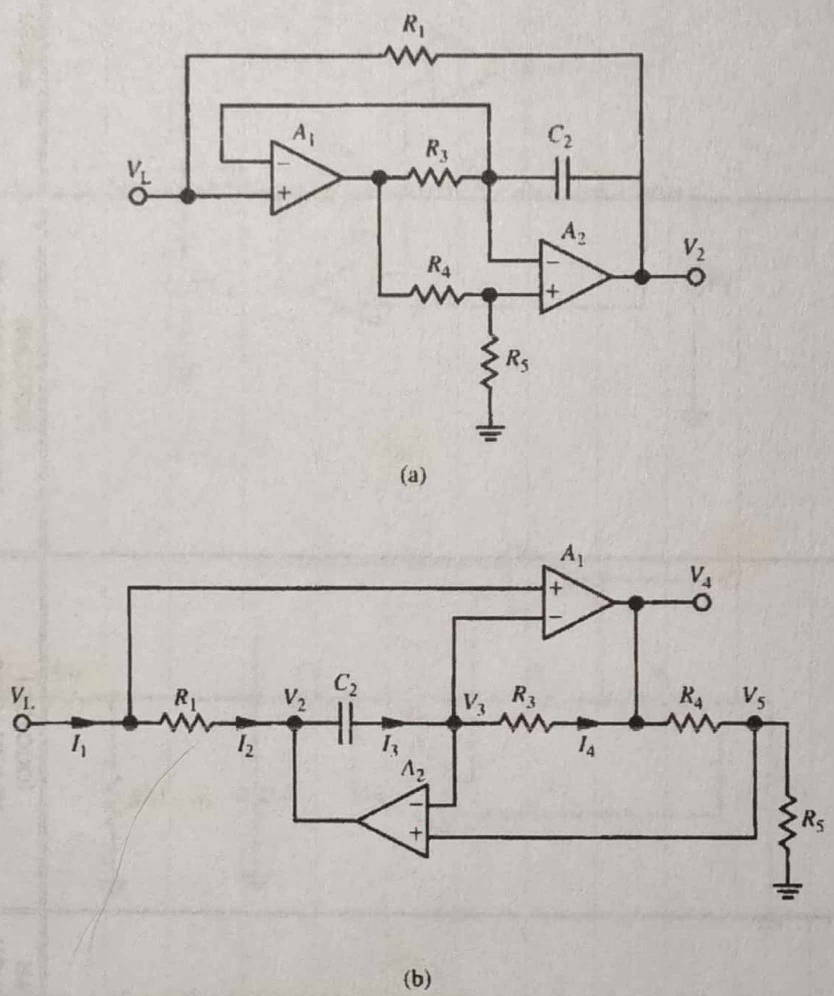


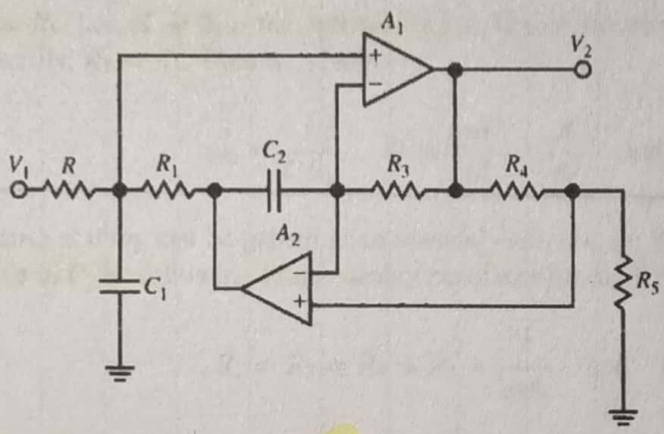
Figure 4.45 The general impedance converter (GIC). (a) Circuit obtained by combing Figs. 4.44a and b; (b) standard representation of the GIC.

We shall return to the GIC II structure later when we realize general high-order filters by simulating *LC* ladders. For our present situation, biquads, we use the GIC I.

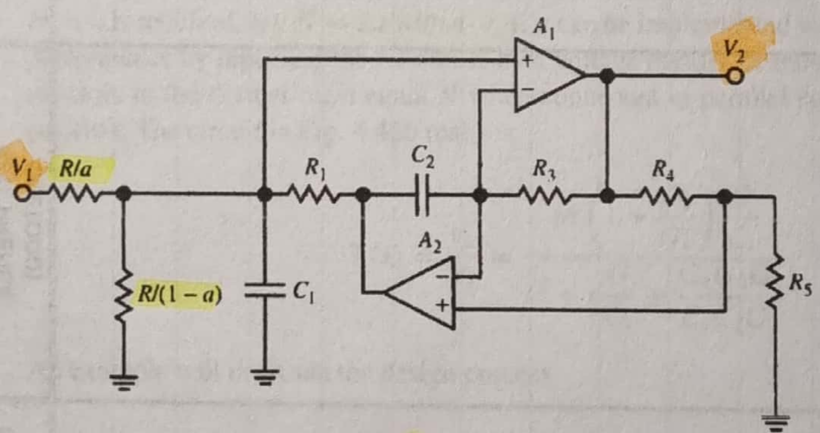
We have succeeded in building an opamp circuit whose input impedance looks like an inductor. Consequently, if we substitute that circuit, Fig. 4.45b, for the inductor in Fig. 4.39, we ought to arrive at an active bandpass biquad. There is still a small problem to be resolved: the biquad output voltage V_L must be taken from the inductor terminal that is not an opamp output and, hence, cannot be loaded without affecting filter performance. Naturally, we could use an opamp buffer, but that would require use of a third opamp. Fortunately, a simpler solution is available: recall that we found in Eq. (4.163) that the opamp output V_4 in Fig. 4.45b is proportional to V_L,

$$V_4 = V_L\left(1 + \frac{G_5}{G_4}\right) = V_L H \qquad (4.167)$$

We shall, therefore, use the output V_4 as the output terminal of our bandpass. The final circuit is shown in Fig. 4.46a, where we have the filter output relabeled V_2. This circuit is among the best performing second-order bandpass sections available. Its transfer function is obtained



(a)



(b)

Figure 4.46 (a) The GIC bandpass circuit for gain $H = 1 + R_4/R_5$; (b) the circuit for gain aH, $a < 1$. The optimal component choice is given in Eq. (4.170).

by substituting Eq. (4.161) for L in Eq. (4.153) and remembering that the output is taken at $V_4 = HV_L$. We obtain

$$T(s) = \frac{V_2}{V_1} = \frac{s\left(1 + \dfrac{G_5}{G_4}\right)\dfrac{G}{C_1}}{s^2 + s\dfrac{G}{C_1} + \dfrac{G_1 G_3 G_5}{C_1 C_2 G_4}} \tag{4.168a}$$

Let us again pick $C_1 = C_2 = C$. The bandpass circuit then has the center frequency ω_0, quality factor Q, and midband gain H

$$\omega_0 = \sqrt{\frac{G_1 G_3 G_5}{C^2 G_4}}, \quad Q = \frac{\omega_0 C}{G} = \frac{1}{G}\sqrt{\frac{G_1 G_3 G_5}{G_4}}, \quad H = 1 + \frac{G_5}{G_4} \tag{4.169a}$$

In principle, any value of $H > 1$ can be used. However, we shall demonstrate shortly that

$R_4 = R_5$, i.e., $H = 2$, is the optimal choice. Let us, therefore, choose $R_4 = R_5$ and select, arbitrarily, $R_3 = R_1$. Then we obtain

$$\omega_0 = \frac{1}{CR_1}, \quad Q = \frac{\omega_0 C}{G} = \frac{R}{R_1}, \quad \text{and} \quad H = 2 \tag{4.169b}$$

Because nothing can be gained from unequal resistors, we have for given ω_0 and a suitable choice of C the following simple design equations for the GIC bandpass circuit:

$$R_1 = R_3 = R_4 = R_5 = \frac{1}{\omega_0 C} \quad \text{and} \quad R = QR_1 \tag{4.170}$$

H is fixed at the value 2. A design with $H > 2$ is possible according to Eq. (4.181a) by selecting $R_4 > R_5$, but as shown in the following it will cause the errors in ω_0 and Q to increase. If $H < 2$ is required, say $H = 2a$ with $a < 1$, it can be implemented with no ill effects on circuit performance by replacing the resistor R by a voltage divider as shown in Fig. 4.46b. The two resistors in the divider must equal R when connected in parallel so as not to affect the pole position. The circuit in Fig. 4.46b realizes

$$T(s) = \frac{V_L}{V_1} = \frac{sa\left(1 + \dfrac{G_5}{G_4}\right)\dfrac{G}{C_1}}{s^2 + s\dfrac{G}{C_1} + \dfrac{G_1 G_3 G_5}{C_1 C_2 G_4}} \tag{4.168b}$$

An example will illustrate the design process.

EXAMPLE 4.11

Repeat the design of Example 4.10, but for a GIC biquad. The filter should have 0-dB midband gain; use LM741 opamps.

Solution

Specified were $f_0 = 12.5$ kHz and $Q = 10$. Choosing $C = 5$ nF, we obtain according to Eq. (4.170)

$$R_1 = R_3 = R_4 = R_5 = \frac{1}{\omega_0 C} = \frac{10^4}{3.927}\ \Omega = 2.55\ \text{k}\Omega, \qquad R = 25.5\ \text{k}\Omega$$

Because 0-dB gain is specified, i.e., $aH = 2a = 1$, we need to set $a = 0.5$. The resistor R is split, therefore, into two 51-kΩ resistors in parallel.

$$R/a = R/(1-a) = 51\ \text{k}\Omega$$

The circuit and its performance are shown in Fig. 4.47. We see that $f_0 = 12.3$ kHz for an error of -1.6% and $Q \approx 10.08$, increased by $\approx 0.8\%$ over the specifications. The midband gain is 0 dB as designed.