Phase-Shift Modulator

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An old saying is that “if you want an amplifier try building an oscillator and if you want an oscillator then try building an amplifier”! The reason is that it is often challenging to build an amplifier that is free from instability. However, with a phase-shift modulator, the amplifier really *is* an oscillator – one with an input that allows for control of the low-frequency portion (i.e. the audio band) of the output.

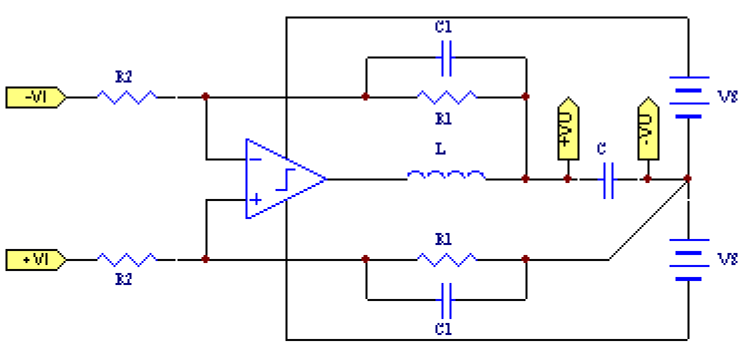
Advantages

- Great simplicity
  - The amplifier is essentially a high-power comparator
- Instability, a concern with typical amplifiers, is not so much of an issue
  - The amplifier is already an oscillator
- Feedback is easily taken after the output filter
  - This results in load insensitivity with respect to frequency response

Disadvantages

- Limited loop gain unless higher-order loops are employed
  - Higher-order loops may adversely impact the stability of the amplifier
- Frequency variability
  - More difficult to filter, especially at duty cycle extremes
  - Not as simple to integrate multiple channels

Basic Circuit Diagram



List of Variables

D	Duty cycle, a value from 0 to 1
V_S	Supply voltage of split $\pm V_S$ supply
C	Output filter capacitance
L	Output filter inductance
R	Load connected after filter
ω_n	Natural frequency of output filter
ω_s	Switching frequency of amplifier
ω_0	Switching frequency at idle
τ	Amplifier propagation delay

Other variables are as shown in the basic circuit diagram.

Open-Loop Gain

The determination of open-loop gain for this class-D amplifier differs from a typical linear amplifier in that the *slope* at the zero crossing of the residual carrier is used, rather than the actual gain in the forward path (remember, the forward path in a class-D is simply a comparator).

A given error at the output results in a shift of the residual carrier. This in turn results in a duty cycle change that tends to counteract the error based on the slope of the residual carrier at the zero crossing.

$$v = L\frac{di}{dt} \qquad i = C\frac{dv}{dt}$$

$$\left.\frac{dv}{dt}\right|_{\max} = \frac{i_{\max}}{C} \qquad i_{\max} = \frac{D(1-D)V_S}{f_S L}$$

$$\left.\frac{dv}{dt}\right|_{\max} = \frac{2\pi D(1-D)V_S\omega_n^2}{\omega_S} \qquad \Delta v = \frac{2\pi D(1-D)V_S\omega_0^2}{\omega_S}\Delta t$$

The open loop gain A_{OL} is derived by taking the output voltage ΔV_O that results from a given error voltage ΔV_E and simply dividing the two expressions.

$$A_{OL} = \frac{\Delta V_O}{\Delta V_E}$$

$$\Delta V_E = \frac{2\pi D(1-D)V_S\omega_n^2}{\omega_S}\Delta t$$

$$\Delta V_O = (2D-1)V_S - (2D'-1)V_S$$

$$\Delta V_O = 2(D-D')V_S$$

$$D' = \frac{t_{on} - 2\Delta t}{t_{on} - 2\Delta t + t_{off} + 2\Delta t} = D - \frac{2}{T_S}\Delta t$$

$$D' = D - \frac{\omega_S}{\pi}\Delta t$$

$$\Delta V_O = \frac{2\omega_S V_S}{\pi}\Delta t$$

$$\Delta V_O = \frac{1}{\pi^2 D(1-D)}\left(\frac{\omega_S}{\omega_n}\right)^2 \Delta V_E$$

$$\frac{\Delta V_O}{\Delta V_E} = \frac{1}{\pi^2 D(1-D)}\left(\frac{\omega_S}{\omega_n}\right)^2$$

$$A_{OL} = \frac{1}{\pi^2 D(1-D)}\left(\frac{\omega_S}{\omega_n}\right)^2$$

Setting D=0.5 (i.e. the amplifier at idle) yields the *small-signal open-loop gain*:

$$A_{OL} = \frac{4}{\pi^2}\left(\frac{\omega_S}{\omega_n}\right)^2$$

The switching frequency versus duty cycle has been determined empirically. It is the same as for a hysteresis modulator for 0.1<D<0.9 and very close for 0.05<D<0.95. The frequency tends to be higher than that of a hysteresis modulator at the duty cycle extremes, which is an advantage for filtering.

$$\omega_S = 4D(1-D)\omega_0$$

Substituting this into the open-loop gain expression above yields the *large-signal open-loop gain*:

$$A_{OL} = \frac{16}{\pi^2}D(1-D)\left(\frac{\omega_0}{\omega_n}\right)^2$$

Condition for Oscillation

The condition for oscillation is when the net phase shift around the loop is 180°. If a simple lead-lag RC network is used as part of the feedback network (as is shown in the basic circuit diagram), then this condition is met when:

$$\tan^{-1}\omega R_1 C_1 - \tan^{-1}\frac{\omega R_1 C_1}{1+\frac{R_1}{R_2}} - \omega\tau - \tan^{-1}\frac{\omega\frac{L}{R}}{1-\omega^2 LC} = 0$$

These contributions are due to (respectively):

- Phase lead from feedback network
- Phase lag from feedback network
- Phase lag from propagation delay
- Phase lag from output filter

The phase contribution from the output filter tends to be roughly -180°, since the switching frequency is usually well above the corner frequency of the output filter. The propagation delay of the amplifier is a fairly constant value for a given design. This leaves the phase contribution of the feedback network as the primary variable for adjusting the switching frequency.