On RIAA Equalization Networks*[1]

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Most current disk preamplifiers, including some very expensive models, have audibly inaccurate RIAA equalization. This severely limits any conclusions that can be drawn from A/B testing of such preamplifiers. These errors are due in part to the perpetuation in print of incorrect formulas for the design of the RIAA equalization networks commonly employed. Other factors include the existence of an uncorrected high-frequency zero too close to the top of the audio band in many noninverting designs, and failure to take adequate account of the limited available loop gain. The situation is surveyed, and tables taking in account the above problems are given for the design of both inverting and noninverting RIAA deemphasis and preemphasis circuits. Examples are furnished to illustrate the various configurations.

0. INTRODUCTION

This paper has been stimulated by the writer's experiences with disk preamplifiers over the past few years. As readers will be aware, many hypothetical causes have been put forward for the subjectively perceived differences between such preamplifiers when A/B tested against each other, and much mystique currently surrounds their design and evaluation. One fact, however, is indisputable, and that is that frequency response differences exceeding a few tenths of a decibel in magnitude between disk preamplifiers are audible. Such deviations tend to be broad band in extent, since they arise from gain and component errors within the RIAA deemphasis circuit. After examining many disk preamplifiers it has become apparent to the writer that this is a problem of significant, if not major, proportions. It is, moreover, not confined to lower priced components only. Some of the most expensive and highly regarded disk preamplifiers on the market deviate audibly from correct RIAA equalization.

There seem to be three major causes for these errors.

1) What the writer, after examining numerous books and schematic diagrams, can only put down to the use of incorrect design equations for the calculation of the resistor and capacitor values used in the equalization networks.

2) Failure to take into account the fact that there is an additional high-frequency corner in the response of an equalized noninverting amplifier stage (the almost universally used configuration), which causes its response to deviate at high frequencies from that required by the RIAA curve. If this corner is placed too close to the top of the audio band, and no corrective action is taken, audible deviations will occur at high audio frequencies.

3) Failure to correctly take into account the limited loop gain available from the amplifier circuit. Many discrete disk preamplifiers have a loop gain at low frequencies which is inadequate to cause them to adhere to the low-frequency portion of the RIAA curve, while many integrated operational amplifiers display insufficient high-frequency loop gain due to their low gain–bandwidth products.

We shall comment further on these points in the sequel. Point 1) is perhaps the most surprising, for there is nothing

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[1] Since the presentation of this paper to the 61st Convention of the Audio Engineering Society in New York on 1978 November 6 the existence of an as yet unpublished thesis on these topics by Paul Skritek ("Optimierung von Entzerrernetzwerken," Diplomarbeit, Institut für Allgemeine Elektrotechnik, Technische Universität, Vienna, Austria, 1977 March) has been brought to the writer's attention. Skritek's approach is somewhat different, but his thesis includes most of the results reported here. In fact, he develops formulas for a very wide class of networks, including all the RIAA cases, and also considers the effects of finite amplifier gain and gain–bandwidth on their performance. The writer has also discovered an article by James Sugden ("Equalizasion," *Audio Annual 1968* (Link House Publications, Croydon, U.K., 1968), pp. 34–37) which correctly develops formulas for two cases of passive and active networks, and which, after Skritek's thesis and [33], represents the single most comprehensive treatment of RIAA equalization networks yet found in print.

extraordinarily difficult about analyzing the standard RIAA equalization configurations.

In case the reader feels that the writer is grossly exaggerating the widespread nature of the problem, we would like to refer him to [1]–[18], drawn from many diverse sources, in support of our contention.[2] As will shortly become apparent, these circuits all suffer from one or more of maladies 1)–3) without showing signs of any adequate corrective action having been taken. All is, however, not bleak, for we have come across a few circuits which do correct for some or all of these sources of error. Without wanting to play favorites, we list some of these circuits [19]–[29], but they are few and far between.

This paper is intended to answer points 1)–3) by providing design formulas for RIAA networks used both passively and actively around inverting or noninverting amplifier stages, and will also give some guidelines for those cases when the loop gain is insufficient for this factor to be ignored. A search of the literature has failed to turn up much in the way of correct formulas; the only sources found which correctly treat a few particular aspects of the problem are [30]–[33]. (See also footnote 1.) It would therefore appear that the time is ripe for a discussion of this topic in some detail. It is hoped that this paper will help fill the gap.

1. THE CIRCUITS AND THEIR CHARACTERISTICS

As is well known, the RIAA disk recording/reproduction standard specifies equalization time constants of $T_3 = 3180\ \mu s$, $T_4 = 318\ \mu s$, and $T_5 = 75\ \mu s$, corresponding to turnover frequencies $f_3 = 50.05$ Hz, $f_4 = 500.5$ Hz, and $f_5 = 2122$ Hz, respectively.[3] The recent IEC amendment [34] to this standard, not yet adopted by the RIAA, adds a further rolloff of time constant $T_2 = 7950\ \mu s$, corresponding to a frequency of $f_2 = 20.02$ Hz, which is applied only on replay. (The reason for this apparently strange nomenclature will shortly become apparent.) Such equalization is commonly achieved by means of frequency-dependent negative feedback around the disk preamplifier stages. The feedback network generally incorporates one of the four electrically equivalent R/C networks N, shown in Fig. 1, for this purpose. The four networks N are listed in the order of popularity, that of Fig. 1(a) being the most popular configuration, while that of Fig. 1(d) is the least frequently used. Also given are their complex impedance formulas, which are easily calculated (see, for example, [35] or [36]). Throughout this paper we shall assume that the components are labeled such that $R_1 > R_2$ and $C_1 > C_2$. (This results in the apparently "reversed" labeling of network 1(c).) Thus $R_1C_1 > R_2C_2$, and so R_1 and C_1 principally determine T_3, while R_2 and C_2 principally determine T_5.

The networks N can be used actively or passively to perform RIAA pre- or deemphasis functions. Of the possible configurations those which appear to be of the most practical utility are listed in Figs. 2–5.[4] Also shown in Figs. 2–5 is the stylized frequency response (Bode plot) of each configuration, $G(\omega)$ representing the magnitude of the gain at angular frequency ω. At this stage it is assumed that the amplifier shown has infinite open-loop gain and can be treated as an ideal operational amplifier. We shall comment later on the very real restrictions and modifications that are necessitated by practical circuits which do not meet these ideal requirements. Two points are at once apparent.

1) There is an additional unavoidable high-frequency turnover with time constant T_6 (corresponding to a frequency f_6, say) which appears in Fig. 3 even when $R_3 = 0$. This departure from the ideal RIAA deemphasis curve does not arise in the inverting case (Fig. 2), unless we deliberately set $R_3 \neq 0$. As mentioned in the Introduction, the appearance of f_6 has almost universally been ignored in practice. While this is not serious if f_6 is at least two octaves above the audio band, this is frequently not the case, as an examination of the circuits cited in the Introduction will show. We shall see, however, that f_6 can be *exactly* compensated for by adding a passive single-pole R/C low-pass filter at the output of the equalized preamplifier, and thus need not concern us unduly. Another reason for wishing to continue the 6-dB per octave RIAA deemphasis beyond f_6 is to prevent ultrasonic signals (from either tracing distortion or radio-frequency pickup) from reaching subsequent possibly slew-rate-limited stages in the chain.[5] It should also be pointed out that the inclusion of R_3 in any of the active circuits under consideration may be *necessary* to enable them to be stabilized.

2) The addition of capacitor C_0 introduces a further pole/zero pair, namely, ω_1 and ω_2, which provides a low-frequency rolloff in the circuits of Figs. 2 and 3 and thus enables a degree of infrasonic filtering of warp and rumble signals to be achieved. If T_2 is chosen equal to $7950\ \mu s$, the inclusion of C_0 will provide equalization as required by the IEC amendment [34]. The reasons behind our labeling of the RIAA time constants T_3–T_5 is now clear. We shall always assume that $T_1 > T_2 > T_3 > T_4 > T_5 > T_6$.

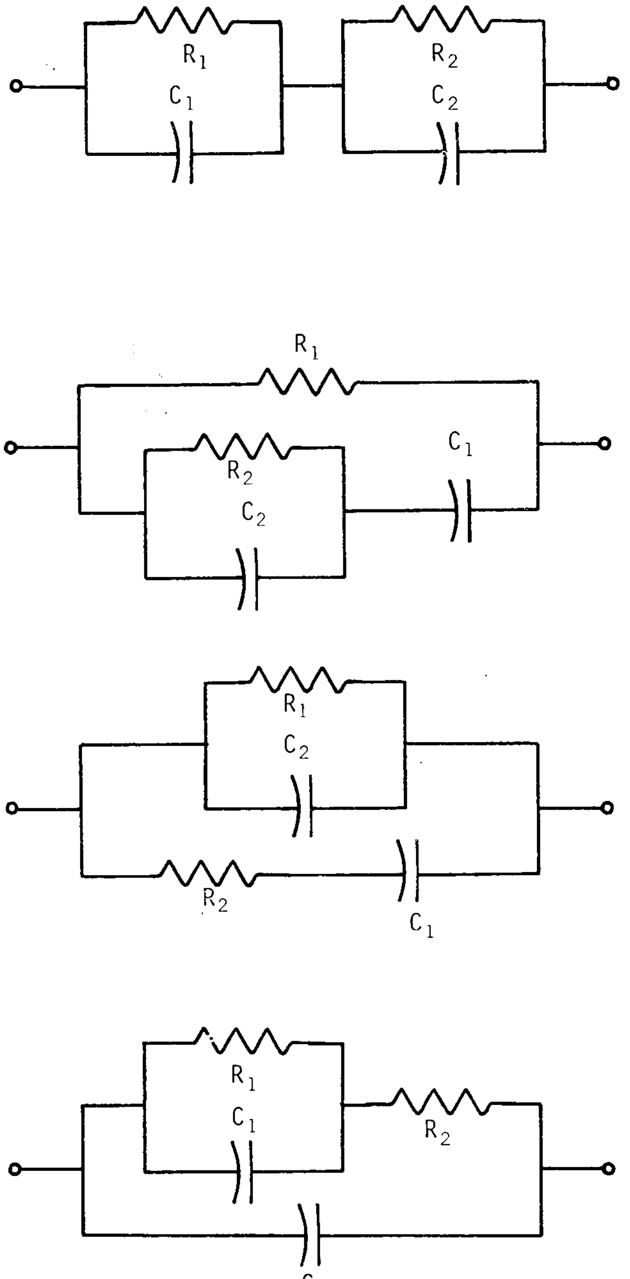
The notes appended to the circuits of Figs. 2–5 will be seen to follow from our calculations in the next three sections. They also refer to the appropriate design table to be

[2] The writer would like to express his appreciation to Walter G. Jung for kindly furnishing him with many of the references cited.

[3] We shall consistently use the symbol f_i to refer to the frequency and ω_i to the angular frequency of a pole/zero of time constant T_i. These quantities are related by $\omega_i = 2\pi f_i$, $\omega_i = 1/T_i$, $i = 1, \cdots, 7$.

[4] The two remaining possibilities which have been omitted for practical reasons are 1) active noninverting preemphasis circuit (this is not feasible due to the enormous HF open-loop gain requirement necessitated by the fact that the minimum signal gain is unity) and 2) passive deemphasis circuit (its wide variation in output impedance renders the circuit of Fig. 2(a) preferable, especially since gain is required in any case); but see Section 8.

[5] For the same reason, the presence of the T_6 corner in the preemphasis circuits of Figs. 4 and 5 is desirable, provided that it lies at least two octaves above the audio band. One cannot continue preemphasizing at 6 dB per octave much beyond this point. Hence R_3 *should* be used in the circuit of Fig. 4, with T_6 carefully chosen.



(a)

$$Z(s) = \frac{(R_1 + R_2)\left[1 + \frac{R_1 R_2}{R_1 + R_2}(C_1 + C_2)s\right]}{(1 + R_1 C_1 s)(1 + R_2 C_2 s)}$$

$$R_A = R_1 + R_2$$

(b)

$$Z(s) = \frac{R_1[1 + R_2(C_1 + C_2)s]}{1 + [R_1 C_1 + R_2(C_1 + C_2)]s + R_1 C_1 R_2 C_2 s^2}$$

$$R_A = R_1$$

(c)

$$Z(s) = \frac{R_1(1 + R_2 C_1 s)}{1 + [R_1(C_1 + C_2) + R_2 C_1]s + R_1 C_1 R_2 C_2 s^2}$$

$$R_A = R_1$$

(d)

$$Z(s) = \frac{(R_1 + R_2)\left[1 + \frac{R_1 R_2}{R_1 + R_2} C_1 s\right]}{1 + [R_1(C_1 + C_2) + R_2 C_2]s + R_1 C_1 R_2 C_2 s^2}$$

$$R_A = R_1 + R_2$$

Fig. 1. Four commonly used equalization networks N.

used for each configuration, and it will be our purpose in these sections to derive the appropriate formulas upon which these tables are based.

2. CALCULATING THE POLES AND ZEROS FOR FIG. 2

In this section we analyze the inverting deemphasis configurations of Fig. 2.[6] The case $R_3 = 0$ will be referred to as the "ideal case," since it is the only one which avoids the undesirable high-frequency zero at ω_6. We thus write down the signal gain equation for the (complex) signal gain $G(s)$ of this circuit (assuming infinite open-loop gain) and find

$$G(s) = -\frac{Z(s) + R_3}{R_0 + 1/(C_0 s)} = -\frac{\{Z(s) + R_3\}C_0 s}{1 + R_0 C_0 s} \quad (1)$$

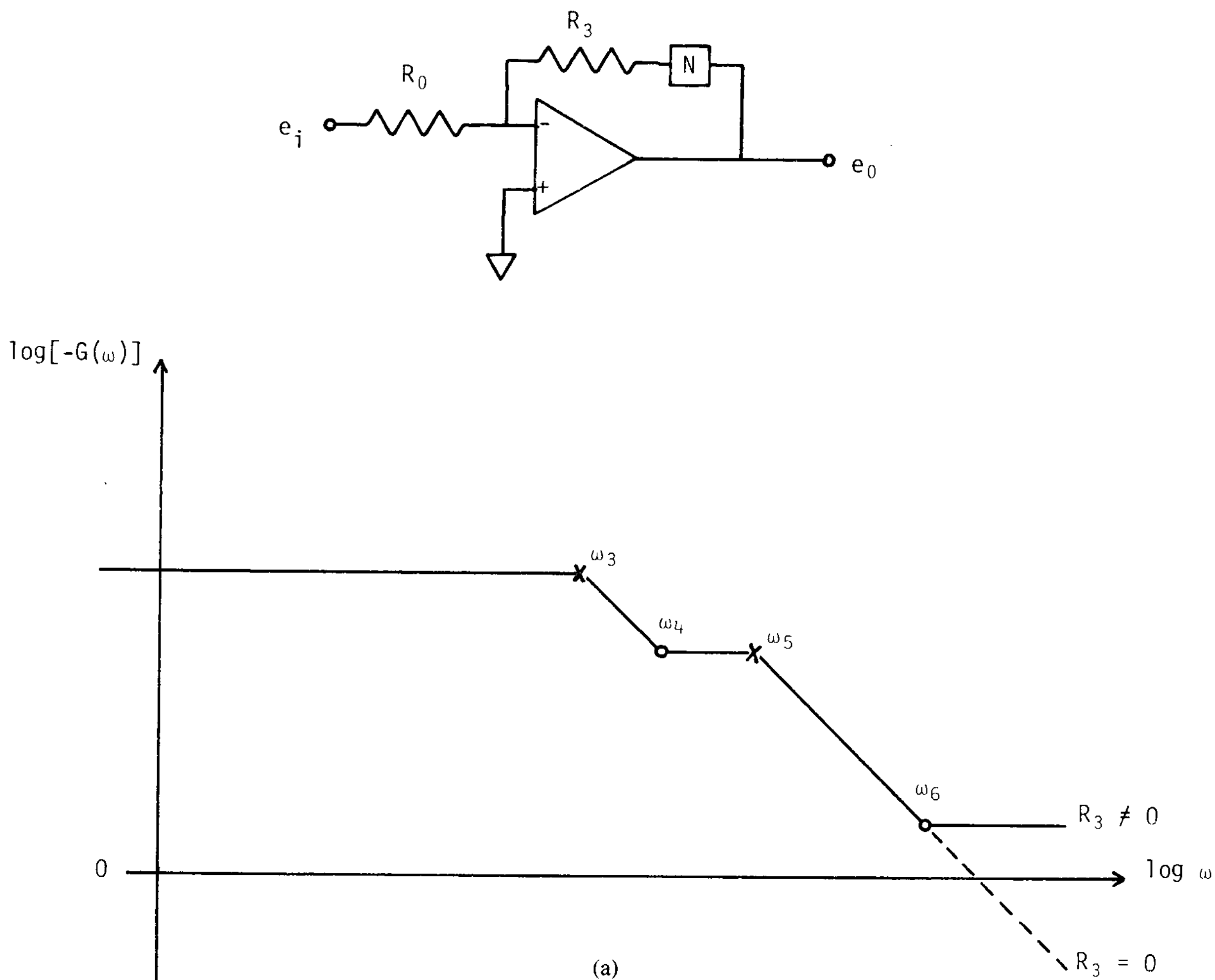
where $Z(s)$ refers to the impedance formulas for the networks N given in Fig. 1. [The case in which C_0 is not present may be obtained by letting $C_0 \rightarrow \infty$ in Eq. (1).] Alternately, we may express $G(s)$ in terms of the time constants T_2–T_6 as

$$G(s) = -\frac{R_A + R_3}{R_0} \cdot \frac{T_2 s(1 + T_4 s)(1 + T_6 s)}{(1 + T_2 s)(1 + T_3 s)(1 + T_5 s)} \quad (2)$$

where the resistance R_A, introduced in Fig. 1, represents the resistance of the network N at 0 Hz (its dc resistance).

Equating the right-hand sides of Eqs. (1) and (2) we can, for each of the four networks of Fig. 1, solve for T_2–T_6 in terms of R_0, R_1, R_2, R_3, C_0, C_1, C_2, thus obtaining formulas for the actually realized time constants of this configuration, and more usefully, we can solve for the

[6] The reader is asked to bear with us through this analysis, for the more common noninverting configurations of Fig. 3 will turn out to be reducible to those of Fig. 2, and the latter are easier to analyze first.

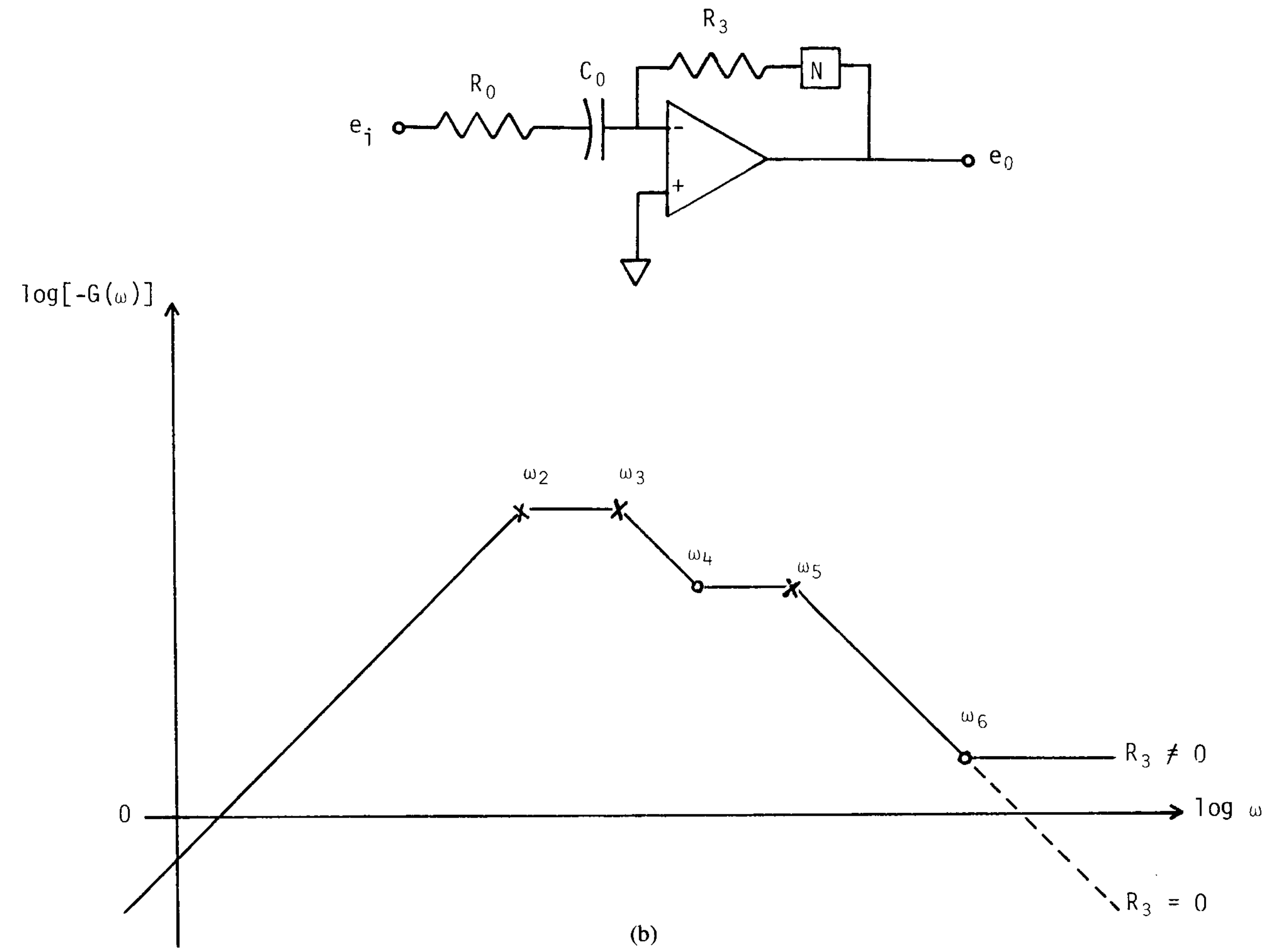


NOTES:
1) R_0 includes source resistance.
2) If $R_3 \neq 0$: Both R_0 and R_3 can be chosen independently. R_0 alone enables gain adjustment without affecting frequency response; R_3 alone determines ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_0, R_3. The ω_6 corner is passively correctable.
3) If $R_3 = 0$: R_0 and *one* of the components of N can be chosen independently. R_0 alone enables gain adjustment without affecting frequency response.
4) Use Tables 1 and 2.

Fig. 2. Active inverting deemphasis circuit. (a) Without C_0.

resistor and capacitor values of the network components in terms of T_2–T_6.[7] These latter formulas can be used in the design of the networks to fulfill the required RIAA function. A different set of formulas results in the case of each of the four networks of Fig. 1. An example of the rather elaborate calculations involved is given in Appendix 1 for the case of the network of Fig. 1(a). The other cases are somewhat more complicated. The results are summarized in Table 1(a)–(d), referring, respectively, to the networks of Fig. 1(a)–(d). The first column in Table 1 gives the design formulas for the ideal case $R_3 = 0$, and the second column lists the corresponding formulas when $R_3 \neq 0$. For simplicity, some of these formulas are given in an approximate form only in the second column. These approximations are to first order in R_3, and are valid to a very high degree of accuracy, provided $R_3 << R_2$, a situation occurring in practice. Table 2 gives the formulas for the magnitude $G(\omega)$ of the complex gain $G(s)$ at angular frequency ω, and is to be used in conjunction with Table 1 in the design process. The design notes appended to Fig. 2 now become relevant. In solving for the formulas given, it is found that both R_0 and R_3 (if nonzero) can be chosen independently. For this reason the formulas in the middle third of Table 1 are "normalized" to give each of the unknown quantities R_1, R_2, C_0, C_1, C_2 in terms of R_0 and R_3 *only*, assuming that the T_i have been chosen in any particular case. Practical design is thus simplified. We shall have more to say about this aspect later. Of considerable significance are the formulas in the first column of

[7] Note that the zeros and poles all lie on the negative real axis in the complex frequency plane.



(b)

NOTES:
1) R_0 includes source resistance.
2) If $R_3 \neq 0$: Both R_0 and R_3 can be chosen independently. Changing R_0 adjusts gain but also affects ω_2. R_3 alone affects ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_0, R_3. The ω_6 corner is passively correctable.
3) If $R_3 = 0$: R_0 and *one* of the components of N can be chosen independently. R_0 alone affects *both* gain and ω_2, but none of the other ω_i.
4) Use Tables 1 and 2.

Fig. 2. Active inverting deemphasis circuit. (b) With C_0.

Table 1, for they represent the ideal RIAA case, and are modified only slightly in numerical value when $R_3 \neq 0$. It should be noted that, not unexpectedly in this simple case ($R_3 = 0$), the formulas for T_3–T_5 are just precisely those for the time constants corresponding to the negative real zero and poles of the impedance expressions $Z(s)$ given in Fig. 1. They also point up what appears to be a very common error committed to print in some of the references cited in the Introduction, and clearly demonstrated by many of the circuits referred to there. For example, the following two situations are not uncommon, and will be found to be represented in the references cited:

1) Use of the network of Fig. 1(a) with the false design equations

$$R_1C_1 = T_3, \qquad R_2C_2 = T_5, \qquad R_2C_1 = T_4 = 318\ \mu\text{s}. \quad (3)$$

As we see from Table 1(a), in fact the network RC products should be (ignoring R_3)

$$R_1C_1 = T_3, \qquad R_2C_2 = T_5,$$

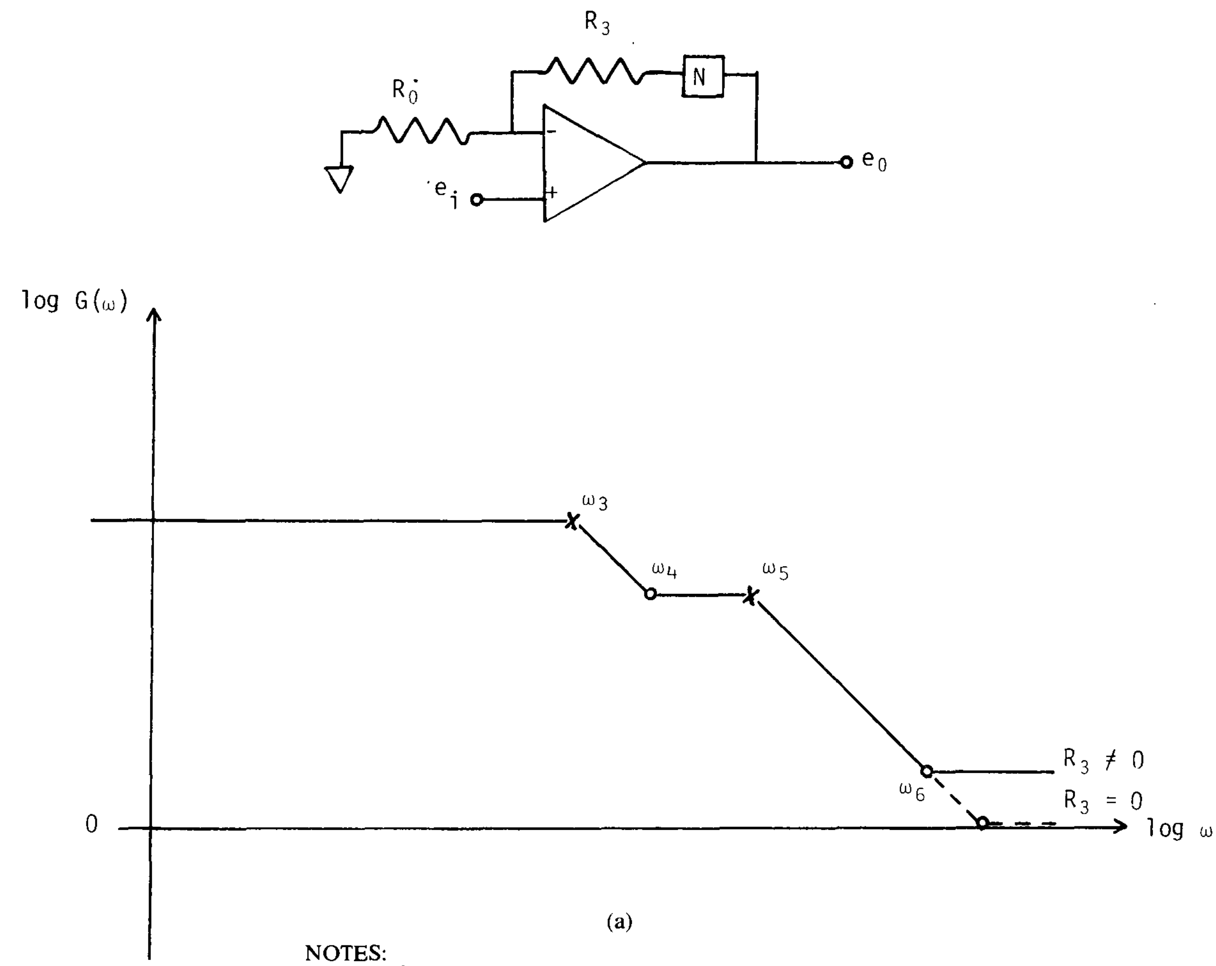
$$R_2C_1 = \frac{T_3(T_4 - T_5)}{T_3 - T_4} = 270\ \mu\text{s}$$

so that the last of formulas (3) is in error by a substantial 18%. This is a very common mistake. The correct formula for T_4, namely,

$$T_4 = \frac{R_1R_2}{R_1 + R_2}(C_1 + C_2)$$

is not difficult to remember, for it represents the time constant of the parallel combination of R_1 and R_2 with the parallel combination of C_1 and C_2.

2) Use of the network of Fig. 1(b) with $R_2C_2 = T_5 = 75\ \mu\text{s}$ instead of the correct value (ignoring R_3)



(a)

NOTES:
1) If $R_3 \neq 0$: Both R_0 and R_3 can be chosen independently. To adjust gain without affecting frequency response, change R_3/R_0 while keeping $(R_0 + R_3)$ fixed. $(R_0 + R_3)$ alone determines ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_0, R_3.
2) If $R_3 = 0$: Only R_0 can be chosen independently. Changing R_0 affects *both* gain and ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_0.
3) The ω_6 corner is passively correctable.
4) Use Tables 1 and 2 with R_3 replaced by $(R_0 + R_3)$ wherever it occurs, and $G(\omega)$ replaced by $-G(\omega)$.

Fig. 3. Active noninverting deemphasis circuit. (a) Without C_0.

$$R_2C_2 = \frac{T_3T_5}{T_3 - T_4 + T_5} = 81.21\ \mu\text{s}.$$

This represents an error of -8%, which is not negligible.

It must thus be realized that the R/C subsections of the networks N interact in determining the overall poles and zeros, and hence the individual RC products for each subsection do *not* give the time constants of the overall network.

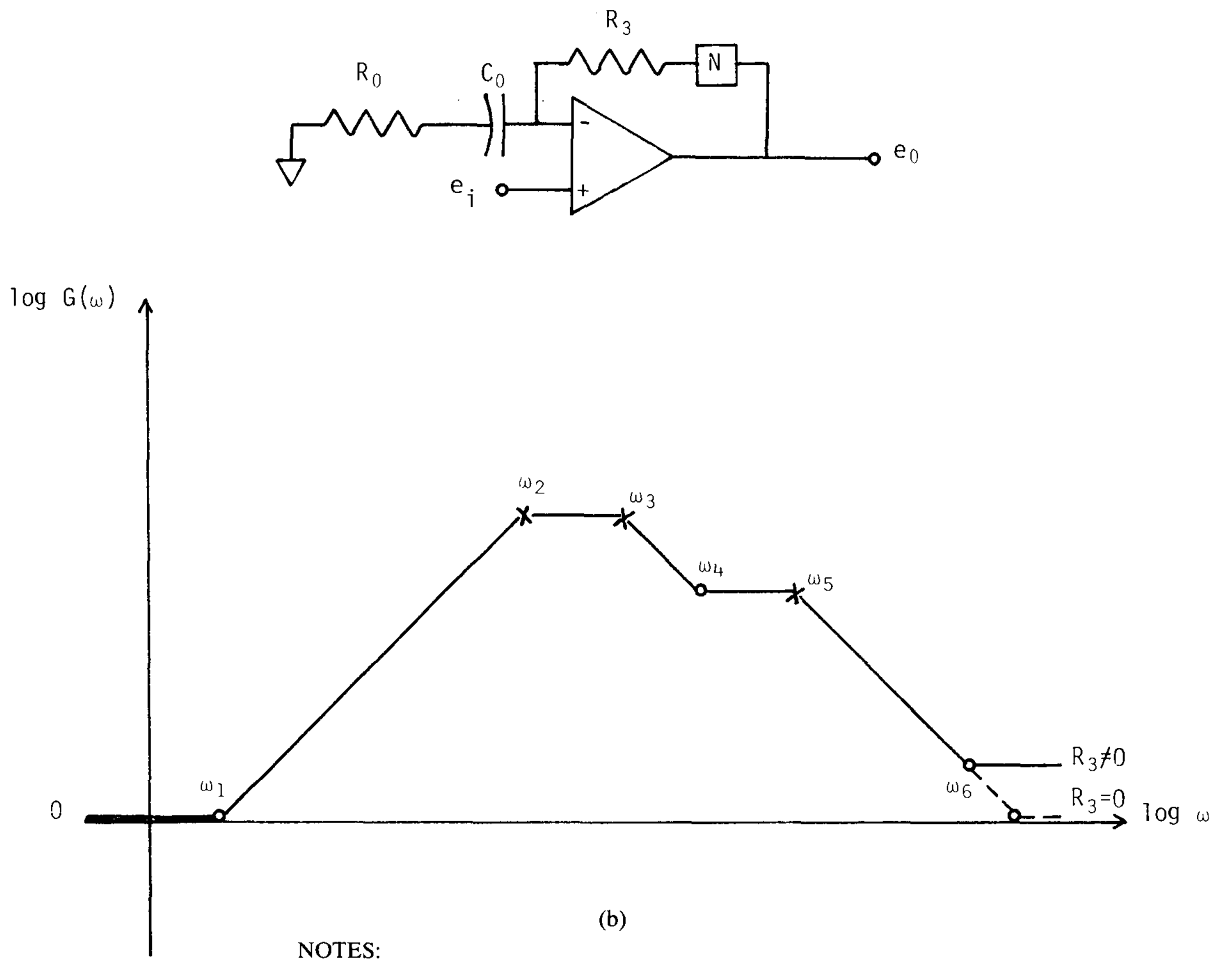
We have placed considerable emphasis on Tables 1 and 2, and for a good reason: With only a few substitutions they will also provide design formulas for the circuits of Figs. 3(a), 4, and 5. Only the circuit of Fig. 3(b) will require a different design table. In fact, since in general R_0 and R_3 should be very much less than R_2 in value, the first column of Table 1 serves as a fairly accurate prototype of the values that will apply in most practical situations. The symbol $[\cdot]$ is used in some of the formulas in Table 1 (and will also be used subsequently) to denote a repetition of the square-bracketed expression that precedes it within the same formula.

As a final note it should be remarked that T_2 is uncoupled from the other time constants T_3–T_6 in the sense that changing or removing C_0 affects only T_2, leaving T_3–T_6 unaltered. (This is not true for the circuit of Fig. 3(b).)

3. EXTENDING THE RESULTS TO FIGS. 3(a), 4, AND 5

The extension of the above results to the active inverting RIAA preemphasis circuit of Fig. 4 is immediate and obvious, for the simple replacement of $G(s)$ by $1/G(s)$ in our previous analysis converts it to this case (that is, the poles and zeros are interchanged). Thus as mentioned in the design notes in Fig. 4, Tables 1 and 2 are easily applied.

The active noninverting RIAA deemphasis circuit of Fig. 3(a) is not much more difficult to handle. For in this case (see Eq. (1)),



(b)

NOTES:
1) If $R_3 \neq 0$: Both R_0 and R_3 can be chosen independently if, say, ω_1 is considered to be dependent. To adjust gain, change R_3/R_0 while keeping $(R_0 + R_3)$ fixed; this also affects ω_2. $(R_0 + R_3)$ alone determines ω_1, ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_0, R_3.
2) If $R_3 = 0$: Only R_0 can be chosen independently. Changing R_0 affects *both* gain and ω_1, ω_2, ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_0.
3) The ω_6 corner is passively correctable.
4) Use Tables 3 and 4.

Fig. 3. Active noninverting deemphasis circuit. (b) With C_0.

$$G(s) = \frac{Z(s) + R_3}{R_0} + 1 = \frac{Z(s) + (R_0 + R_3)}{R_0} \tag{4}$$

which is just precisely the limiting form of Eq. (1) when $C_0 \to \infty$, if we replace R_3 in Eq. (1) by $(R_0 + R_3)$ and delete the minus sign on the right-hand side. Eq. (2) also now applies with the same changes, and so it follows at once that the design Tables 1 and 2 without C_0 also apply directly to the circuit of Fig. 3(a) under the simple substitutions

$$R_3 \to R_0 + R_3 \qquad \text{and} \qquad G(\omega) \to -G(\omega). \tag{5}$$

We see that the poles T_3, T_5 are exactly the same as those of Fig. 2(a); the zeros T_4, T_6 are, however, shifted by the change of R_3 to $(R_0 + R_3)$.

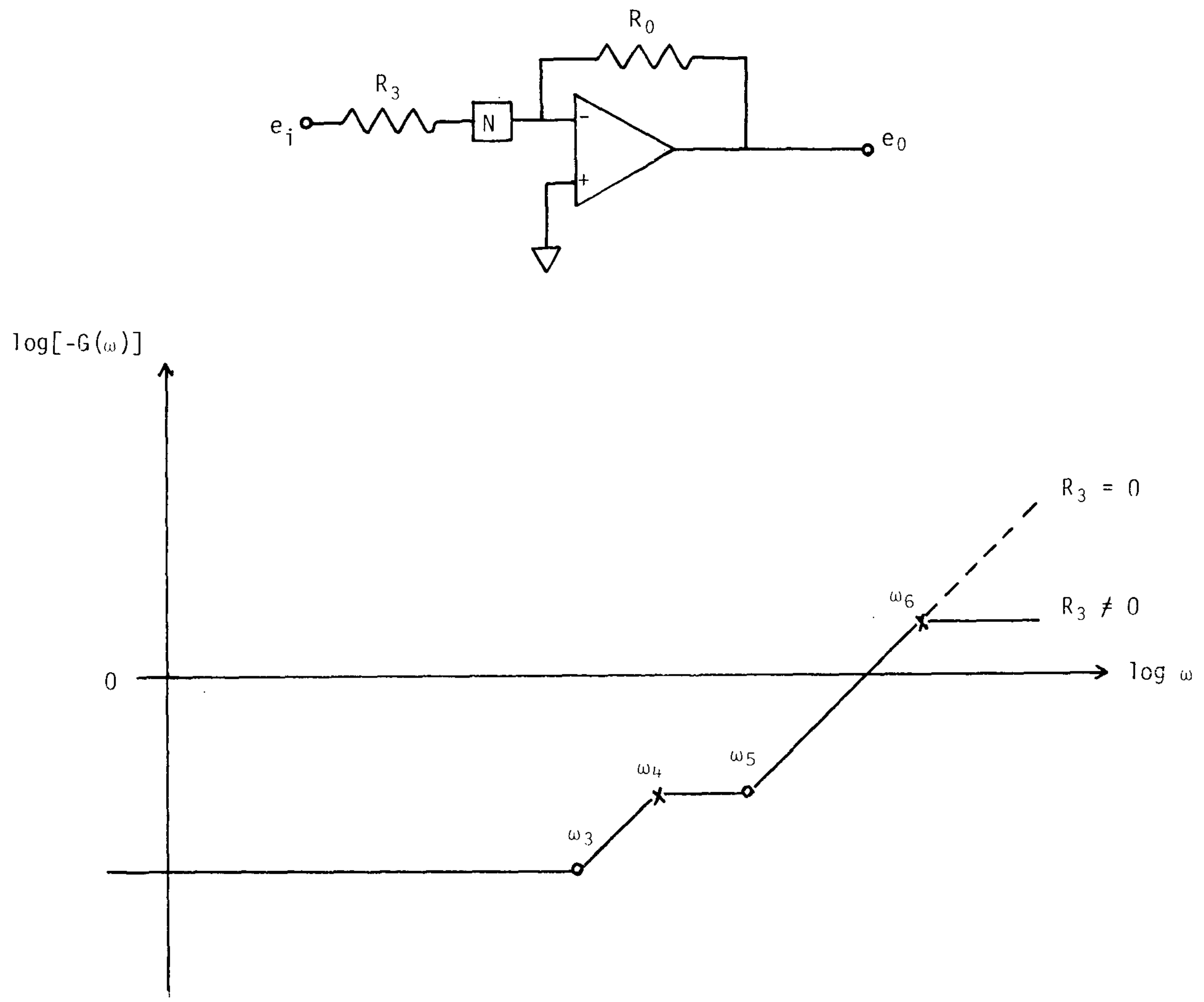
Similarly, the passive preemphasis circuit of Fig. 5 now follows easily from the case of Fig. 3(a), since its gain formula is just the reciprocal of Eq. (4). Thus its design equations also follow from Tables 1 and 2 without C_0 by making the substitutions

$$R_3 \to (R_0 + R_3) \qquad \text{and} \qquad G(\omega) \to -\frac{1}{G(\omega)}\cdot \tag{6}$$

Again, the design notes appended to Figs. 3(a), 4, and 5 should now begin to fall into place. In particular, note that both R_0 and R_3 (if nonzero) can be chosen independently in the design process. In view of the manner in which the time constants are affected by changing R_0 and R_3 in the case of the circuits of Figs. 3(a) and 5, it is preferable to think of the combinations $(R_0 + R_3)$ and R_3/R_0 [or $(R_0 + R_3)/R_0$] as being the independent quantities in these cases. This is so because of the appearance of $(R_0 + R_3)$ in the formulas of Tables 1 and 2 as a result of the substitutions (5) and (6).

4. THE CASE OF FIG. 3(b)

Fig. 3(b) requires a separate treatment. The signal gain formula now reads



NOTES:
1) R_3 includes source resistance and is required for stability. At high frequency the load on the source is R_3.
2) Both R_0 and R_3 can be chosen independently. R_0 alone enables gain adjustment without affecting frequency response; R_3 alone determines ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_3.
3) Use Tables 1 and 2 with $G(\omega)$ replaced by $1/G(\omega)$.

Fig. 4. Active inverting preemphasis circuit.

$$G(s) = \frac{Z(s) + R_3}{R_0 + 1/(C_0 s)} + 1 = \frac{1 + \{Z(s) + (R_0 + R_3)\}C_0 s}{1 + R_0 C_0 s} \quad (7)$$

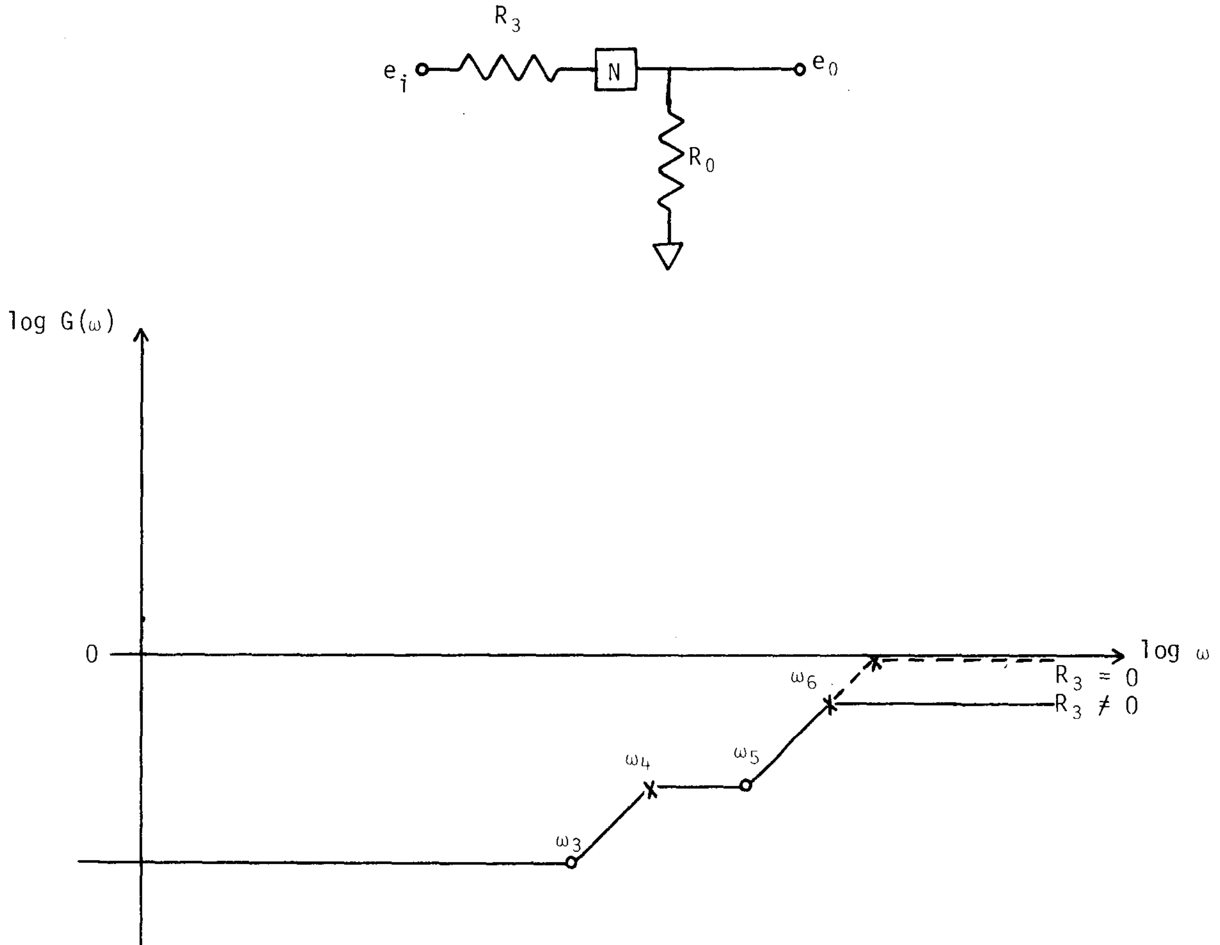
where $Z(s)$ is given in Fig. 1. In terms of the circuit time constants T_1–T_6, $G(s)$ can alternately be expressed as

$$G(s) = \frac{(1 + T_1 s)(1 + T_4 s)(1 + T_6 s)}{(1 + T_2 s)(1 + T_3 s)(1 + T_5 s)} . \quad (8)$$

As $C_0 \rightarrow \infty$, Eq. (7) reduces to Eq. (4), as expected. Once again, the poles T_2, T_3, T_5 are exactly the same as those of Fig. 2(b), and moreover, T_3 and T_5 remain unchanged whether or not C_0 is present, but the location of the zeros T_4, T_6 is different from that of both Figs. 2(b) and 3(a) as a result of the presence of C_0 in the noninverting configuration. This is in contradistinction to the inverting case, where only T_2 was affected by the presence or absence of C_0, and T_3–T_6 remained unchanged.

The analysis proceeds by equating the right-hand sides of Eqs. (7) and (8), obtaining a system of six equations which can be solved for T_1–T_6 in terms of R_0, R_1, R_2, R_3, C_0, C_1, C_2, and also for the resistor and capacitor values of the network components in terms of T_1–T_6. An example of the calculations involved in the case of the network of Fig. 1(a) is presented in Appendix 2, while the results are collected in Table 3(a)–(d) for the networks of Fig. 1(a)–(d), respectively.[8] The points made in the preceding paragraph are apparent from the formulas in the upper third of the table. In Table 4 we give the formulas for $G(\omega)$ for the circuit of Fig. 3(b). Reference should also be made to the design notes in Fig. 3(b). For this configuration only *one*

[8] Again, the symbol [·] used in some of the formulas denotes a repetition of the square-bracketed expression which precedes it within the same formula.



NOTES:
1) R_0 includes shunting effect of load resistance; R_3 includes source resistance. At high frequency the load on the source is $(R_0 + R_3)$.
2) If $R_3 \neq 0$: Both R_0 and R_3 can be chosen independently. To adjust gain without affecting frequency response, change R_3/R_0 while keeping $(R_0 + R_3)$ fixed. $(R_0 + R_3)$ alone determines ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_0, R_3.
3) If $R_3 = 0$: Only R_0 can be chosen independently. Changing R_0 affects *both* gain and ω_4, ω_6, while ω_3, ω_5 are not affected by changing R_0.
4) Use Tables 1 and 2 with R_3 replaced by $(R_0 + R_3)$ wherever it occurs, and $G(\omega)$ replaced by $-1/G(\omega)$.

Fig. 5. Passive preemphasis circuit.

of the network components can be chosen independently, and then all the others are fixed by the values of T_1–T_6. In view of the similarities with Fig. 3(a) and the appearance of R_0 and R_3 in the combination $(R_0 + R_3)$ in Eq. (7), the most logical and convenient choice for independent variable is $(R_0 + R_3)$. The formulas in the middle third of Table 3 are therefore expressed in terms of $(R_0 + R_3)$ and T_1–T_6. Since T_1 is an artifact of this circuit configuration and not of primary importance to us from the RIAA point of view, an alternative and more useful way of considering these equations is by choosing *both* $(R_0 + R_3)$ and $(R_0 + R_3)/R_0$ independently, and looking upon T_1 as a *dependent* quantity, related to the others by the formula (from Table 3):

$$\frac{R_0 + R_3}{R_0} = \frac{T_1 T_4 T_6}{T_2 T_3 T_5} \geq 1 \; . \qquad (9)$$

This formula is seen to provide a constraint on the allowable values of T_1–T_6, for we must always have the inequality satisfied. If $R_3 = 0$, it reduces to the constraint

$$T_1 T_4 T_6 = T_2 T_3 T_5 \qquad (10)$$

and R_0 remains as the only independent variable in this case. We see that a small value for T_6, which is desirable for high-frequency RIAA equalization accuracy, then necessitates a large value for T_1, which results in a long dc stabilization time for the circuit—an undesirable artifact. The time constants thus must be played off one against the other in a practical circuit.

In the next section we shall discuss design procedures using all the formulas so far developed.

Table 1(a). Design formulas for active inverting deemphasis circuits of Fig. 2, using network of Fig. 1(a).

Quantity	$R_3 = 0$ Formula	$R_3 = 0$ RIAA/IEC	$R_3 \neq 0$ Formula
T_2	R_0C_0	7950.00 μs	R_0C_0
T_3	R_1C_1	3180.00 μs	R_1C_1
T_4	$\frac{R_1R_2}{R_1+R_2}(C_1+C_2)$	318.00 μs	$\approx \frac{R_1R_2}{R_1+R_2}(C_1+C_2) + \frac{(R_1C_1-R_2C_2)^2}{(R_1+R_2)^2(C_1+C_2)}R_3$
T_5	R_2C_2	75.00 μs	R_2C_2
T_6	0		$\approx \frac{C_1C_2}{C_1+C_2}R_3$
R_1/R_3			$\frac{T_5(T_3-T_4)(T_3-T_6)}{T_4T_6(T_3-T_5)}$
R_2/R_3			$\frac{T_3(T_4-T_5)(T_5-T_6)}{T_4T_6(T_3-T_5)}$
R_0C_0	T_2	7950.00 μs	T_2
R_3C_1			$\frac{T_3T_4T_6(T_3-T_5)}{T_5(T_3-T_4)(T_3-T_6)}$
R_3C_2			$\frac{T_4T_5T_6(T_3-T_5)}{T_3(T_4-T_5)(T_5-T_6)}$
R_1C_1	T_3	3180.00 μs	T_3
R_2C_2	T_5	75.00 μs	T_5
R_1C_2	$\frac{T_5(T_3-T_4)}{T_4-T_5}$	883.33 μs	$\frac{T_5^2(T_3-T_4)(T_3-T_6)}{T_3(T_4-T_5)(T_5-T_6)}$
R_2C_1	$\frac{T_3(T_4-T_5)}{T_3-T_4}$	270.00 μs	$\frac{T_3^2(T_4-T_5)(T_5-T_6)}{T_5(T_3-T_4)(T_3-T_6)}$
R_1/R_2	$\frac{T_3-T_4}{T_4-T_5}$	11.778	$\frac{T_5(T_3-T_4)(T_3-T_6)}{T_3(T_4-T_5)(T_5-T_6)}$
C_1/C_2	$\frac{T_3(T_4-T_5)}{T_5(T_3-T_4)}$	3.600	$\frac{T_3^2(T_4-T_5)(T_5-T_6)}{T_5^2(T_3-T_4)(T_3-T_6)}$

5. HOW TO DESIGN RIAA CIRCUITS

By this stage the reader should already have a fairly good idea of the correct design procedure, making use of Tables 1–4 as appropriate. We can, however, usefully make a number of additional remarks. We shall assume that T_3–T_5 are given their RIAA values and that T_2, if present, is given either its IEC value of 7950 μs or else is suitably chosen to determine the circuit's low-frequency rolloff point. In any event it is assumed that the values of T_2–T_5 are known and fixed beforehand.

Clearly, once the circuit configuration (Figs. 2–5) has been selected, the next decision is between the four electrically equivalent networks of Fig. 1, and here a choice must be based on practical factors. Please bear in mind that we are still assuming adequate loop gain at all relevant frequencies to ensure adherence to the frequency response curve dictated by the feedback network. We shall show in Section 7 how to deal with cases in which this assumption is not valid. As is evident from the first column of Table 1, the R_1/R_2 and C_1/C_2 ratios are different for each of the four networks. Since, in practice, the range of available capacitor values is more restricted than that of resistor values, a reasonable first question to ask is which networks have a capacitor ratio that is available from, say, the standard E24 series of capacitors.[9] Now as the formulas in the second column of Table 1 show, the capacitor and resistor ratios change from their "ideal" values, given in the first column, as R_3 [or $(R_0 + R_3)$ in the case of Figs. 3 and 5] increases in value from zero. So our question must be in two parts:

1) In the ideal case $R_3 = 0$, which networks realize available E24 capacitor ratios?

2) In the case $R_3 \neq 0$ (or $R_0 + R_3 \neq 0$ for Figs. 3, 5), which networks realize available E24 capacitor ratios *and* give T_6 sufficiently small that their high-frequency zeros lie well above the audio band?

A bit of calculating using a table of E24 values and Table 1 leads to Table 5 and an answer to our questions:

1) In the ideal case only the networks of Fig. 1(a) and (d) are achievable using standard E24 capacitor values. The only three possible "ideal" designs calculated from the first column of Table 1 are given in Table 5(a), with closest E96 resistor values in parentheses.[9] Of course, if one is willing to parallel capacitors to form C_1 and C_2, an infinity of designs is possible.

[9] The E24 series of component values comprises 24 values spanning each decade, the ratio of each value to its predecessor being $10^{1/24}$. The E96 series, in the ratio $10^{1/96}$, contains 96 component values in each decade.

Table 1(b). Design formulas for active inverting deemphasis circuits of Fig. 2, using network of Fig. 1(b).

	$R_3 = 0$		$R_3 \neq 0$
Quantity	Formula	RIAA/IEC	Formula
T_2	R_0C_0	7950.00 μs	R_0C_0
T_3	$\frac{1}{2}\{[R_1C_1 + R_2(C_1 + C_2)] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$	3180.00 μs	$\frac{1}{2}\{[R_1C_1 + R_2(C_1 + C_2)] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_4	$R_2(C_1 + C_2)$	318.00 μs	$\approx R_2(C_1 + C_2) + \frac{C_1^2}{C_1 + C_2} R_3$
T_5	$\frac{1}{2}\{[R_1C_1 + R_2(C_1 + C_2)] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$	75.00 μs	$\frac{1}{2}\{[R_1C_1 + R_2(C_1 + C_2)] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_6	0		$\approx \frac{C_1C_2}{C_1 + C_2} R_3$
R_1/R_3			$\frac{T_3T_5}{T_4T_6} - 1$
R_2/R_3			$\frac{R_2}{R_1} \cdot \frac{R_1}{R_3}$
R_0C_0	T_2	7950.00 μs	T_2
R_3C_1			$R_1C_1 \cdot \frac{R_3}{R_1}$
R_3C_2			$R_1C_2 \cdot \frac{R_3}{R_1}$
R_1C_1	$T_3 - T_4 + T_5$	2937.00 μs	$\frac{T_3T_5}{R_2C_2}$
R_2C_2	$\frac{T_3T_5}{T_3 - T_4 + T_5}$	81.21 μs	$\frac{T_3T_5 - T_4T_6}{T_3 - T_4 + T_5 - T_6}$
R_1C_2	$\frac{T_3T_5(T_3 - T_4 + T_5)}{(T_3 - T_4)(T_4 - T_5)}$	1007.20 μs	$\frac{T_3T_5}{R_2C_1}$
R_2C_1	$\frac{(T_3 - T_4)(T_4 - T_5)}{T_3 - T_4 + T_5}$	236.79 μs	$T_3 + T_5 - R_1C_1 - R_2C_2$
R_1/R_2	$\frac{(T_3 - T_4 + T_5)^2}{(T_3 - T_4)(T_4 - T_5)}$	12.403	$\frac{R_1C_2}{R_2C_2}$
C_1/C_2	$\frac{(T_3 - T_4)(T_4 - T_5)}{T_3T_5}$	2.916	$\frac{R_2C_1}{R_2C_2}$

2) As R_3 (or $R_0 + R_3$) increases from zero, the C_1/C_2 ratio decreases from its value given in the first column of Table 1, and simultaneously T_6 increases in value from zero. In seeking the best designs possible in this case, which represents the most frequent situation, it is best to proceed backwards. Starting with the formula for C_1/C_2 from the second column of Table 1, we solve it for T_6 in terms of T_3–T_5 and C_1/C_2. Then from the E24 series we choose capacitor values yielding a C_1/C_2 ratio just less than that given in the first column, and calculate the corresponding value of T_6 from our formula. This value of T_6 is then used in the second column to calculate all other component values. In this way we construct the designs given in Table 5(b), listed in the order of decreasing f_6. These are believed to represent the best such designs possible. Again, many more are possible if we are willing to parallel capacitors. Note that all four networks N are represented in Table 5(b). This table can be used to construct very accurate and cheap designs, using few components, for the circuits of Figs. 2, 3(a), 4, and 5, and to a high degree of accuracy, also Fig. 3(b).

Overall it would appear that the network of Fig. 1(a) is perhaps not undeservedly the most popular of the four. An interesting question which springs to mind is whether any one of the networks offers an advantage over the others as regards the ease with which it can be "trimmed" for accuracy. To begin with, trimming is a difficult procedure, for each component affects at least two of the finally realized time constants of the network. Furthermore, to be able to trim accurately one must have either a precision RIAA circuit for reference or else be able to measure over a dynamic range of >40 dB and over a frequency range of >3 decades to an accuracy of tenths of a decibel. This is not an easy task. In fact, it is sufficiently difficult that the writer would suggest that a much better and easier procedure in practice is to produce an accurate design in the first place, and not rely on trimming to adjust the circuit for accuracy. This is, in fact, the whole thesis of this paper. This said, it is interesting to examine Table 6, a table of relative sensitivities of the main network RIAA

Table 1(c). Design formulas for active inverting deemphasis circuits of Fig. 2 using network of Fig. 1(c).

Quantity	$R_3 = 0$ Formula	$R_3 = 0$ RIAA/IEC	$R_3 \neq 0$ Formula
T_2	R_0C_0	7950.00 μs	R_0C_0
T_3	$\frac{1}{2}\{[R_1(C_1+C_2)+R_2C_1] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$	3180.00 μs	$\frac{1}{2}\{[R_1(C_1+C_2)+R_2C_1] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_4	R_2C_1	318.00 μs	$\approx R_2C_1 + C_1R_3$
T_5	$\frac{1}{2}\{[R_1(C_1+C_2)+R_2C_1] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$	75.00 μs	$\frac{1}{2}\{[R_1(C_1+C_2)+R_2C_1] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_6	0		$\approx C_2R_3$
R_1/R_3			$\frac{T_3T_5}{T_4T_6} - 1$
R_2/R_3			$\frac{R_2}{R_1} \cdot \frac{R_1}{R_3}$
R_0C_0	T_2	7950.00 μs	T_2
R_3C_1			$R_2C_1 \cdot \frac{R_3}{R_2}$
R_3C_2			$R_1C_2 \cdot \frac{R_3}{R_1}$
R_1C_1	$\frac{(T_3-T_4)(T_4-T_5)}{T_4}$	2187.00 μs	$T_3 + T_5 - R_1C_2 - R_2C_1$
R_2C_2	$\frac{T_3T_4T_5}{(T_3-T_4)(T_4-T_5)}$	109.05 μs	$\frac{T_3T_5}{R_1C_1}$
R_1C_2	$\frac{T_3T_5}{T_4}$	750.00 μs	$\frac{T_3T_5}{R_2C_1}$
R_2C_1	T_4	318.00 μs	$\frac{T_3T_4(T_5-T_6)+T_5T_6(T_3-T_4)}{T_3T_5-T_4T_6}$
R_1/R_2	$\frac{(T_3-T_4)(T_4-T_5)}{T_4^2}$	6.877	$\frac{R_1C_1}{R_2C_1}$
C_1/C_2	$\frac{(T_3-T_4)(T_4-T_5)}{T_3T_5}$	2.916	$\frac{R_2C_1}{R_2C_2}$

time constants (T_3–T_5) to changes in the values of the components R_1, R_2, C_1, and C_2. They are calculated from the formula

$$S_x^{T_i} = \frac{x}{T_i} \cdot \frac{\partial T_i}{\partial x}, \qquad i = 3,4,5$$

where x is one of R_1, R_2, C_1, or C_2, and represent the percentage change in T_i caused by a 1% change in the component x from its ideal value given in the first column of Table 1. Table 6 must be interpreted with care, but it does show that the network of Fig. 1(c) is the best, and that of Fig. 1(b) the worst, from the interaction (and hence also from the trimming) point of view. A suitable trimming procedure for the Fig. 1(a) network would be to fix R_1, say, and first adjust C_1 at 100 Hz to trim T_3; then adjust R_2 at 1 kHz to trim T_4; and finally adjust C_2 at 10 kHz to trim T_5. Of course, the procedure must be iterated, and is made more complicated by the effect each component change has on the overall gain, as is evident from Tables 2 and 4.

The next point to make is that, for all deemphasis circuits with $T_6 \neq 0$ (that is, Fig. 2 with $R_3 \neq 0$ and all cases of Fig. 3), the high-frequency zero thus introduced can be *exactly* canceled by adding an identical high-frequency pole at the output of the circuit. A passive R/C low-pass filter of time constant T_6 will do this, and if T_6 is small enough, will not significantly degrade output impedance. For example, the second and third designs given in Table 5(b), with $T_6 = 0.4$ μs, can be corrected with a filter having $R = 1.1$ kΩ and $C = 360$ pF. Such a filter *should* be incorporated, especially in those designs where f_6 is rather close to the audio band. Failure to do so will then lead to a rising response (relative to RIAA) in the top octave of the audio band.

This brings us to the next point. In a practical circuit T_6 usually cannot be made arbitrarily small, for decreasing T_6 is equivalent to decreasing R_3 for the circuits of Figs. 2 and 4 or ($R_0 + R_3$) for the circuits of Figs. 3 and 5. Practical questions of amplifier loading and stabilization will generally prevent us from decreasing these components too far, although noise considerations *per se* would dictate using the smallest possible values. In particular, R_3 may be *required* in order to ensure amplifier closed-loop stability without excessive reduction in the gain–bandwidth product and slewing rate. Also, T_1 should be made as small as possible in Fig. 3(b), for it determines the length of time the circuit will take to stabilize its dc operating levels. However, T_1 and T_6 are interrelated according to Eqs. (9),

Table 1(d). Design formulas for active inverting deemphasis circuits of Fig. 2, using network of Fig. 1(d).

	$R_3 = 0$		$R_3 \neq 0$
Quantity	Formula	RIAA/IEC	Formula
T_2	R_0C_0	7950.00 μs	R_0C_0
T_3	$\frac{1}{2}\{[R_1(C_1 + C_2) + R_2C_2] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$	3180.00 μs	$\frac{1}{2}\{[R_1(C_1 + C_2) + R_2C_2] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_4	$\frac{R_1R_2}{R_1 + R_2} C_1$	318.00 μs	$\approx \frac{R_1R_2}{R_1 + R_2} C_1 + \frac{R_1^2C_1}{(R_1 + R_2)^2} R_3$
T_5	$\frac{1}{2}\{[R_1(C_1 + C_2) + R_2C_2] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$	75.00 μs	$\frac{1}{2}\{[R_1(C_1 + C_2) + R_2C_2] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_6	0		$\approx C_2R_3$
R_1/R_3			$\frac{R_1}{R_2} \cdot \frac{R_2}{R_3}$
R_2/R_3			$\frac{R_2C_2}{R_3C_2}$
R_0C_0	T_2	7950.00 μs	T_2
R_3C_1			$R_2C_1 \cdot \frac{R_3}{R_2}$
R_3C_2			$\frac{T_3T_4T_5T_6}{T_3T_4(T_5 - T_6) + T_5T_6(T_3 - T_4)}$
R_1C_1	$\frac{(T_3 - T_4)(T_4 - T_5) + T_4^2}{T_4}$	2505.00 μs	$T_3 + T_5 - \frac{(T_3T_5 - T_4T_6)}{T_4T_6} R_3C_2$
R_2C_2	$\frac{T_3T_4T_5}{(T_3 - T_4)(T_4 - T_5) + T_4^2}$	95.21 μs	$\frac{T_3T_5}{R_1C_1}$
R_1C_2	$\frac{T_3T_5(T_3 - T_4)(T_4 - T_5)}{T_4[(T_3 - T_4)(T_4 - T_5) + T_4^2]}$	654.79 μs	$T_3 + T_5 - R_1C_1 - R_2C_2$
R_2C_1	$\frac{T_4[(T_3 - T_4)(T_4 - T_5) - T_4^2]}{(T_3 - T_4)(T_4 - T_5)}$	364.24 μs	$\frac{T_3T_5}{R_1C_2}$
R_1/R_2	$\frac{(T_3 - T_4)(T_4 - T_5)}{T_4^2}$	6.877	$\frac{R_1C_1}{R_2C_1}$
C_1/C_2	$\frac{[(T_3 - T_4)(T_4 - T_5) + T_4^2]^2}{T_3T_5(T_3 - T_4)(T_4 - T_5)}$	3.826	$\frac{R_1C_1}{R_1C_2}$

(10), and the gain formula, and so decreasing T_1 results in an increase in T_6 and a change in gain. As an example, if T_2 is chosen to be 7950 μs for an IEC design, we find from Eq. (9) that

$$T_1T_6 \geq 5.96 \times 10^{-6} \, [\mathrm{s}^2]$$

and so for $T_6 = 0.4$ μs we would have $T_1 \geq 14.9$ s. For a non-IEC design using Fig. 3(b), T_2 would be larger and so T_1 would be even greater for the same gain. In general, although T_2–T_5 are specified by the RIAA/IEC, T_1 and T_6 and the gain are at our disposal. Since the error caused by T_6 can be exactly compensated for, in a practical circuit we may have to increase T_6 in order to obtain an acceptably small value for T_1 and a suitable gain. Reference to the frequency response curve of Fig. 3(b) shows that T_1 is affected by both the circuit's gain and the location of T_2. If T_2 is specified beforehand, changing T_1 necessitates a change in gain.

A final important practical consideration is the circuit's 1-kHz gain. This can be calculated using Tables 2 or 4 as appropriate. In fact, it may be useful in the course of design to work backwards from these tables, starting with a given desired 1-kHz gain together with Eqs. (9) and (10) and calculating the corresponding values of T_1 and/or T_6 to realize this gain before proceeding to use these values in Tables 1 and 3. Speaking about gain, the design notes in Figs. 2–5 give important information concerning gain adjustment in these circuits. Referring to the upper third of Tables 1 and 3 it is seen that, when changing gain in the circuits of Figs. 2 and 4, R_3 should be held fixed and only R_0 varied, while for the circuits of Figs. 3 and 5, $(R_0 + R_3)$ should be held fixed as R_3/R_0 is varied (that is, the tapping point along $R_0 + R_3$ is varied). This procedure will ensure that the *only* frequency response casualty will be T_2. Any other procedure will affect the important RIAA time constant T_4. This point is of considerable significance, and it appears to be generally ignored in practice.

The only major design problem which can yet affect our considerations above is the lack of suitable loop gain to guarantee adherence to these formulas. We address this problem in Section 7, but first an example.

6. AN EXAMPLE

For the purposes of illustration let us consider the most

Table 2. Gain formulas for active inverting deemphasis circuits of Fig. 2.

Quantity	$R_3 = 0$, Without C_0	$R_3 = 0$, With C_0
$G(0)$	$-\frac{R_A}{R_0}$	0
$G(\infty)$	0	0
$G(\omega)$	$-\frac{R_A}{R_0}\sqrt{\frac{1+T_4^2\omega^2}{(1+T_3^2\omega^2)(1+T_5^2\omega^2)}}$	$-\frac{R_A}{R_0}\sqrt{\frac{T_2^2\omega^2(1+T_4^2\omega^2)}{(1+T_2^2\omega^2)(1+T_3^2\omega^2)(1+T_5^2\omega^2)}}$

Quantity	$R_3 \neq 0$, Without C_0	$R_3 \neq 0$, With C_0
$G(0)$	$-\frac{R_A+R_3}{R_0} \equiv -\frac{R_3}{R_0}\cdot\frac{T_3T_5}{T_4T_6}$	0
$G(\infty)$	$-\frac{R_3}{R_0}$	$-\frac{R_3}{R_0}$
$G(\omega)$	$-\frac{R_A+R_3}{R_0}\sqrt{\frac{(1+T_4^2\omega^2)(1+T_6^2\omega^2)}{(1+T_3^2\omega^2)(1+T_5^2\omega^2)}}$	$-\frac{R_A+R_3}{R_0}\sqrt{\frac{T_2^2\omega^2(1+T_4^2\omega^2)(1+T_6^2\omega^2)}{(1+T_2^2\omega^2)(1+T_3^2\omega^2)(1+T_5^2\omega^2)}}$
	$\equiv -\frac{R_3}{R_0}\cdot\frac{T_3T_5}{T_4T_6}\sqrt{\frac{(1+T_4^2\omega^2)(1+T_6^2\omega^2)}{(1+T_3^2\omega^2)(1+T_5^2\omega^2)}}$	$\equiv -\frac{R_3}{R_0}\cdot\frac{T_3T_5}{T_4T_6}\sqrt{\frac{T_2^2\omega^2(1+T_4^2\omega^2)(1+T_6^2\omega^2)}{(1+T_2^2\omega^2)(1+T_3^2\omega^2)(1+T_5^2\omega^2)}}$

difficult design case, namely, the circuit of Fig. 3(b), which also represents a sizable proportion of higher priced commercial circuits. Let us set as our criteria a 1-kHz gain of around 35 dB and a frequency response as dictated by RIAA/IEC, that is,

$$T_2 = 7950\ \mu s, \quad T_3 = 3180\ \mu s, \quad T_4 = 318\ \mu s, \quad T_5 = 75\ \mu s.$$

Reference to a straight-line approximation to the RIAA/IEC frequency response curve defined in [34] shows us that its idealized gain at 20 Hz (corresponding to T_2) is +19.9 dB relative to that at 1 kHz. Hence the desired idealized signal gain at 20 Hz is 54.9 dB, and with reference to Fig. 3(b) we conclude that $f_1 = 0.0360$ Hz, giving $T_1 = 4.419$ s. Then Eq. (9) shows that $T_6 \geq 1.349\ \mu s$, with equality if and only if $R_3 = 0$. The case $R_3 = 0$ corresponds to a high-frequency zero at 118.0 kHz, which is reasonably placed two octaves above the audio band. Let us choose the network of Fig. 1(a) and set $R_0 = 1$ kΩ. Then the following component values are easily calculated from the middle third of Table 3(a) for each of two possible designs (assuming sufficient loop gain):

1) $R_3 = 0$; $f_6 = 118.0$ kHz, $R_0 = 1$ kΩ [Fig. 1(a)], and

$$\begin{aligned} R_1 &= 511.813\ \text{k}\Omega, & R_2 &= 42.722\ \text{k}\Omega \\ C_1 &= 6.213\ \text{nF}, & C_2 &= 1.756\ \text{nF} \\ & & C_0 &= 7.950\ \mu\text{F} \end{aligned}$$

with high-frequency zero correction filter of 1 kΩ and 1.349 nF. The 1-kHz gain follows from Table 4 as 35.0 dB, as desired.

2) $R_3 = 1$ kΩ, say, to help stabilize the amplifier by increasing the high-frequency noise gain to 2;[10] $f_6 = 59.0$ kHz, $R_0 = 1$ kΩ [Fig. 1(a)], and

$$\begin{aligned} R_1 &= 511.596\ \text{k}\Omega, & R_2 &= 41.940\ \text{k}\Omega \\ C_1 &= 6.216\ \text{nF}, & C_2 &= 1.788\ \text{nF}, \\ & & C_0 &= 7.950\ \mu\text{F} \end{aligned}$$

with high-frequency correction filter of 2 kΩ and 1.349 nF. Again, Table 4 confirms the 1-kHz gain as 35.0 dB.

To give some idea of the accuracy achievable through the use of the formulas, we give in Table 7[11] the measured RIAA frequency response error of a circuit of the Fig. 2(a) type, using the theoretically calculated component values.

It is of interest to note the small changes in the values of R_1, R_2, C_1, and C_2 between these two designs. If it is desired to experiment in order to bring some of the component values closer to standard available values, one can try changing ($R_0 + R_3$), T_6, the 1-kHz gain, and/or the network to Fig. 1(b)–(d). In this way the design can be optimized.

7. TAKING INADEQUATE LOOP GAIN INTO ACCOUNT

In this last section we shall consider what can be done if the amplifier in one of our circuits does not have enough loop gain at some frequencies to ensure adequate (say,

[10] The noise gain of an amplifier, being the gain experienced by input-referred noise and error, is the reciprocal of the feedback loop attenuation β. It is equal to the signal gain for an ideal noninverting configuration, and is important, for together with the open-loop gain, it determines a feedback amplifier's stability.

[11] Table 7 is extracted from [26, Table IV] and reproduced with the kind permission of the publisher (The Audio Amateur, P.O. Box 176, Peterborough, NH 03458).

Table 3(a). Design formulas for active noninverting deemphasis circuit of Fig. 3(b), using network of Fig. 1(a).

Quantity	Formula
T_1, T_4, T_6	With $T_1 > T_4 > T_6$, they are $-\left(\frac{1}{\text{roots}}\right)$ of the cubic in s: $0 = 1 + [(R_0 + R_3)C_0 + (R_1 + R_2)C_0 + R_1C_1 + R_2C_2]s + [(R_0 + R_3)C_0(R_1C_1 + R_2C_2) + R_1R_2\{C_0(C_1 + C_2) + C_1C_2\}]s^2 + [(R_0 + R_3)C_0R_1C_1R_2C_2]s^3$
T_2	R_0C_0
T_3	R_1C_1
T_5	R_2C_2
$R_0/(R_0 + R_3)$	$\frac{T_2T_3T_5}{T_1T_4T_6} \rightarrow$ constraint $T_1T_4T_6 = T_2T_3T_5$ if $R_3 = 0$
$R_1/(R_0 + R_3)$	$\frac{T_5(T_1 - T_3)(T_3 - T_4)(T_3 - T_6)}{T_1T_4T_6(T_3 - T_5)}$
$R_2/(R_0 + R_3)$	$\frac{T_3(T_1 - T_5)(T_4 - T_5)(T_5 - T_6)}{T_1T_4T_6(T_3 - T_5)}$
R_0C_0	T_2
$(R_0 + R_3)C_1$	$\frac{T_1T_3T_4T_6(T_3 - T_5)}{T_5(T_1 - T_3)(T_3 - T_4)(T_3 - T_6)}$
$(R_0 + R_3)C_2$	$\frac{T_1T_4T_5T_6(T_3 - T_5)}{T_3(T_1 - T_5)(T_4 - T_5)(T_5 - T_6)}$
R_1C_1	T_3
R_2C_2	T_5
R_1C_2	$\frac{T_5^2(T_1 - T_3)(T_3 - T_4)(T_3 - T_6)}{T_3(T_1 - T_5)(T_4 - T_5)(T_5 - T_6)}$
R_2C_1	$\frac{T_3^2(T_1 - T_5)(T_4 - T_5)(T_5 - T_6)}{T_5(T_1 - T_3)(T_3 - T_4)(T_3 - T_6)}$
R_1/R_2	$\frac{T_5(T_1 - T_3)(T_3 - T_4)(T_3 - T_6)}{T_3(T_1 - T_5)(T_4 - T_5)(T_5 - T_6)}$
C_1/C_2	$\frac{T_3^2(T_1 - T_5)(T_4 - T_5)(T_5 - T_6)}{T_5^2(T_1 - T_3)(T_3 - T_4)(T_3 - T_6)}$

0.2-dB) adherence of the signal gain to that dictated by the feedback network.[12] This will be the case if the loop gain is less than 30–40 dB at any frequency at which the open-loop and noise gain curves are parallel (no relative phase shift), or less than 15–20 dB at any freqeuncy at which the open-loop and noise gain curves have a relative slope of 6 dB per octave (90° relative phase shift).[13] This occurs frequently in practice in disk preamplifiers, and two particular situations are common:

1) The discrete amplifier with large open-loop bandwidth but inadequate low-frequency open-loop gain. Here the main RIAA errors are in the region of the ω_2 and ω_3 poles.

2) The integrated operational amplifier with large low-frequency open-loop gain but small open-loop bandwidth, resulting in inadequate high-frequency loop gain. Here the main errors are around the pole at ω_5.

These errors take the form of deviations in both gain and pole position. These two situations are illustrated diagrammatically in Fig. 6 for the circuit of Fig. 3(b). The solid lines represent the originally intended RIAA response, and the dashed lines show the response actually realized. One solution is, of course, to reduce the desired 1-kHz signal gain in order to increase the available loop gain. If this is not practicable, other solutions must be sought, and these are the topic of the present discussion. Since our aim is mainly to illustrate a design procedure by means of which these errors can be avoided, we shall restrict the discussion to the circuit of Fig. 3(b). It can, however, be applied to the other circuits, but with somewhat greater difficulty.

In the diagrams of Fig. 6 the unprimed quantities are the ones used in the formulas developed earlier for the case of infinite open-loop gain, realizing closed-loop gain G, while the primed quantities are those actually realized due to the finite open-loop gain. Our aim is to force the primed ω_i to take on the *desired* RIAA values by *deliberately* choosing the unprimed ω_i differently. Then the shape of the achieved (dashed) curves will be correct, although their gain will be somewhat below that predicted by Table 4. This is indeed possible.

[12] The writer would like to thank John Vanderkooy for bringing home to him the importance of a discussion of this topic, and for suggesting a possible analytical approach.

[13] See footnote 10.

Table 3(b). Design formulas for active noninverting deemphasis circuit of Fig. 3(b), using network of Fig. 1(b).

Quantity	Formula
T_1, T_4, T_6	With $T_1 > T_4 > T_6$, they are $-\ (\frac{1}{\text{roots}})$ of the cubic in s: $0 = 1 + [(R_0 + R_3)C_0 + R_1C_0 + R_1C_1 + R_2(C_1 + C_2)]s + [(R_0 + R_3)C_0\{R_1C_1 + R_2(C_1 + C_2)\} + R_1R_2\{C_0(C_1 + C_2) + C_1C_2\}]s^2 + [(R_0 + R_3)C_0R_1C_1R_2C_2]s^3$
T_2	R_0C_0
T_3	$\frac{1}{2}\{[R_1C_1 + R_2(C_1 + C_2)] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_5	$\frac{1}{2}\{[R_1C_1 + R_2(C_1 + C_2)] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
$R_0/(R_0 + R_3)$	$\frac{T_2T_3T_5}{T_1T_4T_6} \rightarrow$ constraint $T_1T_4T_6 = T_2T_3T_5$ if $R_3 = 0$
$R_1/(R_0 + R_3)$	$\frac{T_3T_5(T_1 - T_3 + T_4 - T_5 + T_6)}{T_1T_4T_6} - 1$
$R_2/(R_0 + R_3)$	$\frac{R_2}{R_1} \cdot \frac{R_1}{R_0 + R_3}$
R_0C_0	T_2
$(R_0 + R_3)C_1$	$R_1C_1 \cdot \frac{R_0 + R_3}{R_1}$
$(R_0 + R_3)C_2$	$R_1C_2 \cdot \frac{R_0 + R_3}{R_1}$
R_1C_1	$\frac{T_3T_5}{R_2C_2}$
R_2C_2	$\frac{T_3T_5(T_1 - T_3 + T_4 - T_5 + T_6) - T_1T_4T_6}{(T_3 + T_5)(T_1 - T_3 + T_4 - T_5 + T_6) - (T_1T_4 + T_1T_6 + T_4T_6 - T_3T_5)}$
R_1C_2	$\frac{T_3T_5}{R_2C_1}$
R_2C_1	$T_3 + T_5 - R_1C_1 - R_2C_2$
R_1/R_2	$\frac{R_1C_2}{R_2C_2}$
C_1/C_2	$\frac{R_2C_1}{R_2C_2}$

Let

$$G(s) = \frac{N(s)}{D(s)} \quad \text{and} \quad G'(s) = k\frac{N'(s)}{D'(s)} \qquad (11)$$

where k is a constant, and $N(s)$, $N'(s)$ and $D(s)$, $D'(s)$ are the polynomials in s in the numerators and denominators of the gain formulas $G(s)$ and $G'(s)$, respectively, from Eq. (8). If $A_v(s)$ denotes the open-loop gain, the familiar gain formula

$$G'(s) = \frac{A_v(s)}{1 + A_v(s)/G(s)} = \frac{G(s)}{1 + G(s)/A_v(s)}$$

for a noninverting amplifier yields

$$k\frac{N'(s)}{D'(s)} = \frac{N(s)}{D(s) + N(s)/A_v(s)} \qquad (12)$$

as the relation between the N and the D. We now specialize Eq. (12) to the two cases of Fig. 6.

7.1 Constant Open-Loop Gain: $A_v = A_{v0}$

This is a good approximation to a wide open-loop bandwidth amplifier. Then we deduce from Eq. (12) that

$$\left.\begin{aligned} k &= \frac{1}{1 + 1/A_{v0}}, \qquad N(s) = N'(s), \\ D(s) &= D'(s) - \frac{N'(s) - D'(s)}{A_{v0}}. \end{aligned}\right\} \qquad (13)$$

The first important point to note is the formula for k, relating the 0-Hz gains $G(0)$ and $G'(0)$. Clearly, as $A_{v0} \rightarrow \infty$, these become equal as expected. The second point is the somewhat surprising fact that the zeros ω_1', ω_4', and ω_6' are *not shifted* in frequency by the finite loop gain error. It is only the poles ω_2', ω_3', and ω_5' which are shifted according to the last of Eqs. (13). Again, as $A_{v0} \rightarrow \infty$, they tend to their expected values. Our next step is to substitute the forms of N and D from Eq. (8) into the last of Eqs. (13) and equate coefficients of like powers of s on both sides. If we introduce the notation

$$\left.\begin{aligned} N_1' &= T_1' + T_4' + T_6', \\ N_2' &= T_1'T_4' + T_1'T_6' + T_4'T_6', \\ N_3' &= T_1'T_4'T_6' \\ \\ D_1' &= T_2' + T_3' + T_5', \\ D_2' &= T_2'T_3' + T_2'T_5' + T_3'T_5', \\ D_3' &= T_2'T_3'T_5' \end{aligned}\right\} \qquad (14)$$

Table 3(c). Design formulas for active noninverting deemphasis circuit of Fig. 3(b), using network of Fig. 1(c).

Quantity	Formula
T_1, T_4, T_6	With $T_1 > T_4 > T_6$, they are $-\left(\frac{1}{\text{roots}}\right)$ of the cubic in s: $0 = 1 + [(R_0+R_3)C_0 + R_1C_0 + R_1(C_1+C_2) + R_2C_1]s + [(R_0+R_3)C_0\{R_1(C_1+C_2)+R_2C_1\} + R_1C_1R_2(C_0+C_2)]s^2 + [(R_0+R_3)C_0R_1C_1R_2C_2]s^3$
T_2	R_0C_0
T_3	$\frac{1}{2}\{[R_1(C_1+C_2)+R_2C_1] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_5	$\frac{1}{2}\{[R_1(C_1+C_2)+R_2C_1] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
$R_0/(R_0+R_3)$	$\frac{T_2T_3T_5}{T_1T_4T_6} \rightarrow$ constraint $T_1T_4T_6 = T_2T_3T_5$ if $R_3 = 0$
$R_1/(R_0+R_3)$	$\frac{T_3T_5(T_1 - T_3 + T_4 - T_5 + T_6)}{T_1T_4T_6} - 1$
$R_2/(R_0+R_3)$	$\frac{R_2}{R_1} \cdot \frac{R_1}{R_0+R_3}$
R_0C_0	T_2
$(R_0+R_3)C_1$	$R_1C_1 \cdot \frac{R_0+R_3}{R_1}$
$(R_0+R_3)C_2$	$R_1C_2 \cdot \frac{R_0+R_3}{R_1}$
R_1C_1	$T_3 + T_5 - R_1C_2 - R_2C_1$
R_2C_2	$\frac{T_3T_5}{R_1C_1}$
R_1C_2	$\frac{T_3T_5}{R_2C_1}$
R_2C_1	$\frac{T_3T_5(T_1T_4 + T_1T_6 + T_4T_6 - T_3T_5) - (T_3+T_5)T_1T_4T_6}{T_3T_5(T_1 - T_3 + T_4 - T_5 + T_6) - T_1T_4T_6}$
R_1/R_2	$\frac{R_1C_1}{R_2C_1}$
C_1/C_2	$\frac{R_2C_1}{R_2C_2}$

we deduce that

$$\left.\begin{aligned} T_2 + T_3 + T_5 &= D_1' - \frac{N_1' - D_1'}{A_{v0}} \\ T_2T_3 + T_2T_5 + T_3T_5 &= D_2' - \frac{N_2' - D_2'}{A_{v0}} \\ T_2T_3T_5 &= D_3' - \frac{N_3' - D_3'}{A_{v0}} \end{aligned}\right\} \quad (15)$$

and so T_2, T_3, T_5 with $T_2 > T_3 > T_5$, are the roots of the cubic in T:

$$T^3 - \left[D_1' - \frac{N_1' - D_1'}{A_{v0}}\right]T^2 + \left[D_2' - \frac{N_2' - D_2'}{A_{v0}}\right]T - \left[D_3' - \frac{N_3' - D_3'}{A_{v0}}\right] = 0. \quad (16)$$

This equation is exact, and when we insert into its coefficients the *desired* (that is, primed) RIAA time constants, its solutions T_2, T_3, T_5 give us the time constants which, together with

$$T_1 = T_1', \qquad T_4 = T_4', \qquad T_6 = T_6' \quad (17)$$

are the ones that must be used in the design tables and formulas of earlier sections for the circuit to realize the desired frequency response with gain error k given by Eq. (13).

A reasonable approximation in this case is $T_5 = T_5'$ in view of the 20-dB greater loop gain available at ω_5. If this simplification is made in system (15), it follows that, as a good approximation, T_2 and T_3, with $T_2 > T_3$, can be obtained more simply as the roots of the quadratic in T:

$$T^2 - \left[T_2' + T_3' - \frac{N_1' - D_1'}{A_{v0}}\right]T + \left[D_3' - \frac{N_3' - D_3'}{A_{v0}}\right] \cdot \frac{1}{T_5'} = 0, \qquad T_5 = T_5'. \quad (18)$$

7.2 Integrating Open-Loop Gain: $A_v = \omega_0/s$

Here ω_0 denotes the unity-gain angular frequency of the amplifier. This is a good middle- to high-frequency approximation of an integrating operational amplifier (the common type) with small open-loop bandwidth. Then the denominator on the right-hand side of Eq. (12) is quartic in

Table 3(d). Design formulas for active noninverting deemphasis circuit of Fig. 3(b), using network of Fig. 1(d).

Quantity	Formula
T_1, T_4, T_6	With $T_1 > T_4 > T_6$, they are $-$ $(\frac{1}{\text{roots}})$ of the cubic in s: $0 = 1 + [(R_0 + R_3)C_0 + (R_1 + R_2)C_0 + R_1(C_1 + C_2) + R_2C_2]s + [(R_0 + R_3)C_0\{R_1(C_1 + C_2) + R_2C_2\} + R_1C_1R_2(C_0 + C_2)]s^2 + [(R_0 + R_3)C_0R_1C_1R_2C_2]s^3$
T_2	R_0C_0
T_3	$\frac{1}{2}\{[R_1(C_1 + C_2) + R_2C_2] + \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
T_5	$\frac{1}{2}\{[R_1(C_1 + C_2) + R_2C_2] - \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2}\}$
$R_0/(R_0 + R_3)$	$\frac{T_2T_3T_5}{T_1T_4T_6} \rightarrow$ constraint $T_1T_4T_6 = T_2T_3T_5$ if $R_3 = 0$
$R_1/(R_0 + R_3)$	$\frac{R_1}{R_2} \cdot \frac{R_2}{R_0 + R_3}$
$R_2/(R_0 + R_3)$	$\frac{R_2C_2}{(R_0 + R_3)C_2}$
R_0C_0	T_2
$(R_0 + R_3)C_1$	$R_2C_1 \cdot \frac{R_0 + R_3}{R_2}$
$(R_0 + R_3)C_2$	$\frac{T_1T_3T_4T_5T_6}{T_3T_5(T_1T_4 + T_1T_6 + T_4T_6 - T_3T_5) - (T_3 + T_5)T_1T_4T_6}$
R_1C_1	$T_3 + T_5 - \frac{T_3T_5(T_1 - T_3 + T_4 - T_5 + T_6) - T_1T_4T_6}{T_1T_4T_6} \cdot (R_0 + R_3)C_2$
R_2C_2	$\frac{T_3T_5}{R_1C_1}$
R_1C_2	$T_3 + T_5 - R_1C_1 - R_2C_2$
R_2C_1	$\frac{T_3T_5}{R_1C_2}$
R_1/R_2	$\frac{R_1C_1}{R_2C_1}$
C_1/C_2	$\frac{R_1C_1}{R_1C_2}$

s, and hence the left-hand side must also have a fourth pole at ω_7', say, as illustrated in Fig. 6(b). Thus Eq. (12) becomes

$$k \frac{N'(s)}{(1 + T_7's)D'(s)} = \frac{N(s)}{D(s) + sN(s)/\omega_0}$$

and we deduce that

$$\left.\begin{aligned} k &= 1, \quad N(s) = N'(s), \\ D(s) &= (1 + T_7's)D'(s) - \frac{sN'(s)}{\omega_0}. \end{aligned}\right\} \tag{19}$$

As in case 7.1, $G(s)$ and $G'(s)$ have the *same* zeros as expressed by Eq. (17). Now, however, $G(0)$ equals $G'(0)$, while the poles ω_2', ω_3', and ω_5' are shifted according to the last of Eqs. (19), and a further pole ω_7' is added. As $\omega_0 \rightarrow \infty$, $\omega_7' \rightarrow \infty$, and the other poles tend to their expected values.

Substituting into the last of Eqs. (19) from Eq. (8), and equating coefficients of like powers of s, we find, in the notation of Eq. (14), that T_2, T_3, and T_5, with $T_2 > T_3 > T_5$, are the roots of the cubic in T:

$$\begin{aligned} T^3 &- \left[D_1' + T_7' - \frac{1}{\omega_0}\right]T^2 \\ &+ \left[D_2' + T_7'D_1' - \frac{N_1'}{\omega_0}\right]T \\ &- \left[D_3' + T_7'D_2' - \frac{N_2'}{\omega_0}\right] = 0 \end{aligned} \tag{20}$$

where

$$T_7' = \frac{N_3'}{D_3'\omega_0} = \frac{T_1'T_4'T_6'}{T_2'T_3'T_5'\omega_0}. \tag{21}$$

Note that by the first formula in the middle third of Table 3 $\omega_7' \leq \omega_0$, with equality if and only if $R_3 = 0$. This is also evident from Fig. 6(b).

Reduction of Eq. (20) to a quadratic equation, in the way in which Eq. (18) was derived from Eq. (16), is *not* justifiable in this case, and the full cubic Eq. (20) should be used.

7.3 General Single-Pole Amplifier Gain:

$$A_v = \frac{A_{v0}}{1 + A_{v0}s/\omega_0}$$

As a generalization, we can combine the cases of Fig.

Table 4. Gain formulas for active noninverting deemphasis circuit of Fig. 3(b).

Quantity	Formula
$G(0)$	1
$G(\infty)$	$\frac{T_1T_4T_6}{T_2T_3T_5}$, which becomes 1 if $R_3 = 0$
$G(\omega)$	$\sqrt{\frac{(1+T_1^2\omega^2)(1+T_4^2\omega^2)(1+T_6^2\omega^2)}{(1+T_2^2\omega^2)(1+T_3^2\omega^2)(1+T_5^2\omega^2)}}$

Whichever case we are dealing with, once the modified values T_1–T_6 have been calculated, the appropriate resistor and capacitor values can be obtained from Table 3.[14] One final comment is warranted. In practice it would appear that a procedure frequently adopted, when it transpires that a design is not following the required RIAA curve due to inadequate loop gain, is to adjust a *single* component's value, for example, R_1 or C_1 in case 7.1 and R_2 or C_2 in case 7.2. *This is incorrect,* for such a change will, according to the formulas in the upper third of Table 3, modify not only the requisite pole T_3' or T_5', but also the zeros T_1', T_4', and T_6' which our analysis shows should be left unchanged. Our whole thesis is that, by appropriate calculation, an extremely accurate design is

[14] Of course, we assume that the shifts involved are not so large that the roots of Eqs. (16), (18), (20), and (22) become complex, for then the configurations under consideration *cannot* be made to follow the RIAA curve, and the amplifier's open-loop gain must be considered to be totally inadequate.

Table 5. Best possible RIAA network designs using E 24 series capacitors (closest E96 series resistors given in parentheses).

(a) Ideal case: $T_6 = 0$—Circuits of Figs. 2 and 4 with $R_3 = 0$.

Network of Fig. 1	C_1	C_2	$\frac{C_1}{C_2}$	R_1	R_2
(a)	2.7 nF	750 pF	3.600	1.178 MΩ (1.18 MΩ)	100.000 kΩ (100 kΩ)
(a)	3.6 nF	1.0 nF	3.600	883.333 kΩ (887 kΩ)	75.000 kΩ (75.0 kΩ)
(d)	1.8 nF	470 pF	3.830	1.392 MΩ (1.40 MΩ)	202.574 kΩ (205 kΩ)

(b) General case: $T_6 \neq 0$—Circuits of Figs. 2 and 4 with $R_3 \neq 0$, or of Figs. 3(a) and 5 with R_3 replaced by $(R_0 + R_3)$ below.

Network of Fig. 1	C_1	C_2	$\frac{C_1}{C_2}$	f_6	R_1	R_2	R_3 or $(R_0 + R_3)$
(a)	4.3 nF	1.2 nF	3.583	448 kHz	739.535 kΩ (732 kΩ)	62.500 kΩ (61.9 kΩ)	380.4 Ω (383 Ω)
(b)	1.8 nF	620 pF	2.903	398 kHz	1.632 MΩ (1.62 MΩ)	130.924 kΩ (130 kΩ)	871.0 Ω (866 Ω)
(c)	1.8 nF	620 pF	2.903	398 kHz	1.214 MΩ (1.21 MΩ)	176.018 kΩ (174 kΩ)	647.9 Ω (649 Ω)
(a)	2.0 nF	560 pF	3.571	261 kHz	1.590 MΩ (1.58 MΩ)	133.929 kΩ (133 kΩ)	1.402 kΩ (1.40 kΩ)
(d)	9.1 nF	2.4 nF	3.792	227 kHz	274.742 kΩ (274 kΩ)	39.748 kΩ (39.2 kΩ)	293.8 Ω (294 Ω)

6(a) and (b) to realize a formula for the general case of a single-pole amplifier with 0-Hz gain A_{v0} and unity-gain angular frequency ω_0 (that is, open-loop bandwidth ω_0/A_{v0}). Once again we find that the zeros are unchanged, $G(0)$ and $G'(0)$ differ by the same factor k as in case 7.1 (Eq. (13)), and T_2, T_3, and T_5, with $T_2 > T_3 > T_5$, are the roots of the cubic in T:

$$T^3 - \left[D_1' + T_7' - \frac{1}{\omega_0} - \frac{N_1' - D_1' - T_7'}{A_{v0}}\right]T^2 + \left[D_2' + T_7'D_1' - \frac{N_1'}{\omega_0} - \frac{N_2' - D_2' - T_7'D_1'}{A_{v0}}\right]T - \left[D_3' + T_7'D_2' - \frac{N_2'}{\omega_0} - \frac{N_3' - D_3' - T_7'D_2'}{A_{v0}}\right] = 0 \quad (22)$$

where now

$$T_7' = \frac{N_3'}{\left(1 + \frac{1}{A_{v0}}\right)D_3'\omega_0} = \frac{T_1'T_4'T_6'}{\left(1 + \frac{1}{A_{v0}}\right)T_2'T_3'T_5'\omega_0}. \quad (23)$$

These formulas clearly generalize those of cases 7.1 and 7.2.

Table 6. T-sensitivities to component variations for the ideal case.

Sensitivity	Fig. 1(a)	Fig. 1(b)	Fig. 1(c)	Fig. 1(d)
$S_{R_1}^{T_3}$	1.000	0.922	0.922	0.993
$S_{C_1}^{T_3}$	1.000	0.998	0.783	0.783
$S_{R_2}^{T_3}$	0.000	0.078	0.078	0.007
$S_{C_2}^{T_3}$	0.000	0.002	0.217	0.217
$S_{R_1}^{T_4}$	0.078	0.000	0.000	0.127
$S_{C_1}^{T_4}$	0.783	0.745	1.000	1.000
$S_{R_2}^{T_4}$	0.922	1.000	1.000	0.873
$S_{C_2}^{T_4}$	0.217	0.255	0.000	0.000
$S_{R_1}^{T_5}$	0.000	0.078	0.078	0.007
$S_{C_1}^{T_5}$	0.000	0.002	0.217	0.217
$S_{R_2}^{T_5}$	1.000	0.922	0.922	0.993
$S_{C_2}^{T_5}$	1.000	0.998	0.783	0.783

Table 7. Measured frequency response error of circuit of Fig. 2(a) type, using theoretically calculated component values (from [26, Table IV]).

Frequency [kHz]	Error [dB]
0.01	−0.024
0.02	0.000
0.03	+0.004
0.04	+0.007
0.05	+0.010
0.07	+0.012
0.1	+0.013
0.2	+0.013
0.3	+0.009
0.4	+0.007
0.5	+0.005
0.7	+0.001
1	0.0 (ref.)
1.5	0.000
2	0.000
3	+0.001
5	+0.003
7	+0.003
10	+0.003
15	+0.001
20	−0.002
30	−0.010
40	−0.020
50	−0.033

achievable without the need for any trimming which, as indicated earlier, is extremely difficult to carry out successfully.

8. ADDENDUM

There appears to be some current interest in the use of passive RIAA deemphasis circuits, and since the design equations for such circuits are contained in the tables already presented, it is felt to be worthwhile to provide basic design data for such circuits here as well. The four networks of Fig. 1, when used in this configuration, give rise only to the two distinct circuits shown in Fig. 7 (no low-frequency rolloff is provided). It is found that the relevant design formulas are precisely those given in column 1 of Table 1(b) and (c) for the circuits of Fig. 7(a) and (b), respectively. Note that these circuits provide ideal high-frequency deemphasis, but the preceding flat preamplifier stage does not have the greatly increased high-frequency overload margin achieved by the corresponding active deemphasis circuits.

9. ACKNOWLEDGMENT

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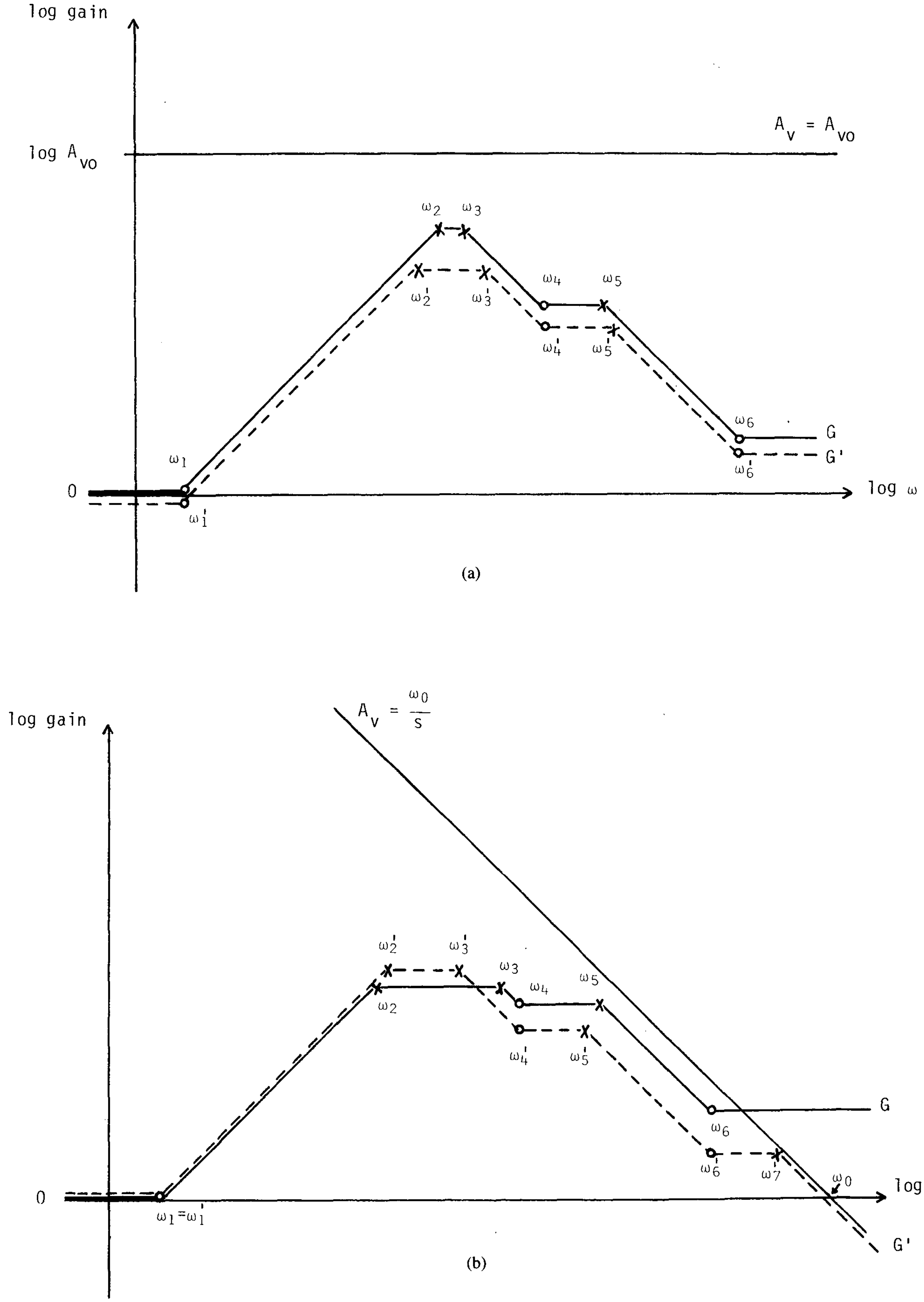

Fig. 6. Effects of inadequate loop gain. (a) Constant open-loop gain: $A_v = A_{v0}$. (b) Integrating open-loop gain: $A_v = \omega_0/s$.

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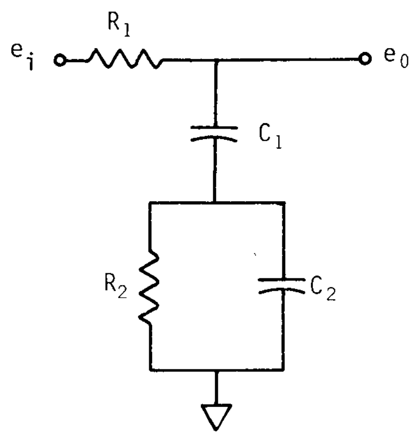
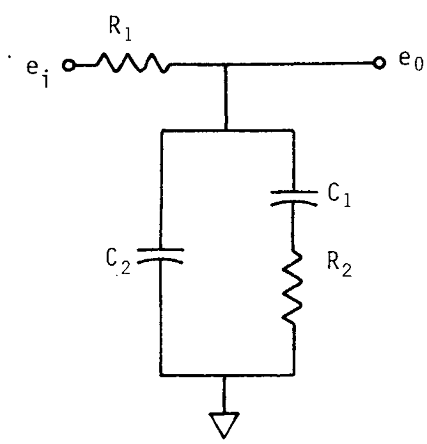
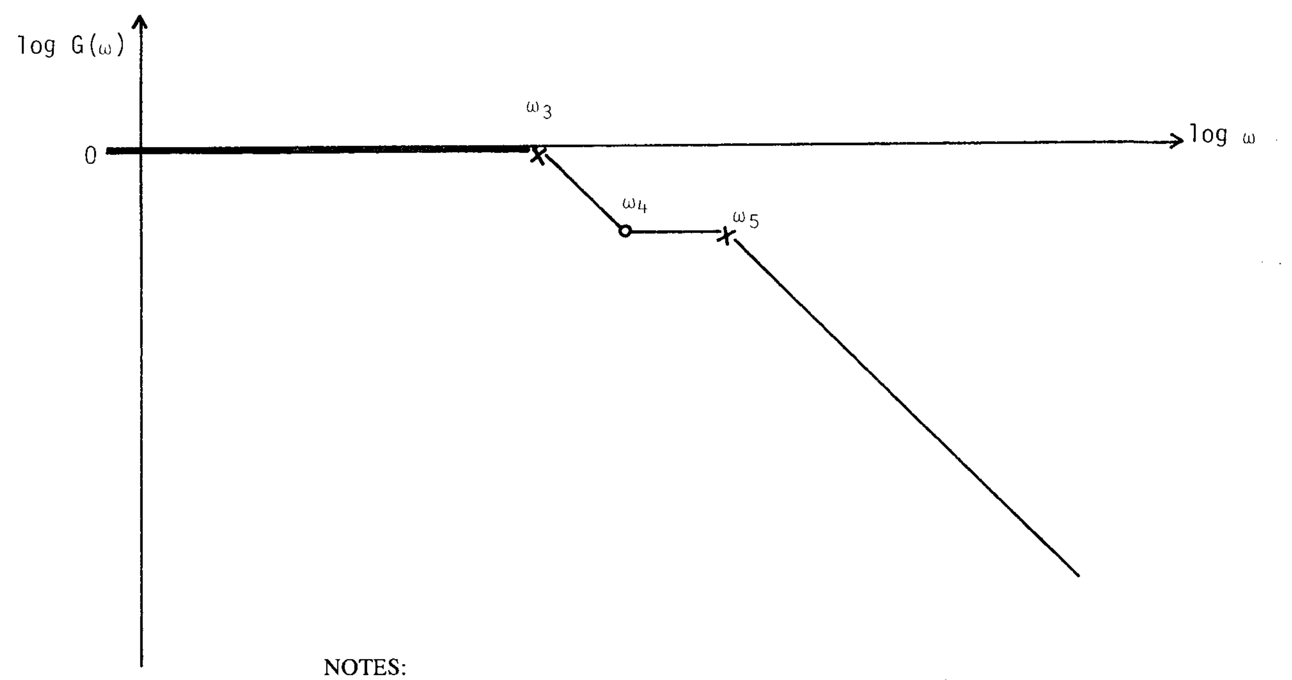


(a) Use Table 1(b) with $R_3 = 0$



(b) Use Table 1(c) with $R_3 = 0$



NOTES:
1) R_1 includes source resistance. The shunting effect of the load resistance modifies R_1, G. At high frequency the load on the source is R_1.
2) *One* of the components can be chosen independently.
3) No gain adjustment is possible; the low-frequency gain is unity.
4) $G(\omega) = \sqrt{(1 + T_4^2\omega^2)/[(1 + T_3^2\omega^2)(1 + T_5^2\omega^2)]}$.

Fig. 7. Passive deemphasis circuits.

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APPENDIX 1

An Example of the Calculations Leading to Tables 1 and 2

As an illustration of the procedure used, we shall consider Fig. 2 with the network of Fig. 1(a). Substituting into Eq. (1) for $Z(s)$, given in Fig. 1(a), and equating the right-hand sides of Eqs. (1) and (2), we obtain, after some simplification:

$$\frac{(R_A + R_3)}{R_0} \cdot \frac{R_0C_0s\left\{1 + \frac{R_1R_2(C_1+C_2) + (R_1C_1 + R_2C_2)R_3}{R_1+R_2+R_3}s + \frac{R_1C_1R_2C_2R_3}{R_1+R_2+R_3}s^2\right\}}{(1+R_0C_0s)\{1 + (R_1C_1 + R_2C_2)s + R_1C_1R_2C_2s^2\}}$$

$$= \frac{R_A + R_3}{R_0} \cdot \frac{T_2s\{1 + (T_4+T_6)s + T_4T_6s^2\}}{(1+T_2s)\{1 + (T_3+T_5)s + T_3T_5s^2\}} .$$

We now equate the coefficients of corresponding powers of s in the numerators and denominators on both sides of this equation, and so reduce it to the following system of five equations:

$$\left.\begin{aligned} T_2 &= R_0C_0 \\ T_3 + T_5 &= R_1C_1 + R_2C_2 \\ T_3T_5 &= R_1C_1R_2C_2 \\ T_4 + T_6 &= \frac{R_1R_2(C_1+C_2) + (R_1C_1 + R_2C_2)R_3}{R_1+R_2+R_3} \\ T_4T_6 &= \frac{R_1C_1R_2C_2R_3}{R_1+R_2+R_3} \end{aligned}\right\} \quad (24)$$

for the five unknowns T_2–T_6, for which it is easily solved:

$$T_2 = R_0C_0$$
$$T_3 = R_1C_1$$
$$T_5 = R_2C_2$$

and, if $R_3 = 0$,

$$T_4 = \frac{R_1R_2}{R_1+R_2}(C_1 + C_2)$$
$$T_6 = 0$$

while, if $R_3 \neq 0$,[15]

$$T_{4,6} = \frac{[R_1R_2(C_1+C_2) + (R_1C_1 + R_2C_2)R_3] \pm \sqrt{[\cdot]^2 - 4R_1C_1R_2C_2R_3(R_1+R_2+R_3)}}{2(R_1+R_2+R_3)} .$$

This latter expression for T_4 and T_6 can now be approximated by standard expansion techniques to derive expressions for T_4 and T_6 which are accurate to first order in R_3, provided $R_3 << R_2$. In this way we obtain the formulas in the upper third of Table 1(a).

From the design point of view it is, however, more useful to have expressions for the values of the resistors and capacitors R_0, R_1, R_2, R_3, C_0, C_1, C_2 in terms of the desired network time constants T_2–T_6. To this end one returns to the system (24), which viewed from this point of view is a system of five equations in seven unknowns, two of which can thus be chosen arbitrarily. We choose to specify R_0 and R_3 beforehand, for they will usually have their values circumscribed by noise and stability considerations. We thus solve system (24) for R_1, R_2, C_0, C_1, C_2 in terms of R_0 and R_3 and, after rather laborious calculations, come up with the formulas which constitute the middle third of Table 1(a). Finally these formulas are combined, eliminating R_0 and R_3, to derive the formulas given in the lower third of the table. These are useful, for they tell us the correct values to expect for the individual RC products and resistor and capacitor ratios in terms solely of T_2–T_6.

It should be remarked that, in the case of the first column of Table 1(a), $R_3 = 0$, and so $T_6 = 0$ and system (24) reduces to a system of only four equations in the six unknowns R_0, R_1, R_2, C_0, C_1, C_2. Now R_0 together with *any one* of the remaining components may be chosen arbitrarily beforehand. The formulas in the second column of Table 1(a) reduce to the "ideal" formulas in the first column as $R_3 \to 0$ (that is, $T_6 \to 0$); the latter thus remain useful as good approximations if R_3 is small. Note that these formulas all remain valid also for the case of Fig. 2(a), that is, as $C_0 \to \infty$. Changing C_0 affects only T_2, leaving T_3–T_6 unchanged. An interesting point is that T_3 and T_5, corresponding to the poles of $G(s)$, occur at *precisely* the poles of $Z(s)$ itself, even when $R_3 \neq 0$, whereas the middle RIAA time constant T_4 is *increased* in value from the zero of $Z(s)$ when $R_3 \neq 0$. This is true for all four networks of Fig. 1 and also for the circuits of Figs. 3, 4, and 5.

Table 2 is derived by putting $s = j\omega$ in Eq. (2) and calculating the magnitude $G(\omega)$ of $G(j\omega)$. The cases with and without C_0 must both be considered, the latter case being obtained from the former by letting $C_0 \to \infty$ (that is, $T_2 \to \infty$). The alternative expressions given in certain cases follow by the use of Table 1. Also given are the limiting values of the low-frequency gain $G(0)$ and the high-frequency gain $G(\infty)$.

[15] Again, the symbol $[\cdot]$ used in some of the formulas denotes a repetition of the square-bracketed expression which precedes it within the same formula.

APPENDIX 2

An Example of the Calculations Leading to Tables 3 and 4

We substitute into Eq. (7) for $Z(s)$ from Fig. 1(a) and equate the right-hand side with that of Eq. (8) to obtain

$$\begin{aligned}(1 &+ [(R_0 + R_3)C_0 + (R_1 + R_2)C_0 + R_1C_1 + R_2C_2]s \\ &+ [(R_0 + R_3)C_0(R_1C_1 + R_2C_2) + R_1R_2\{C_0(C_1 + C_2) + C_1C_2\}]s^2 \\ &+ [(R_0 + R_3)C_0R_1C_1R_2C_2]s^3)\,/ \\ &\quad (1 + R_0C_0s)\{1 + (R_1C_1 + R_2C_2)s + R_1C_1R_2C_2s^2\} \\ = &\quad (1 + (T_1 + T_4 + T_6)s + (T_1T_4 + T_1T_6 + T_4T_6)s^2 + T_1T_4T_6s^3)\,/ \\ &\quad (1 - T_2s)\{1 + (T_3 + T_5)s + T_3T_5s^2\}\end{aligned} \tag{25}$$

Comparison of the numerators and denominators on each side of this equation leads at once to the formulas in the upper third of Table 3(a). Note again that the poles T_3, T_5 are exactly the same as those of $Z(s)$. Unfortunately it is impractical to provide approximate formulas for the zeros T_1, T_4, and T_6, but they can be evaluated from the given cubic equation by standard techniques in any particular case.

To derive the remaining formulas in Table 3(a) one first equates the coefficients of corresponding powers of s in the numerators and denominators of Eq. (25) to obtain the following system of six equations:

$$\left.\begin{aligned} T_2 &= R_0C_0 \\ T_3 + T_5 &= R_1C_1 + R_2C_2 \\ T_3T_5 &= R_1C_1R_2C_2 \\ T_1 + T_4 + T_6 &= (R_0 + R_3)C_0 + (R_1 + R_2)C_0 + R_1C_1 + R_2C_2 \\ T_1T_4 + T_1T_6 + T_4T_6 &= (R_0 + R_3)C_0(R_1C_1 + R_2C_2) + R_1R_2\{C_0(C_1 + C_2) + C_1C_2\} \\ T_1T_4T_6 &= (R_0 + R_3)C_0R_1C_1R_2C_2 \end{aligned}\right\} \tag{26}$$

in the seven unknowns $R_0, R_1, R_2, R_3, C_0, C_1, C_2$. Any *one* of these quantities can thus be chosen arbitrarily, and the obvious choice would appear to be R_0. But in view of the fact that R_0 and R_3 occur in the combination $(R_0 + R_3)$ everywhere in system (26) except its first equation, it turns out that $(R_0 + R_3)$ is a better choice as independent variable. This choice also shows the parallels with the circuit of Fig. 3(a) more clearly. So we choose to solve system (26) for R_0, R_1, R_2, C_1, C_2 in terms of $(R_0 + R_3)$, and after much algebra obtain the formulas in the middle third of Table 3(a). Finally, by combining these results to eliminate $(R_0 + R_3)$ we deduce the lower third of the table. In the particular case $R_3 = 0$ system (26) contains a redundant equation, for then the T_i are constrained to satisfy

$$T_1T_4T_6 = T_2T_3T_5,$$

and so reduces to a system of five equations in six unknowns. Table 3(a) still correctly gives the results in this case, in terms of R_0 now.

Table 4 follows as before by setting $s = j\omega$ in Eq. (8).

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