 the p and p^2 terms in the numerator are of equal magnitude, and above ω_4 the p^2 term is dominant, balancing the p^2 in the denominator and giving a level response. Since $p = j\omega$, equality of the two numerator terms occurs at ω_4 given by:

$$\omega_4 = \frac{C_3(R_2 + R_1) + C_1R_3 + C_3R_3}{C_3(R_2 + R_1)C_1R_3} \quad . \tag{4}$$

Since $1/\omega_4 = T_4$, $C_1R_3 = T_3$, and $C_3R_1 = T_2$, these may be substituted in Eq. (4), which with some algebraic rearrangement, leads to the result:

$$R_2 = R_1\left[\frac{T_3T_4}{T_2(T_3 - T_4)} - 1\right] + R_3\left[\frac{T_4}{T_3 - T_4}\right] \quad . \tag{5}$$

Putting in the values $T_2 = 7950\ \mu s$, $T_3 = 3180\ \mu s$, and $T_4 = 318\ \mu s$ gives:

$$R_2 = R_3/9 - 0.9556R_1 . \tag{6}$$

Thus a more precisely accurate procedure for designing the Fig. 1 circuit to give the IEC-modified RIAA response is to carry out items 1) to 7) as already described above, except that, for item 4), Eq. (6) is used in place of the original equation.

Though most commercial control units provide only one fixed value of gain in the disk input circuit, there is really a very strong case for providing at least two fixed gain settings. Satisfactory results can then be obtained with pickups of exceptionally low sensitivity, playing disks with lower than usual peak recording levels, without there being a danger of clipping when using very-high-sensitivity pickups with heavily recorded disks. Pickups do exist with sensitivities of over 3 mV/cm/s, and on a disk with a peak instantaneous velocity at 1000 Hz of 40cm/s, the output from a disk circuit with a 1000-Hz gain of 100 will be over 12 V—if it can swing as much as this. It is most undesirable that such a level should appear on the Tape Record socket of a control unit, for it would overload many tape-recorder input circuits. It is also more pleasant for the user if the normal position for volume controls is not too near to the zero-volume setting.

The practical circuit of Fig. 3 has been basically designed for a 1000-Hz gain of 10, obtained with the switch in the top position. The actual gain, with the nominal preferred values used, is 10.18. Moving the switch wiper downward increases the gain in 10-dB steps, which are correct to within 0.3 dB with the nominal resistor values shown. These values are chosen so that the switch wiper is fed from a source resistance of value close to 6.8 kΩ in all three positions.

Starting with R_3 = 68 kΩ and R_1 = 750 Ω, the ideal calculated values for the other components are as shown in Table 1.

$R_2(A)$ is the approximate value [see item 4], $R_2(B)$ being the more accurate value calculated from Eq. (6).

Some readers may consider the error in C_3 to be undesirably large, though in view of the rather arbitrary nature of the IEC decision to recommend the 7950 μs bass rolloff, one cannot really produce any very rational argument for correcting it. The nominal corner frequency with the Fig. 3 values is 21.22 Hz instead of the ideal 20.02 Hz. In any case, since C_3 will in practice normally be a tantalum electrolytic, with a ± 20% tolerance, the production limits for the bass rolloff corner frequency with the Fig. 3 circuit are approximately 17–27Hz, unless component selection is carried out.

Fig. 4 shows measured response curves for the Fig. 3 circuit with all component values within ± 1% of the nominal ones. They were obtained using the inverse RIAA circuit of Fig. 5, which was constructed before the advent of Mr. Lipshitz's analysis. The original values, shown in brackets, are nevertheless in accordance with

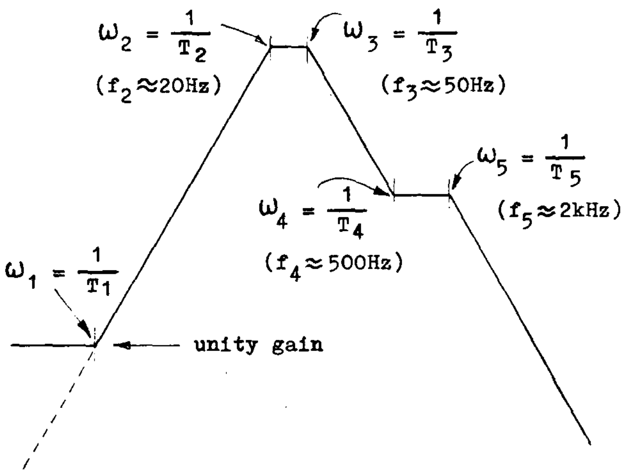

Fig. 2. Theoretical response asymptotes.

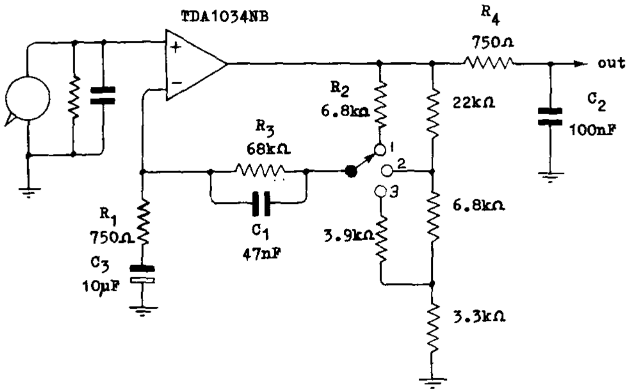

Fig. 3. Practical design for disk input circuit.

Table 1.

Component	Ideal Value	Value Used	Error
C_1	46.76 nF	47 nF	+ 0.50%
$R_2(A)$	6.806 kΩ	6.8 kΩ	− 0.08%
$R_2(B)$	6.839 kΩ	6.8 kΩ	− 0.57%
C_3	10.60 μF	10 μF	− 6%

the formulas in his Table 1(c) for the $R_3 = 0$ condition. I have now, however, corrected the values slightly, in accordance with his formulas for the $R_3 \neq 0$ case, thus allowing for the presence of the 27 Ω resistors in series with the network, and duly acknowledge the usefulness of his contribution in enabling this to be done—even though the corrections in this instance amount to well under 0.1 dB!

The falloff in response above about 40 kHz in the Fig. 4 curves is due to the inverse RIAA network, which gives an unwanted time constant T_6 as enumerated by Mr. Lipshitz. The two lower curves agree very closely indeed with the theoretically predicted results, but a very slight departure from the ideal, of about 0.1 dB, is just discernible in the top curve, particularly at frequencies in the region of 100 Hz. This is caused by the finite open-loop gain of the operational amplifier, which ceases to rise with falling frequency below about 1000 Hz, unlike some other types. At very high frequencies the phase angle of the open-loop gain is nearly in quadrature with that of the feedback network used, giving little departure from the ideal response. It may be noted that with the type of equalization network used in Figs. 1 and 3, the feedback β-value remains well below unity at very high frequencies, making a compensation capacitor for the operational amplifier unnecessary.

In view of the quite small difference between the (*A*) and (*B*) values for R_2 in the table, it is hardly surprising to find that short-circuiting the capacitor C_3 produces an almost perfectly flat low-frequency response in Fig. 4, as indicated by the broken-line curves. The (*A*) value is correct without C_3, whereas the (*B*) value is correct with C_3. The difference in these values would be even smaller if the circuit has been designed for a basic 1000-Hz gain

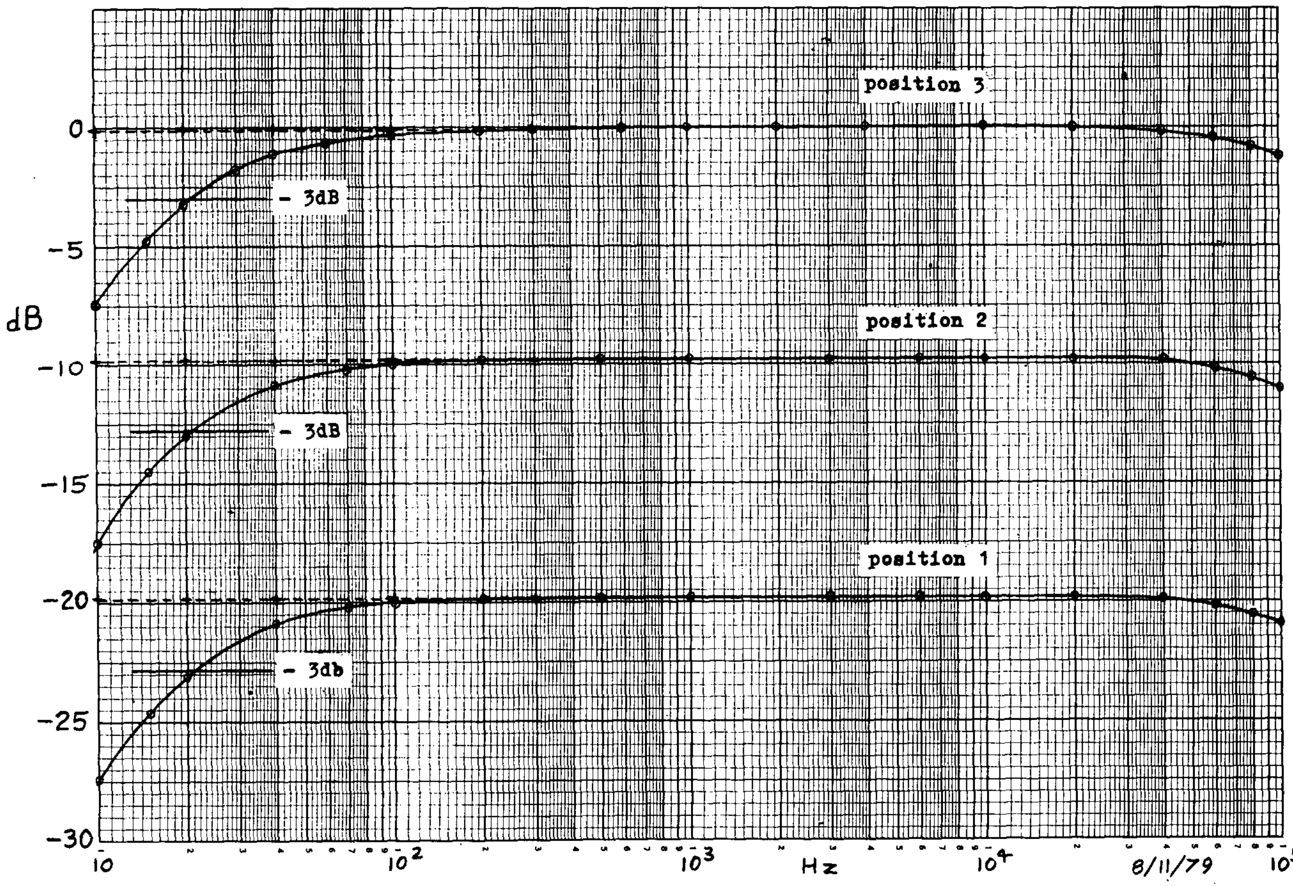

Fig. 4. Measured response curves for Fig. 3 circuit. The broken-line curves were obtained with C_3 short-circuited.

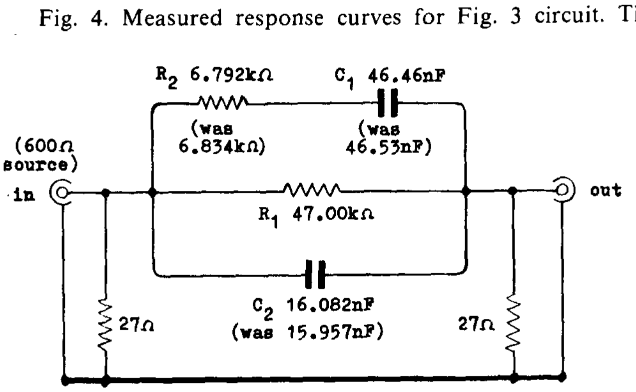

Fig. 5. Inverse RIAA circuit.

greater than the figure of 10 adopted in the Fig. 3 design.

The accuracy of an RIAA equalizer circuit may be quickly checked by feeding it, via an inverse RIAA circuit, from a square-wave source, and Fig. 6(a) and (b) shows the output of the Fig. 3 circuit, set to lowest gain and with C_3 short-circuited, when fed via the circuit of Fig. 5. It is necessary to do such a test at quite a low output level from the RIAA circuit, for the waveform at the output of the inverse RIAA circuit contains very large spikes, such as are never delivered by a pickup, and these can easily cause overloading of the disk circuit, resulting in misleading waveforms. Fig. 6(c), (d), and (e) shows the effects of 5% errors in the values of the network components indicated. The effects are fairly distinctive, and if variable components are provided, it is not very difficult to set such a circuit up, by pure trial and error, to give RIAA equalization accurate at all audio frequencies to within an acceptably small fraction of 1 dB. It is necessary to make sure, of course, that the square-wave generator and oscilloscope are beyond reproach. However, I do agree with Mr. Lipshitz that it is much more sensible to get the equalizer design right by calculation in the first place.

An obvious feature of the Fig. 1 type of circuit, in which it differs from those treated by Mr. Lipshitz, is that the output impedance is not ideally zero. While this may be regarded as a disadvantage, it is hardly so in most practical circumstances, for there will usually be a buffer stage offering a constant high-impedance load to the disk circuit and other signal sources, the buffer output feeding the Tape Record socket and, when required, the volume control. It is easy to arrange matters so that C_2 gives the required 75-μs time constant in association with the parallel value of R_4 and the buffer stage input resistance.

An alternative way to introduce the 7950-μs bass rolloff is to make C_3 so large that it has little effect at audio frequencies, and then incorporate an ac coupling with this time constant in the feed to the subsequent circuits. This has the advantage that a smaller value of capacitor is used to determine the time constant, so that it need not be an electrolytic,[5] but there is the disadvantage that a given rate of rise of supply voltage, on switching the unit on, will produce a bigger output disturbance, or "plonk."

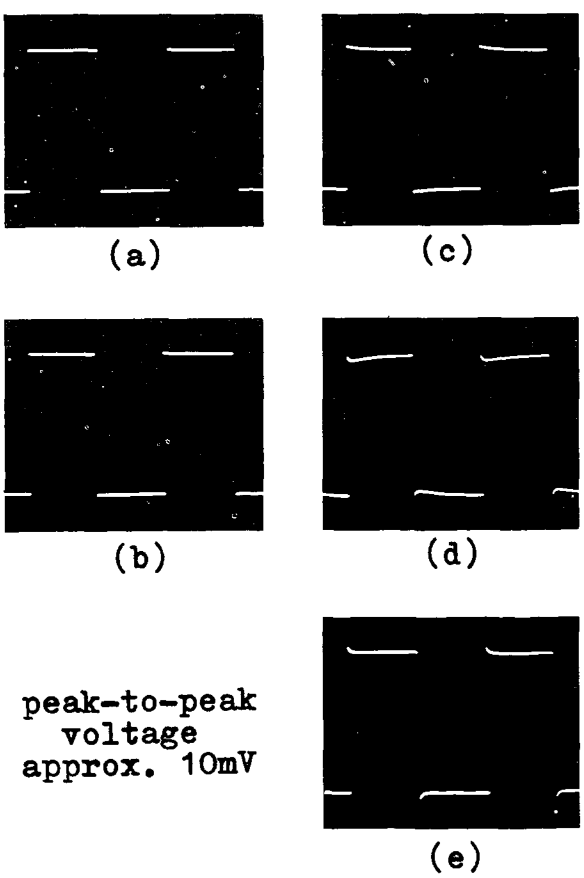

Fig. 6. Output waveforms from circuit of Fig. 3, fed from Levell TG200DM square-wave source via circuit of Fig. 5. (a) 1000 Hz, all values as calculated. (b) 50 Hz, all values as calculated. (c) 1000 Hz, R_4 5% low. (d) 50 Hz, C_1 5% high. (e) 50 Hz, R_2 5% high.

Fig. 7 shows one of three possible CR networks that can be connected immediately after the operational amplifier in a circuit of the Fig. 1 type, in place of R_4 and C_2, to provide the 7950- and 75-μs time constants.

To avoid too much attenuation, R_5 will normally be made much greater than R_4, and this also helps to minimize interaction between the low-frequency and the high-frequency sections, thus permitting the use of simple design formulas with very little error. Under these conditions it would be expected that the two time constants representing the response of the Fig. 7 network would be given fairly closely by:

$$T_2 = C_3(R_4 + R_5) \tag{7}$$

$$T_5 = C_2 \frac{R_4 R_5}{R_4 + R_5} \tag{8}$$

and the approximate design procedure is therefore to choose C_2, C_3, R_4, and R_5 to satisfy these equations when $T_2 = 7950$ μs and $T_5 = 75$ μs. It is of interest, however, to know how much error is introduced by using this simple method of design, and this may be determined without cumbersome mathematics in the following way.

The transfer function for the Fig. 7 network is given in Eq. (9):

$$\frac{V_2}{V_1} = \frac{pC_3R_5}{1 + p[C_3R_5 + (C_2 + C_3)R_4] + p^2 C_2 C_3 R_4 R_5} \tag{9}$$

from which it is easily deduced that at the "midband" frequency, which is the geometric mean of the two corner frequencies, the phase shift is zero, only the p term in the

[5] Another possibility would be to raise the impedance of the whole circuit in Fig. 3, thus permitting a nonelectrolytic capacitor to be conveniently used for C_3, but this has the disadvantage that the noise performance is degraded. From the latter point of view it would, indeed, be slightly beneficial to *lower* the impedances, but unless lowered by a factor of 10, it will be found that other preferred values give less accurate conformity to the correct RIAA time constants.

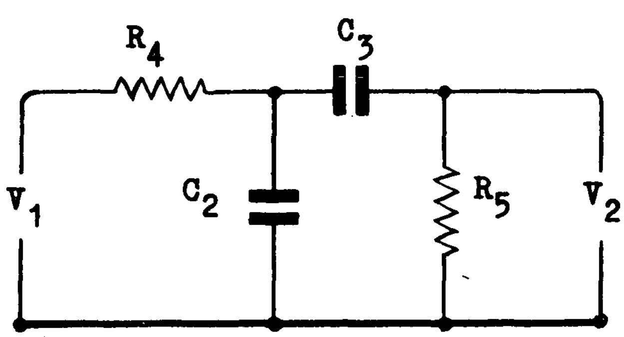

Fig. 7. Circuit for providing 7950-μs and 75-μs time constants.

denominator remaining, and the transmission is given by

$$\left[\frac{V_2}{V_1}\right]_{\text{midband}} = \frac{1}{1 + R_4/R_5 + C_2R_4/C_3R_5} \,. \quad (10)$$

At extremely low frequencies, only the 1 in the denominator of Eq. (9) is significant, so that the low-frequency asymptote (LFA) is the line given accurately by

$$\left|\frac{V_2}{V_1}\right|_{\text{LFA}} = \omega C_3R_5 \,. \quad (11)$$

At extremely high frequencies, only the p^2 term in the denominator is significant, so that the high-frequency asymptote (HFA) is the line given accurately by

$$\left|\frac{V_2}{V_1}\right|_{\text{HFA}} = \frac{1}{\omega C_2R_4} \,. \quad (12)$$

If the circuit gave a low-frequency response following the asymptote (11) at very low frequencies, and having a corner frequency corresponding precisely to the time constant T_2, then the level of the horizontal medium-frequency asymptote (MFA) representing this would be equal to the level reached by the low-frequency asymptote at $\omega = 1/T_2$, which, from Eq. (11), is given by

$$\left|\frac{V_2}{V_1}\right|_{\text{MFA}} = \frac{C_3R_5}{T_2} \,. \quad (13)$$

Since we are choosing the component values to satisfy Eq. (7), it follows from Eqs. (7) and (13) that the horizontal asymptote for the ideal wanted response is as given by Eq. (14):

$$\left|\frac{V_2}{V_1}\right|_{\text{MFA}} = \frac{R_5}{R_4 + R_5} \,. \quad (14)$$

The same result can be obtained by considering the high-frequency asymptote, the time constant T_5, and Eqs. (12) and (8).

Suppose now that a suitable practical value for R_4 is selected, and that R_5 is made 10 times R_4. Eq. (14) then shows that the horizontal asymptote level is at −0.828 dB relative to unity gain. If the total response were equivalent to the sum of two component responses, each perfectly in accordance with the asymptotes, then it may be calculated that, at the midband frequency, each of these responses would be below the horizontal asymptote by 0.041 dB, so that the total response would be below the asymptote by twice this amount, that is, by 0.082 dB. It would therefore be below unity gain by 0.828 + 0.082 = 0.910 dB.

Finally C_3 and C_2 are determined from Eqs. (7) and (8) and the values substituted in Eq. (10) to give the actual amount by which the midband gain is below unity, and the resultant figure is 0.918 dB. It is now seen that the actual response is below the ideal midband response by only 0.918 − 0.910 = 0.008 dB, and the error will be less than this at other frequencies. If R_5 is more than 10 times R_4, there will be an even smaller error, so that under normal conditions Eqs. (7) and (8) may be regarded as highly accurate for design purposes.

In conclusion, I would like to mention that in 1957 a short and excellent Letter to the Editor from W. H. Livy of EMI Studios[6] pointed out an error in an article by another author and gave, concisely and clearly, the correct formulas for Mr. Lipshitz's Fig. 1(a), (b), and (c) networks. It was stated that these designs could be used either in inverting feedback circuits or to provide accurate passive deemphasis.

Also a textbook chapter[7] of which I am the author gives correct design formulas, independently derived, for the above three networks, and points out that significant errors occur if the circuits are used with noninverting amplifiers, and too much resistance is inserted in series with the networks. One or two figures are very slightly in error (by well under 0.1 dB) owing to having used that virtually obsolete device, the slide rule, at the time they were calculated! It is curious that neither Mr. Livy nor myself appears to have been aware of the fourth possible basic network configuration, that of Mr. Lipshitz's Fig. 1(d). In a somewhat similar context, it may be noted that the part of my Fig. 1 circuit involving R_2, R_3, and C_1 may be replaced, if preferred, by a network having a single resistor between the V_{fb} and V' points, shunted by a series combination of R and C.

ADDENDUM

It has been found that if polyester-dielectric capacitors are used in networks such as those of Figs. 3 and 5, then the fall in specific permittivity with rising frequency, and the large loss angle, which are characteristics of this material, produce a detectable high-frequency error and a visible (small overshoot) imperfection in the square-wave response. Polystyrene and polycarbonate capacitors are free from this defect, whose magnitude, however, is sufficiently small to make it of little practical consequence.

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AUTHOR'S REPLY

I would like to thank Mr. Baxandall for his detailed and penetrating comments on the above paper.[8] It is clear that he has devoted a considerable amount of time, thought and care to formulating his comments, and

[6] W. H. Livy, "Disc Replay Equalizers," *Wireless World*, vol. 63, p. 29 (1957 Jan.). The first formula contains an evident misprint; there is a term appearing as t_1t_3/t_3 which should be t_1t_3/t_2. It is given correctly in Table 1.

[7] P. J. Baxandall, "High-Fidelity Amplifiers," in *Radio, TV and Audio Technical Reference Book*, S. W. Amos, Ed. (Newnes-Butterworths, 1977), chap. 14. In the disk equalizer circuit of Fig. 14.19 the input-coupling electrolytic capacitor is shown with the wrong polarity.

[8] See footnote 1, p. 47.

they certainly do represent a significant aid to understanding the behavior of RIAA circuits. I apologize for omitting Mr. Baxandall's contribution[7] from my bibliography; I simply was unaware of it. Having now perused the relevant section of his chapter, I must congratulate Mr. Baxandall on the elegance of the analysis presented there, which is, of course, perfectly correct. The Livy letter[6] was also unknown to me until it was brought to my attention by Mr. A. W. Fitchett of Wellington, New Zealand, subsequent to the publication of my article.

To take up Mr. Baxandall's main points, with which I am substantially in agreement, I would like to comment as follows:

1) I agree that with modern high-gain–bandwidth integrated operational amplifiers such as the NE5534 the effects of finite loop gain can generally be ignored. This is, however, certainly not true of the simple discrete two- or three-transistor (or tube) circuit so common in the past. My purpose in Section 7 was to show how even this limitation need not prevent RIAA accuracy from being achieved.

2) Trimming for accuracy certainly is possible, and the sensitivity Table 6 can serve as a guide to the optimal trimming procedure. Nevertheless the profusion of errors in commercial circuits in the past indicates to me that the designers were either unable or unwilling to achieve accuracy by trimming. I agree with Mr. Baxandall that it is more sensible to get the design right in the first place. (Due to the separation of T_5 from T_1–T_4 in Mr. Baxandall's Fig. 1 design, trimming is inherently easier than it is for the circuit configurations considered in my paper.)

3) Mr. Baxandall's Fig. 1 circuit is a good one, and its reduced overload margin and slightly higher distortion at high frequencies, compared with the standard topology of my Fig. 3, is not a significant problem in practice. His analysis of its design is beautifully clear and valuable. I feel obliged, however, to point out that it is *inherently* an easier design to analyze, for the same reasons mentioned in point 2) above.

4) Although the *precise* value of T_1 is of no great concern, because T_1 determines the dc settling time of the circuit at switch-on, it must be considered in the overall design.

5) The clarification of my use of expressions like R_1C_1/R_2C_1 in Tables 1(b)–(d) and 3(b)–(d) is valuable, and I should have explained my meaning. In Table 1(b), for example, for $R_3 \neq 0$ one should first compute R_2C_2 from the given expression in terms of T_3–T_6, and then use the remaining formulas to evaluate, in order, R_1C_1, R_2C_1, R_1C_2, R_1/R_2, and finally C_1/C_2. Apart from making it unnecessary to display unwieldy formulas, this procedure minimizes the amount of redundant calculation necessary.

6) Mr. Baxandall's optimization of component values in his Fig. 3 circuit, including the clever gain switching without frequency response penalties, is noteworthy, and the overall performance is outstanding in spite of the deceptive simplicity of the circuit. Many North American preamplifier designs unfortunately do not provide a buffer between the RIAA preamplifier and the "tape out" sockets, and so loading of the "tape out" sockets could cause a problem with T_5 in some cases.

7) I must *also* admit to having been unaware of the Fig. 1(d) network configuration until perusing Ref. [36] of my paper. It is not frequently employed for some unknown reason.

Finally I would like to draw attention to two points concerning my paper:

a) In the passive deemphasis circuits of Fig. 7 the loading effect of the subsequent stage can of course be taken into account. Let R_L denote the load resistance on the output circuits in Fig. 7. Apart from a slight gain renormalization by the voltage divider factor $R_L/(R_L + R_1)$, it is now only necessary that R_1 be chosen such that it is the *parallel* combination of R_L and R_1 [that is, $R_LR_1/(R_L + R_1)$], which gives the desired value specified for R_1 in Table 1.

b) The last line of Eq. (25) contains a misprint. The factor $(1 - T_2s)$ should have read $(1 + T_2s)$.

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COMMENTS ON "SHAPED TONE-BURST TESTING"

It was with great interest that I read the above-mentioned Engineering Report[9] by Linkwitz.

For approximately 20 years, in a previous employment while working on envelope compression, expansion, compansion, limiting circuits and on spatial listening investigations, I have found that the use of a pulsed-tone test signal is essential in the determination of transient characteristics. Without these facilities one is completely lost when moving from the traditional steady-state testing into the real world of audio. I have long advocated this method for the testing and evaluation of all audio and acoustic devices and it is gratifying to read similar views expressed in the article.

However, I have a few criticisms, particularly of the method described in obtaining the pulsed or shaped tone burst. My findings over the years are that it is highly desirable to be able to independently vary rise, dwell, fall, and repetition times of the envelope, and that the widths and repetitions need to range about 0.1 ms to about 2 s.

An obvious method of obtaining this specification is to shape the pulse prior to multiplication with the tone in an analog multiplier chip. As it is possible to balance an integrated-circuit multiplier to better than −70 dB relative to the full amplitude burst, this provides a far

[9] S. Linkwitz, *J. Audio Eng. Soc.*, vol. 28, pp. 250–258 (1980 April).

