Use transconductance amplifiers to make programmable active filters. The OTAs offer a 10-octave tuning range by using predictable V_{BE}-versus-I_c relationships.

Operational transconductance amplifiers (OTAs) can provide control of many active filter parameters over a 60-dB dynamic range. Ranges this wide are possible because the controlling signals are currents and not voltages. And current scan represent filter variables such as corner frequency, center frequency. Q or bandwidth, better than a voltage, since the relationship between the collector current and the base-to-emittor voltage drop of a silicon transistor is predictable over extremely wide ranges.

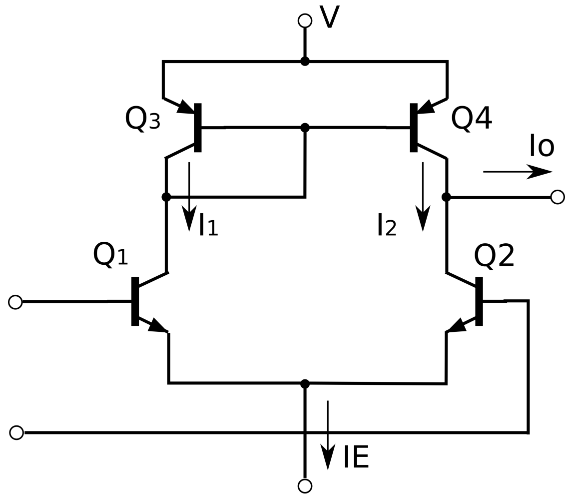
Over the last few years state-variable and biquad active filter circuits have been adopted by many companies as universal filter building blocks. Analog multipliers scale the impedance elements of these filters, thereby achieving parameter control.[1] Commercially available multipliers, however, are limited when it comes to voltage range, temperature drift and linearity. Because of these drawbacks, programmable filters based on analog voltage multipliers are generally restricted to operation over a 3-to-6-octave range.
But by exploiting the transistor's V_{BE} *vs* I_C, relationship in an OTA, you can use current signals to provide filter control over many decades.

The operational transconductance amplifier consists of a matched pair of npn transistors, connected as a variable transconductance multiplier, and a matched pair of pnp transistors, connected as a current mirror (Fig. 1). The small signal relationship between the collector difference currents i_1-i_2 and the base-emitter difference voltages v_1-v_2 can be expressed by

$$i_1 - i_2 = (i_E/2V_T)(v_1 - v_2)$$

in which V_T is the thermal voltage and is typically 26 mV at room temperature, and i_E is the combined emitter current.

Due to the current-mirror action of the pnp pair, the collector current of Q_4 equals i_1. the collector current of Q1. The difference current, i_1-i_2 (shown as i_o) is proportional to the product of the emitter current and the [1] 1.



1. The basic operational transconductance amplifier uses two pnp transistor: and two npn's to provide a linear conversion of voltage to current.

differential voltage applied to the npn pair. This product holds up over 3 very wide range of common-emitter current values-- a property that makes the OTA very powerful in programmable filter synthesis and wide-range gain control.[2,3] Let's see how this can be put to use by examining on integrator-- the basic building block of active filters.

Programmable integrators from OTAs

A programmable integrator can be built with an op amp and a transistor may (Fig. 2). To ensure the validity of a small-signal approximation for Q_1 and Q_2, the differential input signal is scaled down with resistor R_1 and R_2. For the values given in Fig. 2, the difference voltage becomes

$$v_1 - v_2 = (V^- - V^+)\, R_2/(R_1 + R_2)$$
$$= 3.2 \times 10^{-3}\, (V^- - V^+).$$

Thus an input signal of ±5V, pk-pk, is attenuated into a ±16 mV swing at the bases of Q_1 and Q_2 Transistor Q_5 connected in diode form, provides emitter bias for the current-mirror transistors, while the 100-kΩ potentiometer compensates for the overall voltage offset of the OTA.
The pot should be adjusted so that the difference current is zero for $V^- = V^+ = 0$.

The transfer function of the circuit can easily be found if you use Laplace transforms:

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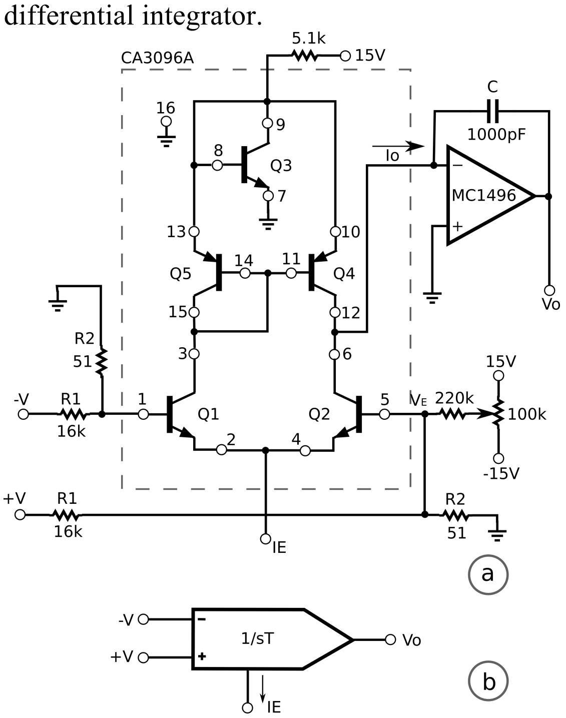
$$V_o = -I_o/SC = (i_E/2V_T)(v_1 - v_2)/SC$$
which can be simplified to
$$Vo = (V^+ - V^-)/S\tau,$$
where
$$\tau = [2\, V_T\, C\, (R_1 + R_2)/R_2\,]\,/i_E.$$
These equations show that the circuit integrates the differential input signal and that the constant of integration, τ, is inversely proportional to i_E. The constant, τ, can be programmed externally if i_E is controlled with a variable current generator. When this is all put together, you have a programmable differential integrator.



2. To build a programmable integrator, the OTA can be combined with an op amp and some resistors to form the basic building block of an active filter.

Of course, if one input is grounded, the circuit is an inverting or non-inverting integrator. To minimize any error introduced by the op amp at low current levels, an op amp with very low input bias current, such as one with a FET input or super-beta transistor input, should be used.

Combined integrators make a filter

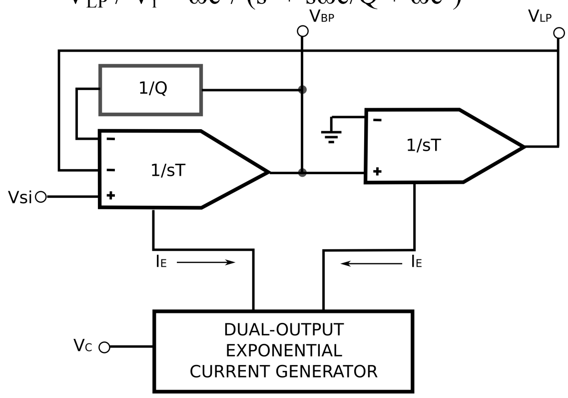
Two integrators can be combined to make a programmable filter that exhibits simultaneous second-order, low-pass and band-pass responses (Fig. 3). To build a circuit using this block diagram, you also need a dual-output current generator to control both integrators. The actual circuit is shown in Fig. 4 (offset adjustment pots have not been included, for the sake of simplicity). Now let's see how it works.

If we start with the block diagram of Fig. 3 and use Laplace transforms to find the band-pass and low-pass outputs, we get:

$$V_{BP} = (V_i - V_{LP} - V_{BP}/Q)/s\tau \text{ and}$$
$$V_{LP} = V_{BP}/s\tau,$$

where 1/Q represents the fraction of V_{BP} fed back to the input of the first integrator. After combining these equations and rearranging terms, we get:

$$V_{BP} / V_i = s\omega c / (s^2 + s\omega c/Q + \omega c^2) \text{ and}$$
$$V_{LP} / V_i = \omega c^2/ (s^2 + s\omega c/Q + \omega c^2)$$

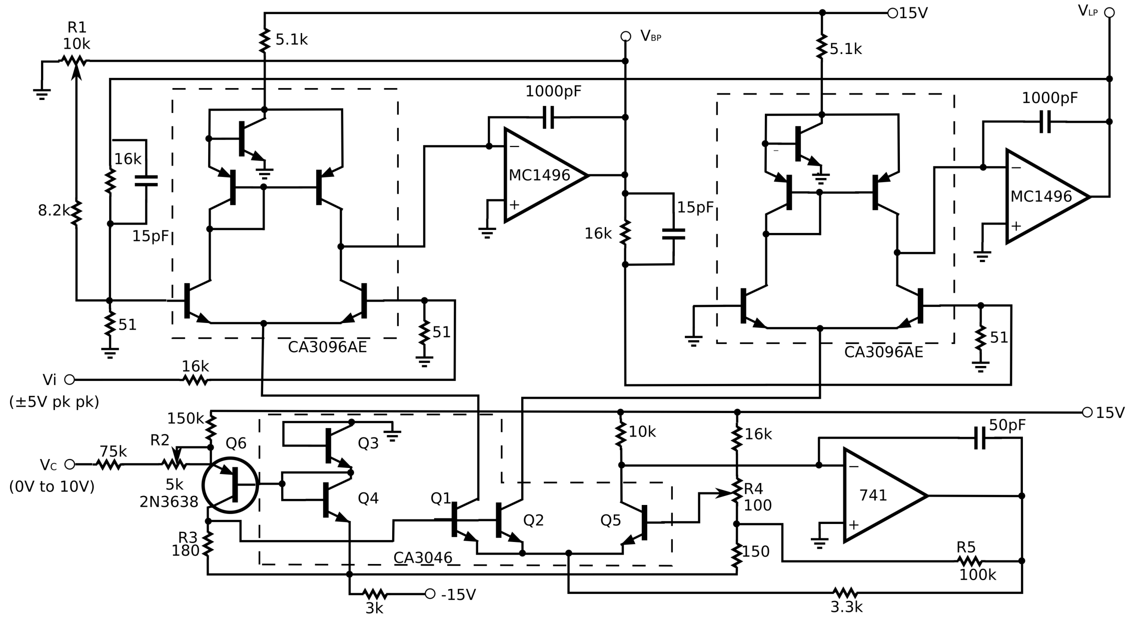


3. This block diagram of a second-order integrator shows how you can obtain both band-pass and low-pass outputs from a single programmable filter.

where
$$\omega c = 1/\tau \approx 62 \times 10^6\,(i_E).$$
The center frequency of the bandpass can now be calculated :
$$f_c = \omega c/2\Pi \approx 10^7(i_E).$$
This shows that f_c is directly proportional to the common-emitter current that feeds the OTAs. Thus if i_E changes from 1 µA to 1 mA, f_c varies accordingly from 10 Hz to 10 kHz (a three-decade or 10-octave range).

When you have ranges this wide, exponential parameter control is usually more convenient than linear control, since changes can then be expressed in decibels rather than linear steps. The exponential control over a 10-Hz-to-10-kHz range can be accomplished by use of a dual-output exponential current generator. Let's specify the circuit parameters so that a. 1V increment in the control voltage, V_C, results in a one-octave change in i_E and hence in f_c. The relationship between f_c and V_C can then be written as

$$f_c = 10 \times 2^{V_C}\text{ Hz } (0 \le V_C \le 10).$$



4. The complete schematic of the programmable active filter uses but three transistor arrays and three op amps, yet offers a 10-octave tuning range. The filter delivers band-pass and low-pass outputs.

The circuit of Fig. 4 provides this wide range. Its Q factor is set by R. and can range from 0.5 to several hundred. As the Q increases, the circuit will eventually break into oscillation at a frequency of f_c. When used in this mode, the circuit can operate as a voltage-controlled quadrature oscillator, since the low-pass output lags the band-pass output by 90°. For good sine purity, though, a nonlinear limiting element should be added in the feedback loop to ensure output amplitude stabilization. As an added bonus, the Q and f_c are totally independent, and any adjustment of either has no effect on the other.

The circuit works like this . . .

A common problem with state-variable circuits is the so-called Q enhancement at high frequencies which results from the phase lag of finite bandwidth op amps. This effect can be reduced if you use high-speed op amps and counteract the phase lag of the amps with some intentionally inserted phase lead in the interstage coupling circuitry. This is exactly what the 15-pF capacitor-s do when connected in parallel with 16- kΩ coupling resistors.

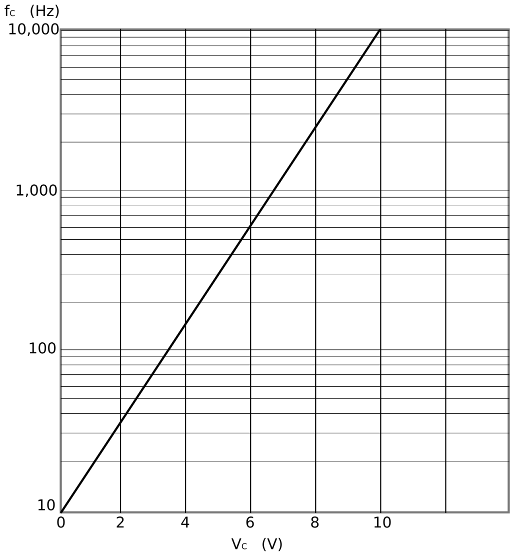
Since the amount of phase lag is likely to vary from one op-amp type to another, the capacitor values shown in Fig. 4 are only indicative of the order of magnitude involved. The optimum capacitor values can be determined experimentally.

The dual-output current generator used in Fig. 4 consists of three matched transistors, Q_1, Q_2, and Q_5, a level shifter, Q_6, and a regulator op amp. The op amp keeps the current through Q_5 constant, and thus develops a temperature tracking reference voltage at the emitter of Q_5 for the emitters of Q_1, and Q_2. Level shifter Q6 scales the input voltage range into a nominal swing of 180 mV at the common bases of Q_1 and Q_2. This permits the collector currents of these transistors to undergo a three-decade change as Vc is swept from 0 to +10 V. Diode-connected transistors Q_3 and Q_4 translate the reference voltage of the entire exponential circuit to below ground, so that the collectors of Q_1 and Q_2 can feed directly to the common emitters of the OTAs.

Potentiometer R_2 adjusts the width of the exponential range, and R_4 shifts the entire range up or down and, in turn, controls the scale factor of the exponential function. The 50-pF capacitor in the feedback circuit of the regulator, together with R_6, prevents high-frequency oscillation of the op amp. Resistor R_5 compensates for any error introduced by the emitter bulk resistances, r_E of Q_1 and Q_3. These resistances could cause the relationship between the collector current and the base-emitter voltage drop to be no longer exponential in the upper current range. The value of R_5 is calculated by the formula:

$$R5 \approx 2 \times 130\ \Omega \times R_6/\ r_E.$$

Since r_E is only about 10 Ω or so, the value of R_5 easily compensates for the bulk-resistance error. Non perfect matching between Q_1 and Q_2 means that the collector currents of these transistors are likely to differ, and this will affect both f_c, and Q. However, this doesn't create a problem, since the mismatch is automatically compensated for during the initial calibration adjustments of R_1, R_2, and R_4. The performance of the filter for a Q setting of about 50 is shown in Fig. 5.



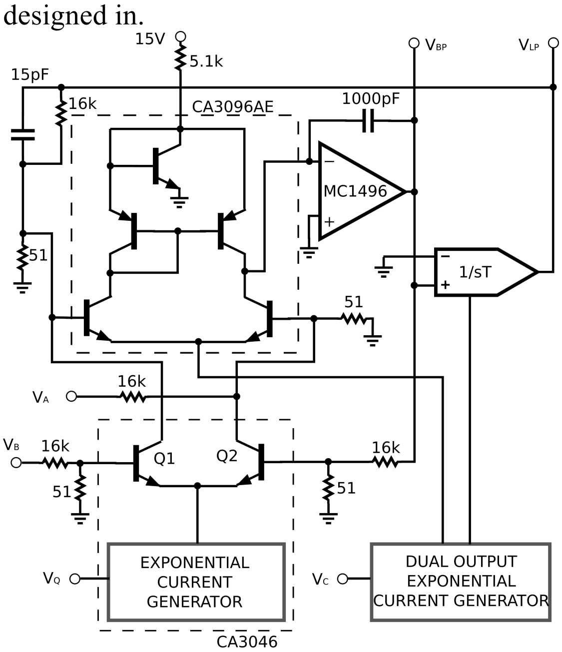
5. The response of the programmable filter is linear (in decibels) over a 10-V tuning range. It has a corner or center frequency within a 10-Hz-to-10-kHz range.

The transfer characteristics of the OTAs, as well as those of the dual-output current generator, depend on V_T, the thermal voltage. The filter thus has a certain amount of temperature sensitivity, and for applications where this is critical, you can minimize it. All you have to do is replace the 51-Ω input resistors (R_2, in Fig. 2) and the 180-Ω current-generator input resistor (R_3 in Fig. 4) with thermistor elements that have compensating characteristics.

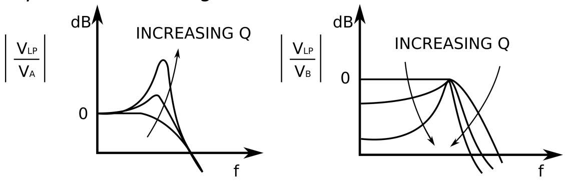
Automatic Q control for the filter

In the circuit of Fig. 4, Q is determined by the setting of R_1. However, the Q can also be made programmable if you add a transconductance multiplier and associated current generator. This modification is shown in Fig. 6, where, for clarity, only the circuit components directly involved in the change are shown schematically; the rest of the circuit is shown in block diagram form. For optimum performance, make provisions to zero the voltage offset between Q_1 and Q_2.

The response of this circuit to input signal levels is the same as for the circuit shown in Fig. 4, except that the value of Q is now determined by a control voltage, V_Q. The gain of the circuit at f_c equals Q, and this may be a problem for large values of Q. This is because the amplifiers in the filter will saturate for large voltage inputs, unless the input is intentionally maintained below a predetermined value. To eliminate this possible saturation problem, an additional input to the filter, V_B, can be designed in.



6. You can program the Q of the tunable filter by adding a transconductance multiplier in the input stage and an exponential current generator to control it.



7. The low-pass response for the modified circuit of Fig.6 shows that for the v_A input, gains will reach values higher than one (a), while for the v_B input, gains are adjusted for a maximum of one (b).

Since the signal V_B is applied to the filter through the multiplier that controls the Q, it undergoes some attenuation. This attenuation is inversely proportional to the value of Q. Thus the gain of the filter at f_c is unity. regardless of the value of Q. The typical low-pass responses to the two input lines are shown in Fig. 7.

References

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2. Jung, W. G., “Get Gain Control of 80 to 100 dB," Electronic Design No. 13, June 21, 1974, pp. 94-99.

3. Franco, "Hardware Design of a Real-Time Musical System" Department of Computer Science Technical Report. University of Illinois. October. 1974.