An Op Amp Tutorial

(Based on material in the book *Introduction to Electroacoustics and Audio Amplifier Design, Second Edition - Revised Printing*, by W. Marshall Leach, Jr., published by Kendall/Hunt, © 2001.)

An op amp has two inputs and one output. The circuit is designed so that the output voltage is proportional to the difference between the two input voltages. In general, an op amp can be modeled as a three-stage circuit as shown in Fig. 1. The non-inverting input is v_{I1}. The inverting input is v_{I2}. The input stage is a differential amplifier (Q_1 and Q_2) with a current mirror load ($Q_3 - Q_5$). The diff amp tail supply is the dc current source I_Q. The second stage is a high-gain stage having an inverting or negative gain. A capacitor connects the output of this stage to its input. This capacitor is called the compensating capacitor. Other names for it are lag capacitor, Miller capacitor, and pole-splitting capacitor. It sets the bandwidth of the circuit to a value so that the op amp is stable, i.e. so that it does not oscillate. The output stage is a unity-gain stage which provides the current gain to drive the load.

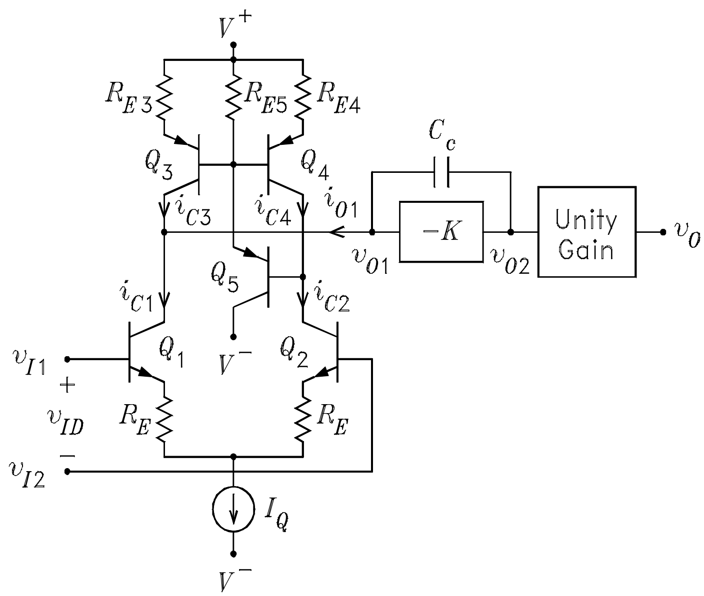

Figure 1: Op amp model.

If we assume that Q_1 and Q_2 are matched, that Q_3 and Q_4 are matched, that base currents can be neglected, and that the Early effect can be neglected, we can write the following equation for i_{O1}:

$$i_{O1} = i_{C1} - i_{C3} = i_{C1} - i_{C4} = i_{C1} - i_{C2} \tag{1}$$

But $i_{C1} + i_{C2} = I_Q$ and $i_{C1} = I_Q/2 + i_{c1}$. Thus we obtain

$$i_{O1} = 2i_{C1} - I_Q = 2i_{c1} \tag{2}$$

Open-Loop Transfer Function

We wish to solve for the transfer function for V_o/V_{id}, where V_{id} is the difference voltage between the two op amp inputs. First, we solve for the current I_{c1} as a function of V_{id}. For the diff amp, let us assume that the transistors are matched, that $I_{E1} = I_{E2} = I_Q/2$, the Early effect can be neglected, and the base currents are zero. In this case, the small-signal ac emitter equivalent circuit of the diff amp is the circuit given in Fig. 2(a). In this circuit, r_{e1} and r_{e2} are the intrinsic emitter resistances given by

$$r_{e1} = r_{e2} = r_e = \frac{V_T}{I_E} = \frac{2V_T}{I_Q} \tag{3}$$

Note that the dc tail supply I_Q does not appear in this circuit because it is not an ac source. From the emitter equivalent circuit, it follows that

$$I_{c1} = I_{e1} = \frac{V_{id}}{2\left(r_e + R_E\right)} \tag{4}$$

where $I_{c1} = I_{e1}$ because we have assumed zero base currents.

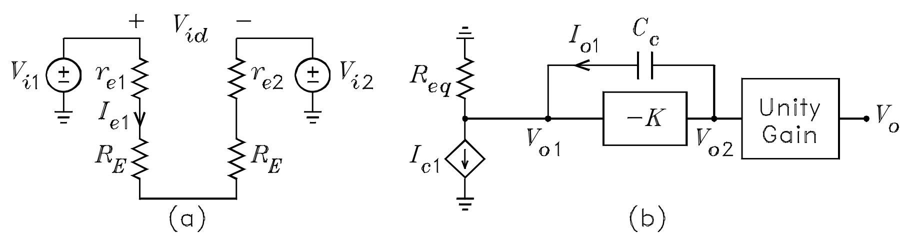

Figure 2: Circuit for calculating I_{e1}. (b) Circuit for calculating V_o.

Figure 2(b) shows the equivalent circuit which we use to calculate V_o. We assume that R_{eq} is the effective load resistance on the current $2I_{c1}$. In this case, the current which flows through the compensating capacitor C_c is given by

$$I_{o1} = 2I_{c1} + \frac{V_{o1}}{R_{eq}} = \frac{V_{id}}{r_e + R_E} - \frac{V_{o2}}{KR_{eq}} \tag{5}$$

where we have used Eq. (4) and the relation $V_{o1} = -V_{o2}/K$. The voltage V_{o2} is given by

$$V_{o2} = V_{o1} + \frac{I_{o1}}{C_c s} = \frac{-V_{o2}}{K} + \left[\frac{V_{id}}{r_e + R_E} - \frac{V_{o2}}{KR_{eq}}\right]\frac{1}{C_c s} \tag{6}$$

If we assume that the output stage has a gain that is approximately unity, then $V_o \simeq V_{o2}$. Let $G(s) = V_o/V_{id}$. It follows from Eq. (6) that $G(s)$ is given by

$$G(s) = \frac{V_o}{V_{id}} \simeq \frac{V_{o2}}{V_{id}} = \frac{KR_{eq}}{r_e + R_E} \times \frac{1}{1 + (1+K)\,R_{eq}C_c s} \tag{7}$$

This is of the form

$$G(s) = \frac{A}{1 + s/\omega_1} \tag{8}$$

where A and ω_1 are given by

$$A = \frac{KR_{eq}}{r_e + R_E} \qquad \omega_1 = 2\pi f_1 = \frac{1}{(1+K)\,R_{eq}C_c} \tag{9}$$

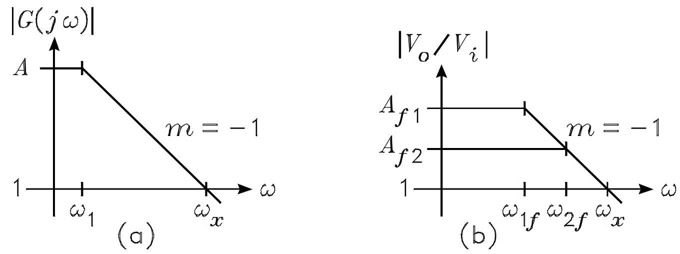

Figure 3: Asymptotic Bode magnitude plots. (a) Without feedback. (b) With feedback.

Gain Bandwidth Product

The asymptotic Bode magnitude plot for $|G(j\omega)|$ is shown in Fig. 3(a). Above the pole frequency ω_1, the plot has a slope of -1 dec/dec or -20 dB/dec. The frequency at which $|G(j\omega)| = 1$ is called the unity-gain frequency or the gain-bandwidth product. It is labeled ω_x in the figure and is given by

$$\omega_x = 2\pi f_x = A\omega_1 = \frac{K}{1+K}\frac{1}{r_e + R_E C_c} \simeq \frac{1}{(r_e + R_E)\, C_c} \tag{10}$$

where the approximation holds for $K \gg 1$. It follows that an alternate expression for $G(s)$ is

$$G(s) = \frac{A}{1 + sA/\omega_x} \tag{11}$$

For maximum bandwidth, f_x should be as large as possible. However, if f_x is too large, the op amp can oscillate. A value of 1 MHz is typical for general purpose op amps.

Example 1 *An op amp is to be designed for* $f_x = 4$ *MHz and* $I_Q = 50\ \mu A$*. If* $R_E = 0$*, calculate the required value for* C_c*.*

Solution. $C_c = 1/(2\pi f_x r_e) = I_Q/(4\pi f_x V_T) = 38.4$ pF, where we assume that $V_T = 0.0259$ V.

Slew Rate

The op amp slew rate is the maximum value of the time derivative of its output voltage. In general, the positive and negative slew rates can be different. The simple model of Fig. 1 predicts that the two are equal so that we can write

$$-SR \le \frac{dv_O}{dt} \le +SR \tag{12}$$

where SR is the slew rate. To solve for it, we use Eqs. (5) and (6) to write

$$V_{o2} = V_{o1} + \frac{I_{o1}}{C_c s} = \frac{-V_{o2}}{K} + \left(2I_{c1} + \frac{-V_{o2}}{KR_{eq}}\right)\frac{1}{C_c s} \tag{13}$$

This can be rearranged to obtain

$$sV_{o2}\left[1+\frac{1}{K}\left(1+\frac{1}{R_{eq}C_c}\right)\right]=\frac{2I_{c1}}{C_c} \tag{14}$$

If we assume that K is large and let $V_{o2}\simeq V_o$, this equation reduces to

$$sV_o\simeq\frac{2I_{c1}}{C_c} \tag{15}$$

The s operator in a phasor equation becomes the d/dt operator in a time-domain equation. Thus we can write

$$\frac{dv_o}{dt}=\frac{2i_{c1}}{C_c} \tag{16}$$

It follows that the slew rate is determined by the maximum value of i_{c1}. The total collector current in Q_1 is the sum of the dc value plus the small-signal ac value. Thus we can write $i_{C1}=I_Q/2+i_{c1}$. This current has the limits $0\leq i_{C1}\leq I_Q$. It follows that the small-signal ac component has the limits $-I_Q/2\leq i_{c1}\leq I_Q/2$. Thus we can write

$$\frac{-I_Q}{C_c}\leq\frac{dv_o}{dt}\leq\frac{+I_Q}{C_c} \tag{17}$$

It follows that the slew rate is given by

$$SR=\frac{I_Q}{C_c} \tag{18}$$

Example 2 *Calculate the slew rate of the op amp of Example 1.*

Solution. $SR=I_Q/C_c=1.30\ \mathrm{V}/\mu\mathrm{s}$.

Relations between Slew Rate and Gain-Bandwidth Product

If C_c is eliminated between Eqs. (10) and (18), we obtain the relation

$$SR=2\pi f_x I_Q\left(r_e+R_E\right)=4\pi f_x V_T\left(1+\frac{I_Q R_E}{2V_T}\right) \tag{19}$$

This equation clearly shows that the slew rate is fixed by the gain-bandwidth product if $R_E=0$. If $R_E>0$, the slew rate and gain bandwidth product can be specified independently.

Example 3 *Emitter resistors with the value $R_E=3\ k\Omega$ are added to the input diff amp in the op amp of Example 1. If f_x is to be held constant, calculate the new value of the slew rate and the new value of C_c.*

Solution. $SR=2\pi f_x I_Q\left(r_e+R_E\right)=5.07\ \mathrm{V}/\mu\mathrm{s}$. $C_c=I_Q/SR=9.86$ pF. The slew rate is greater by a factor of 3.9 and C_c is smaller by the same factor.

The above example illustrates how the slew rate of an op amp can be increased without changing its gain-bandwidth product. When R_E is added, ω_x decreases. To make ω_x equal to its original value, C_c must be decreased, and this increases the slew rate. It can be seen from Eq. (19) that the slew rate can also be increased by increasing I_Q. However, this causes ω_x to increase. To make ω_x equal to its original value, R_E must also be increased. Therefore, the general rule for increasing the slew rate is to either decrease C_c, increase I_Q, or both. All of these make ω_x increase. To bring ω_x back down to its original value, R_E must be increased. The change in R_E does not affect the slew rate.

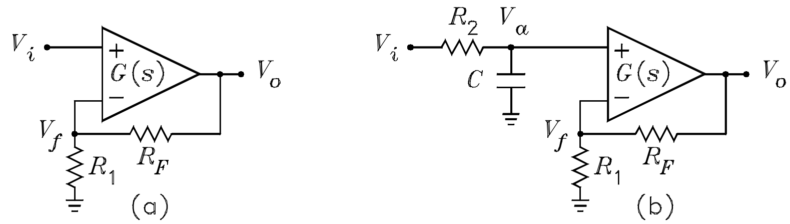

Figure 4: (a) Amplifier with feedback. (b) Amplifier with feedback preceded by a low-pass filter.

Closed-Loop Transfer Function

Figure 4(a) shows the op amp with a two resistor voltage divider connected as a feedback network. The output voltage can be written

$$V_o = G(s)(V_i - V_f) = G(s)(V_i - bV_o) \tag{20}$$

where b is the gain of the voltage divider given by

$$b = \frac{R_1}{R_1 + R_F} \tag{21}$$

Note that $0 \le b \le 1$. Eq. (20) can be solved for V_o/V_i to obtain

$$\frac{V_o}{V_i} = \frac{G(s)}{1 + bG(s)} = \frac{A_f}{1 + s/\omega_{1f}} \tag{22}$$

where Eq. (11) is used for $G(s)$. The dc gain A_f and the pole frequency ω_{1f} are given by

$$A_f = \frac{A}{1 + bA_f} \simeq \frac{1}{b} \tag{23}$$

$$\omega_{1f} = 2\pi f_{1f} = \omega_x \frac{1 + bA}{A} = \frac{\omega_x}{A_f} \simeq b\omega_x \tag{24}$$

where the f in the subscript implies "with feedback" and the approximations assume that $bA \gg 1$.

It can be seen from these equations that

$$A_f \omega_{1f} = A\omega_1 = \omega_x \tag{25}$$

Fig. 3(b) shows the Bode plot for $|V_o/V_i|$ for two values of A_f. As b is increased, A_f decreases and the bandwidth ω_{1f} increases so that the product of the two remain constant. This illustrates why ω_x is called the gain-bandwidth product.

Example 4 *An op amp has the gain bandwidth product $f_x = 8$ MHz. Calculate the upper -3 dB frequency f_u if the op amp is operated at a voltage gain of* 21.

Solution. The upper -3 dB frequency is equal to the pole frequency of the closed-loop transfer function. Thus $f_u = f_{1f} = f_x/A_f = 381$ kHz.

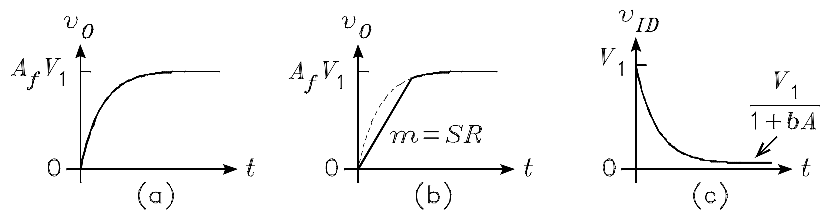

Figure 5: (a) No slewing step response. (b) Step response with slewing. (c) Differential input voltage.

Transient Response

Let the input voltage to the op amp in Fig. 4(a) be a step of amplitude V_1. We can write $v_I(t) = V_1 u(t)$, where $u(t)$ is the unit step function. The Laplace transform of $v_I(t)$ is $V_i(s) = V_1/s$. The Laplace transform of the output voltage is given by

$$V_o(s) = \frac{V_1}{s} \frac{A_f}{1 + s/\omega_{1f}} \tag{26}$$

The time domain output voltage is obtained by taking the inverse Laplace transform to obtain

$$v_O(t) = A_f V_1 [1 - \exp(-\omega_{1f} t)] u(t) \tag{27}$$

A plot of $v_O(t)$ is shown in Fig. 5(a).

The maximum time derivative of $v_O(t)$ occurs at $t = 0$ and is given by

$$\left.\frac{dv_O}{dt}\right|_{\max} = A_f V_1 \frac{d}{dt} \left\{[1 - \exp(-\omega_{1f} t)] u(t)\right\}\Big|_{t=0} = \frac{A_f}{\omega_{1f}} V_1 = \omega_x V_1 \tag{28}$$

If the derivative exceeds the slew rate of the op amp , the output voltage will be distorted as shown in Fig. 5(b), where the non-slewing response is shown by the dashed line. The maximum value of V_1 before the op amp slews is given by

$$V_{1\max} = \frac{SR}{\omega_x} = \frac{SR}{2\pi f_x} = I_Q (r_e + R_E) \tag{29}$$

Example 5 *Calculate the maximum value of V_1 for the op amps of Examples 2 and 3.*

Solution. For Example 1, $V_{1\max} = SR/\omega_x = 51.7$ mV. For Example 2, $V_{1\max} = 202$ mV. This is greater by about a factor of 3.9, i.e. the same as the ratio of the two slew rates.

Input Stage Overload

For the step input signal to the op amp with feedback in Fig. 4(a), the differential input voltage is given by

$$v_{ID}(t) = v_I(t) - b v_O(t) = \frac{V_1}{1 + bA} [1 + bA_0 \exp(-b\omega_x t)] u(t) \tag{30}$$

It follows that $v_{ID}(0) = V_1$ and $v_{ID}(\infty) = V_1/(1+bA)$. A plot of $v_{ID}(t)$ is shown in Fig. 5(c). The peak voltage occurs at $t = 0$. If the op amp is not to slew, the diff amp input stage must not overload with this voltage.

If base currents are neglected, the emitter and collector currents in Q_1 and Q_2 can be written

$$i_{E1} = i_{C1} = I_S \exp\left(\frac{v_{BE1}}{V_T}\right) \qquad i_{E2} = i_{C2} = I_S \exp\left(\frac{v_{BE2}}{V_T}\right) \tag{31}$$

where I_S is the BJT saturation current. The differential input voltage can be written

$$v_{ID} = (v_{BE1} - v_{BE2}) + (i_{E1} - i_{E2}) R_E \tag{32}$$

With the relation $i_{E2} = I_Q - i_{E1}$, these equations can be solved to obtain

$$v_{ID} = V_T \ln\left(\frac{i_{C1}}{I_Q - i_{C1}}\right) + (2i_{C1} - I_Q) R_E \tag{33}$$

The same equation holds for i_{C2} except v_{ID} is replaced with $-v_{ID}$.

Both i_{C1} and i_{C2} must satisfy the inequality $0 \le i \le I_Q$. At either limit of this inequality, one transistor in the diff amp is cut off. Let us consider the diff-amp active range to be the range for which i_{C1} and i_{C2} satisfy $0.05 I_Q \le i \le 0.95 I_Q$. This is the 5% to 95% range for the currents. When the diff amp is operated in this range, it follows from Eq. (33) that v_{ID} must satisfy

$$-v_{ID(\max)} \le v_{ID} \le -v_{ID(\max)} \tag{34}$$

where $v_{ID(\max)}$ is given by

$$v_{ID(\max)} = V_T \ln 19 + 0.9 I_Q R_E \tag{35}$$

If v_{ID} lies in this range, neither transistor in the diff amp can cut off and the op amp cannot exhibit slewing.

Example 6 *Calculate $v_{ID(max)}$ for $I_Q = 50\ \mu A$ for $R_E = 0$ and for $R_E = 3\ k\Omega$.*

Solution. For $R_E = 0$, we have $v_{ID(\max)} = V_T \ln 19 = 76.3$ mV. For $R_E = 3000$, $v_{ID(\max)} =$ 76.3 mV $+0.9 \times 50 \times 10^{-6} \times 3000 = 211$ mV. These values are greater than the values of $V_{1\max}$ calculated in Example 5 because the analysis here is based on the large signal behavior of the BJT.

Full Power Bandwidth

Figure 6 shows the output voltage of an op amp with a sine wave input for two cases, one where the op amp is not slewing and the other where the op amp is driven into full slewing. The full slewing waveform is a triangle wave. The slew-limited peak voltage is given by the slope multiplied by one-fourth the period, i.e.

$$V_{P\text{slew}} = SR \times \frac{T}{4} = \frac{SR}{4f} \tag{36}$$

where $T = 1/f$. When the op amp is driven into full slewing, an increase in the amplitude of the input signal causes no change in the amplitude of the output signal. If the frequency is doubled, the amplitude of the output signal is halved.

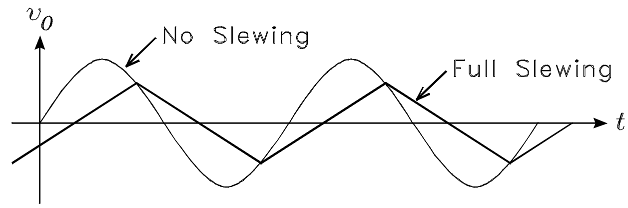

Figure 6: Non-slewing and full slewing output voltage waveforms.

Let the input voltage to the op amp be a sine wave. If the op amp does not slew and is not driven into peak clipping, the output voltage can be written $v_O(t) = V_P \sin \omega t$. The time derivative is given by $dv_O/dt = \omega V_P \cos \omega t$. The maximum value of $|dv_O/dt|$ occurs at $\omega t = n\pi$, where n is an integer, and is given by $|dv_O/dt|_{\max} = \omega V_P$. For a physical op amp , this cannot exceed the slew rate, i.e. $\omega V_P < SR$. It follows that the maximum frequency that the op amp can put out the sine wave without slewing is given by

$$f_{\max} = \frac{SR}{2\pi V_P} \tag{37}$$

Conversely, the peak output voltage without slewing is given by

$$V_{P(\max)} = \frac{SR}{2\pi f} \tag{38}$$

Let V_{clip} be the op amp clipping voltage at midband frequencies. If an op amp is driven at this level and the frequency is increased, the op amp will eventually slew and the maximum output voltage will decrease as the frequency is increased. The full power bandwidth frequency f_{FPBW} is defined as the highest frequency at which the op amp can put out a sine wave with a peak voltage equal to V_{clip}. It is given by

$$f_{\text{FPBW}} = \frac{SR}{2\pi V_{\text{clip}}} \tag{39}$$

Figure 7 shows the peak output voltage versus frequency for a sine wave input signal. At low frequencies, the peak voltage is limited to the op amp clipping voltage V_{clip}. As frequency is increased, the peak voltage becomes inversely proportional to frequency when the op amp is driven into full slewing and is given by $SR/4f$. The figure also shows the peak voltage below which the op amp does not slew. It is given by $SR/2\pi f$.

Example 7 *The op amps of Examples 2 and 3 have clipping voltages of* ± 13 *V. Calculate the full power bandwidth frequency if the op amps are not to slew at maximum output.*

Solution. For the op amp of Example 2, $f_{\text{FPBW}} = 1.3 \times 10^6/(2\pi 13) = 15.9$ kHz. For the op amp of Example 3, $f_{\text{FPBW}} = 5.07 \times 10^6/(2\pi 13) = 62.1$ kHz.

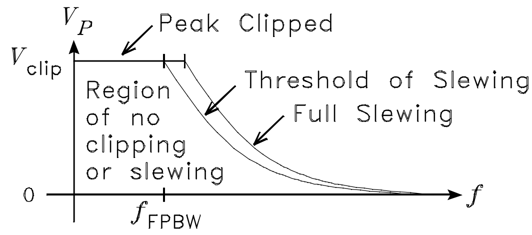

Figure 7: Maximum sine wave output voltage as a function of frequency.

Effect of an Input Low-Pass Filter

Step Input Signal

Figure 4(b) shows the op amp with a low-pass filter preceding its input. By voltage division, the transfer function for the voltage gain of the filter is

$$\frac{V_a}{V_i} = \frac{1/Cs}{R_2 + 1/Cs} = \frac{1}{1 + R_2 Cs} = \frac{1}{1 + s/\omega_a} \tag{40}$$

where $\omega_a = 1/R_2 C$. It follows that the overall transfer function for voltage gain of the op amp and filter is

$$\frac{V_o}{V_i} = \frac{A_f}{(1 + s/\omega_a)(1 + s/\omega_{1f})} \tag{41}$$

The transfer function for the differential input voltage is given by

$$\frac{V_{id}}{V_i} = \frac{V_a - bV_o}{V_i} = \frac{V_a}{V_i}\left(1 - b\frac{V_o}{V_a}\right) \tag{42}$$

With the aid of Eqs. (22) – (24), (40), and (41), this can be reduced to

$$\frac{V_{id}}{V_i} = \frac{A_f (1 + s/\omega_1)}{(1 + s/\omega_a)(1 + s/\omega_{1f})} \tag{43}$$

Let the input voltage be a step of amplitude V_1. Its Laplace transform is $V_i(s) = V_1/s$. It follows that the Laplace transform for V_{id} is

$$V_{id}(s) = \frac{A_f V_1 (1 + s/\omega_1)}{s(1 + s/\omega_a)(1 + s/\omega_{1f})} \tag{44}$$

For the case $\omega_a \neq \omega_{1f}$, the inverse transform of this is

$$v_{ID}(t) = V_1 \left[\frac{\omega_1}{\omega_{1f}} + \frac{\omega_a - \omega_1}{\omega_{1f} - \omega_a}\exp(-\omega_a t) - \frac{\omega_a}{\omega_{1f}} \times \frac{\omega_{1f} - \omega_1}{\omega_{1f} - \omega_a}\exp(-\omega_{1f} t)\right] \tag{45}$$

The maximum value of $v_{ID}(t)$ occurs at the time t_1 which satisfies $dv_{ID}(t_1)/dt = 0$. It is straightforward to show that t_1 is given by

$$t_1 = \frac{1}{\omega_{1f} - \omega_a}\ln\left(\frac{\omega_{1f} - \omega_1}{\omega_a - \omega_1}\right) \tag{46}$$

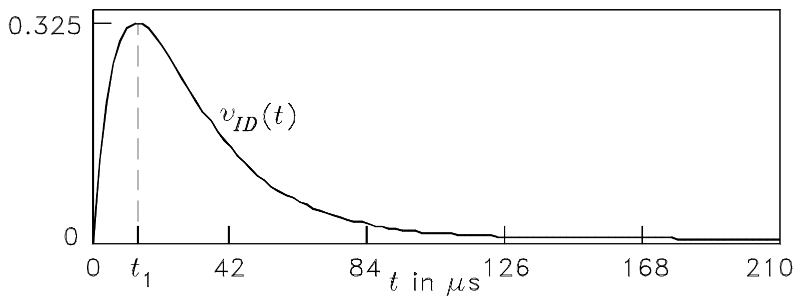

Figure 8: Plot of v_{ID} as a function of t for Example 8.

The value of $v_{ID}(t_1)$ is

$$v_{ID}(t_1) = V_1 \left[\frac{\omega_1}{\omega_{1f}} + \left(1 - \frac{\omega_1}{\omega_{1f}}\right) \left(\frac{\omega_a - \omega_1}{\omega_{1f} - \omega_1}\right)^{\omega_{1f}/(\omega_{1f} - \omega_a)} \right] \tag{47}$$

If the diff amp is not to leave its linear region, this voltage must not exceed $v_{ID\max}$ given by Eq. (35). When this is true, the op amp cannot slew with the step signal.

Example 8 *An op amp with clipping voltages of* ± 13 *V has the open-loop bandwidth* $f_1 = 5$ *Hz, the closed-loop gain* $A_f = 10$*, and the closed-loop bandwidth* $f_{1f} = 100$ *kHz. The op amp is preceded by a low-pass filter having a bandwidth* $f_a = 50$ *kHz. The op amp input is a voltage step which drives the output to the clipping level. Calculate* t_1 *and* $v_{ID}(t_1)$*.*

Solution. The amplitude of the input step is $V_1 = 13/10 = 1.3$ V. Eqs. (46) and (47) give $t_1 = 13.9$ μs and $v_{ID}(t_1) = 0.325$ V. A plot of $v_{ID}(t)$ versus t is shown in Fig. 8. The low-pass filter has reduced the peak overload of the diff amp by the factor $1.3/0.325 = 4$ or by 12 dB.

Example 9 *The op amp of Example 8 has a diff amp that is biased at* $I_Q = 50$ μA*. Calculate the minimum value of* R_E *if the diff amp is not to leave its linear region for the value of* $v_{ID}(t_1)$*. Assume that* C_c *is adjusted so that* f_{1f} *does not change with* R_E*.*

Solution. In Eq. (35), we let $v_{ID\max} = v_{ID}(t_1) = 0.325$ V. Thus R_E is given by $R_E = (0.325 - V_T \ln 19)/0.9 I_Q = 5.53$ kΩ, where we have assumed that $V_T = 0.0259$ V.

Square-Wave Input Signal

The transient examples that we have looked at so far assume that the op amp input voltage is a step and that the initial value of the output voltage is zero. Transient response measurements on op amps are usually made with a square-wave input signal, not a step. A square wave can be written as a series of steps. Thus it may seem that the results obtained for the step can be applied directly for the square wave. This is true only if the calculations are modified to account for the non-zero initial value of the op amp output voltage.

Let a square wave be applied to an op amp that switches from its negative level to its positive level at $t = 0$. The input and output voltage waveforms are illustrated in Fig. 9(a),

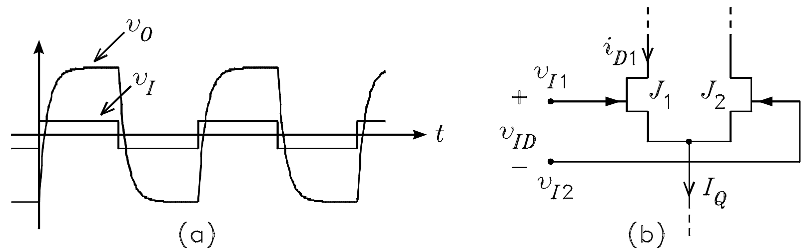

Figure 9: (a) Square wave input and output voltages. (b) JFET diff amp input stage.

where it is assumed that no low-pass filter precedes the op amp input. At $t = 0^-$, let the input voltage be $-V_1$. The output voltage is $-A_f V_1$. At $t = 0^+$, the input voltage switches to $+V_1$, but the output voltage is still at $-A_f V_1$. Thus the differential input voltage at $t = 0^+$ is $v_{ID} = V_1 - v_O/A_f = 2V_1$.

It follows from this result that the results obtained for the step input apply to the square wave input if the amplitude of the step is doubled. This is equivalent to saying that the amplitude of the step must equal the total change in voltage of the square wave between its negative and positive levels. The same conclusion holds when the op amp is preceded by a low-pass filter.

JFET Diff Amp

We have seen above that the addition of emitter resistors to the diff amp transistors reduces the gain bandwidth product of the op amp . If the compensation capacitor is then reduced to bring the gain bandwidth product back up to its original value, the slew rate is increased. Another method of accomplishing this is to replace the BJTs with JFETs. A JFET diff amp is shown in Fig. 9(b). For a specified bias current, the JFET has a much lower transconductance than the BJT. In effect, this makes it look like a BJT with emitter resistors. For this reason, resistors in series with the JFET sources are omitted in the figure. The analysis in this section also applies to the MOSFET diff amp.

The JFET drain current can be written

$$i_D = I_{DSS}\left(1 - \frac{v_{GS}}{V_{TO}}\right)^2 \tag{48}$$

where I_{DSS} is the drain-source saturation current (the value of i_D with $v_{GS} = 0$), V_{TO} is the threshold voltage (which is negative), v_{GS} is the gate to source voltage, and $V_{TO} \le v_{GS} \le 0$. For the drain current in either JFET in the diff amp to be in the range of $0.05I_Q$ to $0.95I_Q$, the maximum differential input voltage is given by

$$v_{ID\max} = |V_{TO}|\left(\sqrt{0.95} - \sqrt{0.05}\right)\sqrt{\frac{I_Q}{I_{DSS}}} = 0.751\,|V_{TO}|\sqrt{\frac{I_Q}{I_{DSS}}} \tag{49}$$

The JFET transconductance is given by

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = \frac{2I_{DSS}}{-V_{TO}}\left(1 - \frac{V_{GS}}{V_{TO}}\right) = \frac{2}{-V_{TO}}\sqrt{I_D I_{DSS}} \tag{50}$$

To convert a formula derived for the op amp with a BJT diff amp into a corresponding formula for the JFET diff amp, the BJT intrinsic emitter resistance r_e is replaced with $1/g_m$ for the JFET. Thus the gain bandwidth product of the op amp with the JFET diff amp is given by

$$\omega_x = 2\pi f_x = \frac{g_m}{C_c} \tag{51}$$

where it is assumed that there is no added resistor in series with the source leads, i.e. $R_E = 0$.