Use off-the-shelf linear ICs for sophisticated audio designs

With a little creativity and a good understanding of common linear circuits, you can produce high-performance audio designs—such as V/F converters, VCAs, preamps and panning circuits—with simple, inexpensive parts.

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Op amps can serve in applications other than audio amplifiers; you can apply them and other off-the-shelf linear ICs to create more sophisticated audio and electronic-music circuits. And although these applications present stringent performance demands, don't assume you'll need excessively complicated and expensive designs; the linear circuits described here achieve high performance at low cost.

As an example of the unusual use of conventional components, consider **Fig 1**'s exponential V/F converter. Suitable for use in music synthesizers, this circuit provides an output that changes its frequency one octave in response to a 1V control-input variation. Similar to conventional nonlinear converters, it exploits

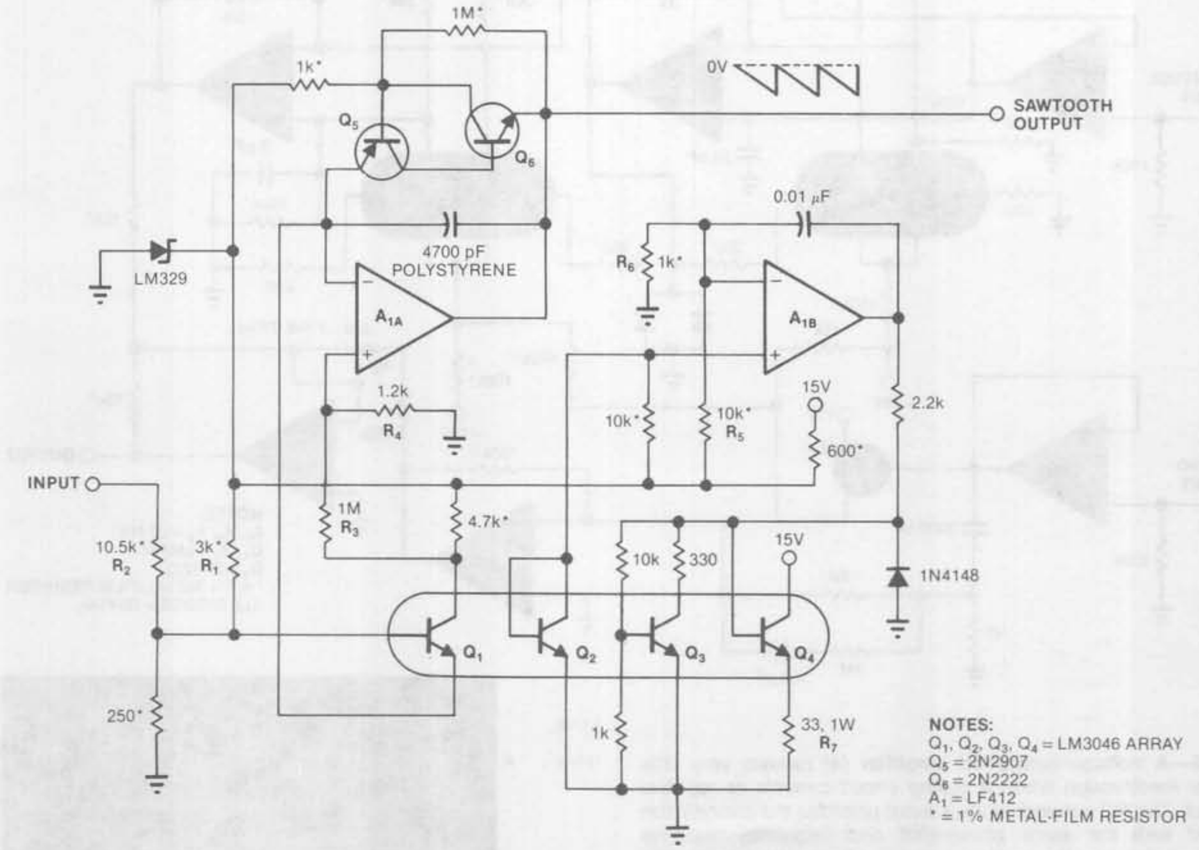

Fig 1—A temperature-compensated transistor array *stabilizes this exponential voltage-to-frequency converter. Q_2 senses the array's temperature and A_{1B} drives chip heater Q_4, maintaining a stable operating environment for logarithmic converter Q_1. The circuit's negative-going sawtooth output results from A_1's integration of Q_1's collector current until it reaches Q_5 and Q_6's threshold.*

A servo loop eliminates temperature-dependent drift

the logarithmic relationship between a transistor's base-emitter voltage and collector current. However, a unique thermal servo loop eliminates temperature-dependent transfer-function errors.

Generating the circuit's negative-going sawtooth output, op amp A_{1A} integrates Q_1's collector current until Q_5 and Q_6 turn on. These feedback transistors then discharge the integrating capacitor, the output rises to 0V and the cycle repeats.

Q_1 is a vital element in this circuit because its V_{BE}/I_C characteristics ensure that A_{1A}'s input current—and thus the output frequency—remain an exponential function of the control voltage. Assisting Q_1, transistors Q_2, Q_3, Q_4 and op amp A_{1B} form a temperature-controlled loop that stabilizes Q_1's operating point by thermally compensating the LM3046 transistor array.

To perform this compensation, Q_2's base-emitter junction senses array temperature, and Q_4 heats the chip. A_{1B} varies this heater transistor's dissipation until Q_2's V_{BE} drop equals the reference level set by R_5 and R_6. Furthermore, Q_3 and R_7 limit Q_4's maximum operating power and ensure proper servo functioning during circuit power-up.

In addition to stabilizing Q_1's collector bias, the LM329 6.9V reference also fixes the Q_5/Q_6 firing point. These two transistors exhibit opposing temperature coefficients, so their switching threshold is compensated to approximately 100 ppm/°C. The polystyrene integrating capacitor's −120-ppm/°C TC cancels remaining firing-level uncertainty.

To establish a 20-Hz quiescent output frequency, R_1 biases the circuit's input. R_2 trims the converter's transfer gain. Op-amp input resistors R_3 and R_4 maintain exponential conformity to within 0.5% from 20 Hz to 15 kHz by providing first-order compensation for Q_1's bulk emitter resistance.

Reduce VCA feedthrough

The next circuit, a voltage-controlled amplifier (VCA), also uses a simple error-correction scheme to improve its performance (**Fig 2**). Commonly employed

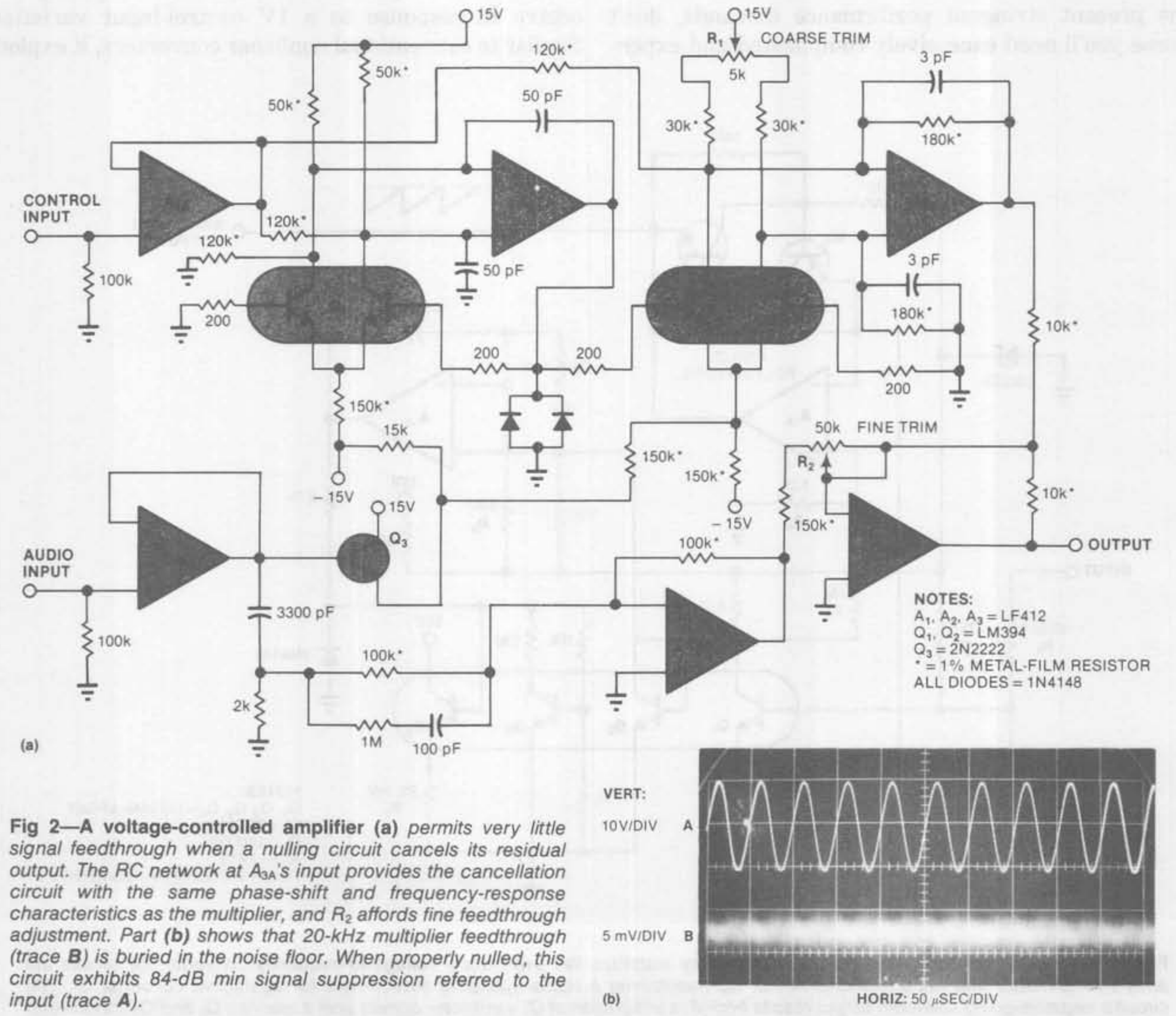

Fig 2—A voltage-controlled amplifier (a) *permits very little signal feedthrough when a nulling circuit cancels its residual output. The RC network at A_{3A}'s input provides the cancellation circuit with the same phase-shift and frequency-response characteristics as the multiplier, and R_2 affords fine feedthrough adjustment. Part* **(b)** *shows that 20-kHz multiplier feedthrough (trace* **B***) is buried in the noise floor. When properly nulled, this circuit exhibits 84-dB min signal suppression referred to the input (trace* **A***).*

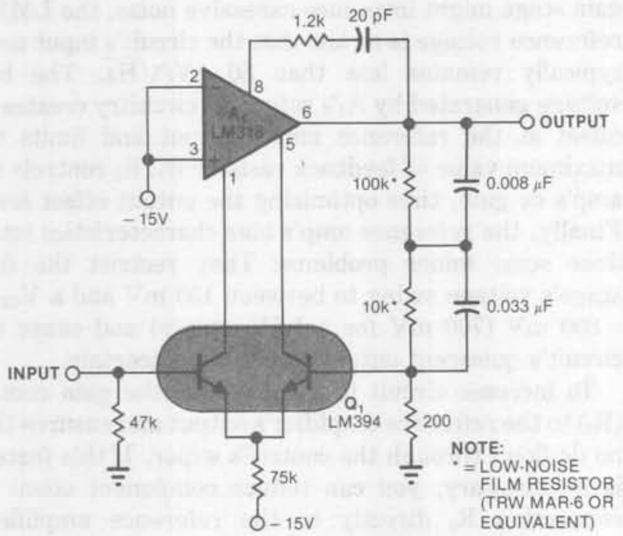

Fig 3—Lower input noise *results when an external transistor pair replaces the first stage of an op amp. This design avoids the loop-instability problems often caused by adding external stages to an op amp because the additional transistors replace rather than complement the LM318's input devices.*

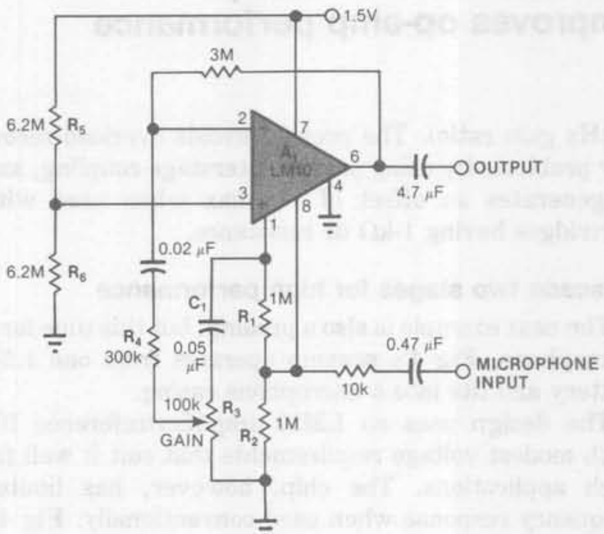

Fig 4—This microphone preamp *operates from supplies as low as 1.5V. Its LM10 reference amplifier provides a voltage gain of 100 with an input noise level of less than 50 nV/$\sqrt{Hz}$. The chip's op-amp section amplifies the reference output by an additional 20 dB.*

in recording-studio mixing consoles, VCAs must permit minimal signal feedthrough when their control inputs reach 0V. Conventional analog multipliers aren't optimal for this application; although they behave well in high-gain regions, they afford inadequate high-frequency signal suppression with their control channels off.

To reduce the feedthrough levels of its simple VCA, **Fig 2a**'s design uses a nulling technique. Op amps A_{1A} and A_{1B} and emitter-follower Q_3 buffer the circuit's control and audio inputs and then feed these signals to a transconductance multiplier composed of A_{2A}, Q_1 and Q_2. A_{2B} converts Q_2's differential collector currents to a single-ended audio output. R_1 allows coarse feedthrough trimming at 10 kHz to approximately −65 dB (relative to the input signal level).

To further reduce feedthrough, A_{3A} and A_{3B} null the multiplier's OFF-state output with the audio input. The RC network at inverter A_{3A}'s input provides phase shift and frequency response similar to the multiplier's feedthrough characteristics—thus, residual signals cancel out when A_{3B} combines the outputs from A_{3A} and A_{2B}. The nulling circuit's gain control, R_2, allows fine feedthrough trimming to −84 dB at 20 kHz.

To adjust this VCA, apply a 20V p-p, 20-kHz sine wave to the audio input, and with the control input grounded, adjust R_1 for minimum output from A_{2B}. Then trim R_2 for the lowest level at A_6's output. **Fig 2b** illustrates the circuit's typical feedthrough signal (trace **B**) for a 20-kHz input (trace **A**) when properly trimmed. Note that circuit noise almost obscures the waveform.

In addition to its excellent feedthrough suppression, this VCA exhibits only 0.05% total harmonic distortion (THD) throughout its 60-kHz power bandwidth. To obtain best circuit performance, construct it on a rigid circuit board, enclose it in a well-shielded box and employ proper grounding and noise reduction.

Replace an op amp's inputs

Instead of correcting a circuit's inherent errors as in the previous designs, you can use your knowledge of an IC's internal workings to eliminate deficiencies before they occur. For example, **Fig 3** illustrates an RIAA-equalized phono preamp with a noise figure less than 2 dB typ; it uses an LM394 ultralow-noise transistor pair at an LM318's compensation inputs instead of the device's internal input transistors.

This technique achieves lower noise than the unaltered op amp without introducing loop instability. Stability criteria become especially critical in RIAA circuits because the equalization function requires 100% feedback at high frequencies. Connecting the op amp's unused inputs to the negative supply shuts off the device's first differential pair and allows the external devices to operate into the LM318's output stages.

The distortion performance of **Fig 3**'s circuit exceeds the measurement capability of most test equipment: THD within the audio band remains less than 0.002% for outputs to 0.1V rms, and 20-kHz distortion rises only to 0.007% at 5V rms. Referred to a 10-mV input, the preamp's noise level equals −90 dB, with absolute values measuring 0.55 µV and 70 pA rms over a 20-kHz bandwidth—levels below the noise generated by most phono cartridges.

Fig 3's phono preamp also performs well in transient-intermodulation (TIM) tests. When fed with a 200-mV input—consisting of 10- and 11-kHz sine waves equally mixed—the circuit generates a 1-kHz output of only 80 µV. The TIM level, therefore, is 0.004% (or 0.0008% if you include the RIAA function's 14-dB (5:1) 10- to

1-kHz gain ratio). The preamp avoids overload-recovery problems by using only dc interstage coupling, and it generates an offset of 1V max when used with cartridges having 1-kΩ dc resistance.

Cascade two stages for high performance

The next example is also a preamp, but this time for a microphone. **Fig 4**'s preamp operates from one 1.5V battery and fits into a microphone casing.

The design uses an LM10 amplifier/reference IC, with modest voltage requirements that suit it well for such applications. The chip, however, has limited frequency response when used conventionally. **Fig 4**'s circuit extends the device's operating range by cascading its reference amplifier and op amp to form a high-gain preamp.

Here are the details. The microphone input drives A_1's reference amplifier, which has a unity-gain bandwidth of 500 kHz. Feedback around this stage yields an ac voltage gain of 100. The op amp, which operates more slowly than the reference amp, provides an additional 20 dB of voltage amplification, resulting in overall circuit gain of 60 dB. The preamp's bandwidth extends to 10 kHz unloaded and reaches 5 kHz with a 500Ω load.

Although using a reference amplifier as a low-level gain stage might introduce excessive noise, the LM10's reference voltage is so low that the circuit's input noise typically remains less than 50 $nV/\sqrt{Hz}$. The bias voltage generated by A_1's reference circuitry creates an offset at the reference amp's output and limits the maximum value of feedback resistor R_1. R_2 controls the amp's dc gain, thus optimizing the output offset level. Finally, the reference amp's bias characteristics introduce some minor problems: They restrict the first stage's voltage swing to between 150 mV and a V_{CC} of −800 mV (700 mV for a 1.5V supply) and cause the circuit's quiescent output to become uncertain.

To increase circuit life, C_1 couples the gain control (R_3) to the reference amplifier's output and ensures that no dc flows through the control's wiper. If this feature is unnecessary, you can reduce component count by connecting R_4 directly to the reference amplifier's output (A_1, pin 1) and using R_1 as a gain control. However, the 70-nA bias current that normally flows through the feedback resistor might increase noise.

DAC pans audio signal

Another approach to solving audio-design problems involves multiplying DACs. **Fig 5** shows a digitally programmable panning circuit that splits an input between two output channels. The DAC input code (D) determines the relative levels of the two channels, and op amps A_{1A} and A_{1B} convert the DAC's complementary current outputs into voltages. The relationship of the

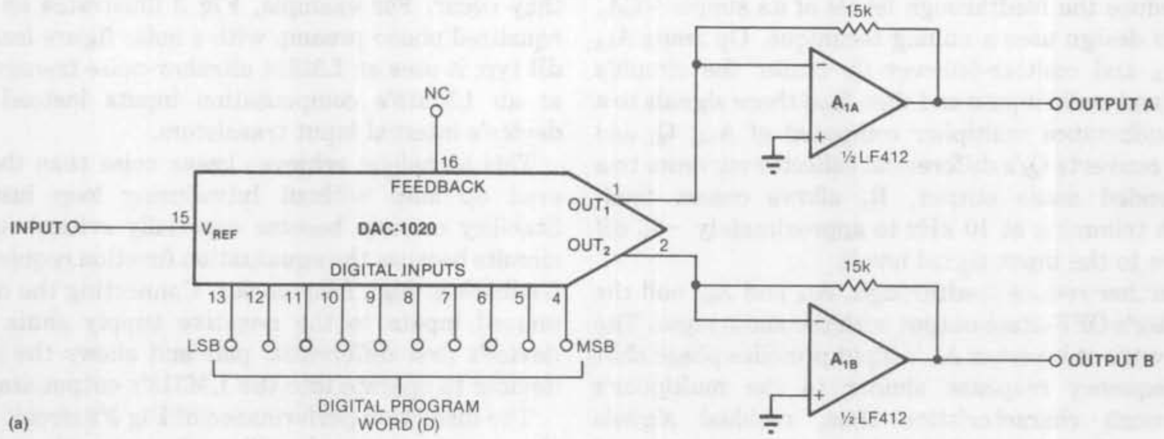

Fig 5—A DAC controls the ratio of two output signals *in the pan-pot circuit shown in* **(a)**. *Op amps A_{1A} and A_{1B} convert the DAC's complementary output currents to voltage signals. The ratio of the pan pot's outputs* **(b)** *changes as a digital ramp drives the DAC input. The sum of the two signals remains constant, regardless of the digital control word.*

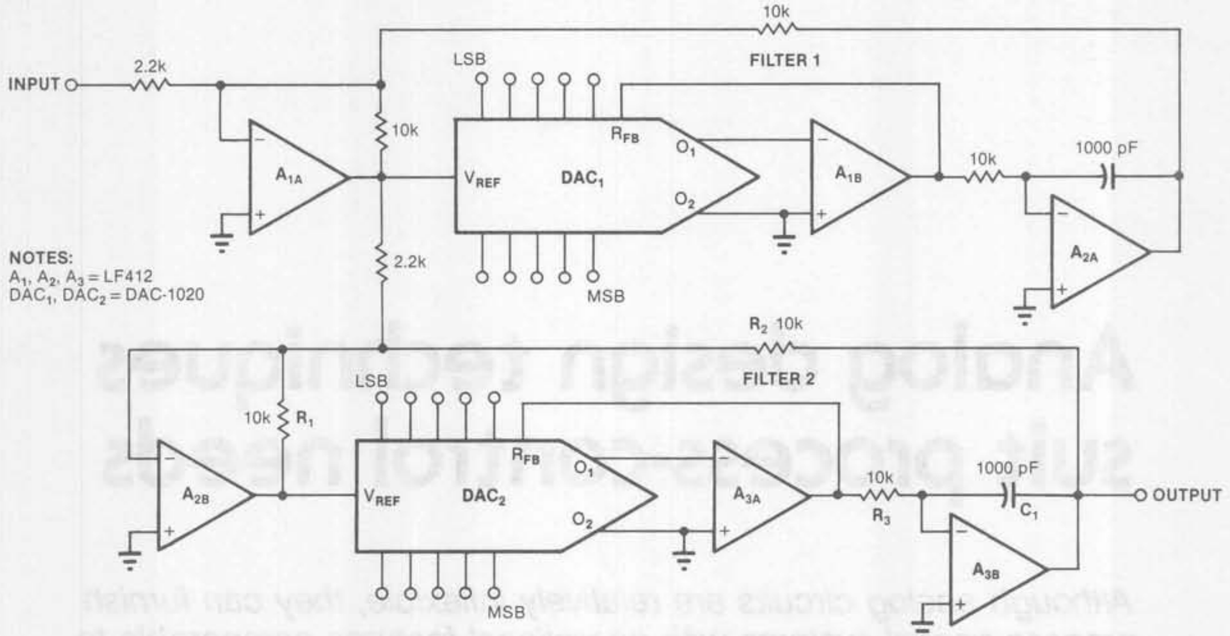

Fig 6—Two DACs control the passband *of this first-order filter. The input codes determine the gains of integrators A_{2A} and A_{3B}.*

two output signals to the digital input in **Fig 5**'s design is given by:

$$\text{OUTPUT A} = -\left[\text{INPUT} \times \left(\frac{3D}{2048}\right)\right],$$

$$\text{OUTPUT B} = -\left[\text{INPUT} \times \frac{3(1024 - D)}{2048}\right]$$

and **Fig 5b** illustrates the circuit's operation when a digital ramp controls the DAC inputs.

This circuit differs from conventional DAC applications because the converter's internal feedback resistor remains unconnected; the circuit's discrete resistors permit better matching of the output channels. Each op amp exhibits 300-ppm/°C gain drift arising from mismatches between the DAC's ladder resistors and op-amp feedback elements. You can eliminate these small errors, though, by using a separate DAC, with complementary digital inputs, for each channel.

You can also employ DACs in programmable-audio-filter designs. **Fig 6**'s circuit, a first-order bandpass network, utilizes two identical state-variable filters. The DACs control the gains of integrators A_{2A} and A_{3B} and thus determine the filters' cutoff frequencies.

You achieve a bandpass characteristic by connecting the first filter's high-pass output (taken from the output of A_{1A}) to the second filter's input. Filter 2's low-pass output then contains only those signals that lie within the passband established by the DACs' input codes.

You can determine filter 2's low-pass cutoff frequency (f_C) as a function of DAC_2's program input (D) with

$$f_C = \frac{R_1}{R_2}\left[\frac{D}{2048\pi R_3 C_1}\right].$$

For **Fig 6**'s components,

$$f_C = \frac{D}{2048\pi(10k)(1 \times 10^{-9})}.$$

Filter 1's high-pass output function equals the derivative of the low-pass expression **(reference)**. **EDN**

Reference

Analog Devices Inc, *Application Guide to CMOS Multiplying D/A Converters,* 1978, pg 32.

Author's biography

Jim Williams, applications manager of National Semiconductor's Linear Applications Group (Santa Clara, CA), specializes in instrument development and analog circuit design. Before joining National, he served as a consultant at Arthur D Little Inc and ran the Instrumentation Development Lab at the Massachusetts Institute of Technology. A former student of psychology at Wayne State University, Jim lists spare-time interests that include tennis, art and collecting antique scientific instruments.