ANALOG DEVICES

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APPLICATION NOTE

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Video VCAs and Keyers Using the AD834 and AD811

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INTRODUCTION

Voltage-controlled amplifiers (VCAs) built from analog multipliers take one of two forms. In the first, the multiplier acts as a voltage-controlled attenuator ahead of a fixed-gain amplifier. This type of VCA is used in applications where only a moderate maximum gain, but a fairly high maximum loss, are needed. In the second, the variable attenuation is placed in the feedback path around an op amp, which, in fact, implements an analog divider, more suitable for applications requiring high gains.

This application note describes practical circuits in which the wide bandwidth of the Analog Devices AD834 Four-Quadrant Multiplier and the AD811 Current-Feedback Op Amp are exploited to provide a video-quality VCA with a maximum gain of 12 dB (×4) or 20 dB (×10), based on the first of the above methods. A slightly modified form of this VCA, using two multipliers whose outputs are summed, provides the first of two video keyer designs; a second design uses global negative feedback around the multipliers to achieve improved accuracy and some simplification.

A VIDEO-QUALITY VCA

The VCA is shown in Figure 1. The AD834 multiplies the signal input by the control voltage. Its outputs are in the form of differential currents from a pair of open collectors, ensuring that the full bandwidth of the multiplier (which exceeds 500 MHz) is available for certain applications. In this case, more moderate bandwidth is obtained using current-to-voltage conversion, provided by the AD811 op amp, to realize a practical amplifier with a single-sided ground-referenced output. Using feedback resistors R8 and R9 of 511 Ω the overall gain ranges from −70 dB for V_G ~0 to +12 dB (a numerical gain of four) when V_G = +1 V.

The −3 dB bandwidth is 90 MHz (Figure 2) and is essentially independent of gain. The response can be maintained flat to within ±0.1 dB from dc to 40 MHz at full gain (Figure 3) with the addition of an optional capacitor of about 0.3 pF across the feedback resistor R8. The circuit produces a full-scale output of ±4 V for a ±1 V input, and can drive a reverse-terminated load of 50 Ω or 75 Ω to ±2 V. Figure 4 shows the typical pulse response.

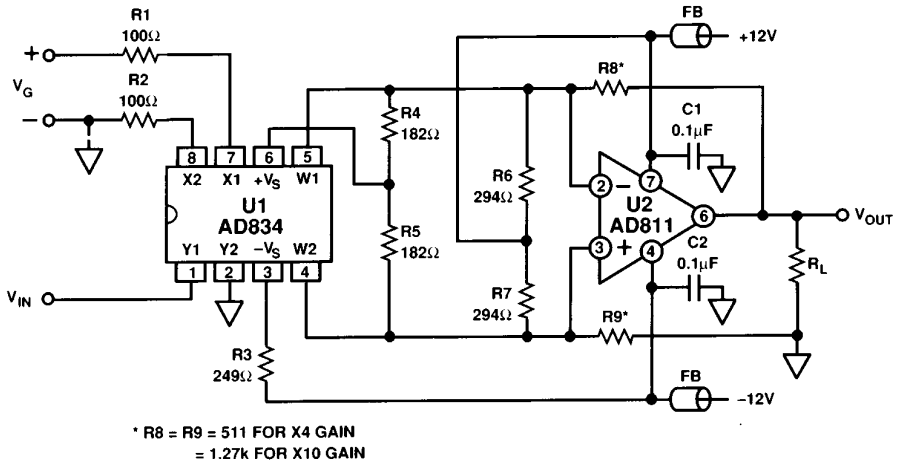


Figure 1. Complete VCA Provides Up to 20 dB of Gain (G = BW = 25 MHz) and a Bandwidth of Over 90 MHz (G = 12 dB)

The gain can be increased to 20 dB (×10) by raising R8 and R9 to 1.27 kΩ, with a reduction of the −3 dB bandwidth to about 25 MHz (also shown in Figure 2) and a maximum output voltage of ±9 V using the ±12 V supplies. It is not necessary to alter R6 and R7 for the high gain version of the amplifier, although an optimized design would raise these slightly to restore the common-mode voltage at the input of the AD811 to +5 V.

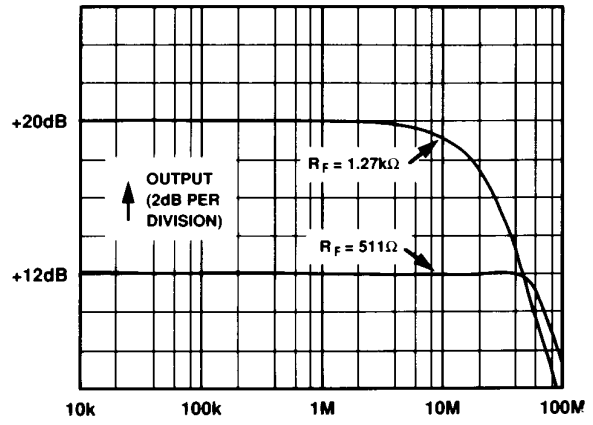


Figure 2. Small-Signal Response of the VCA Shows a −3 dB Bandwidth of 90 MHz for the 12 dB Version and 25 MHz for the 20 dB Version

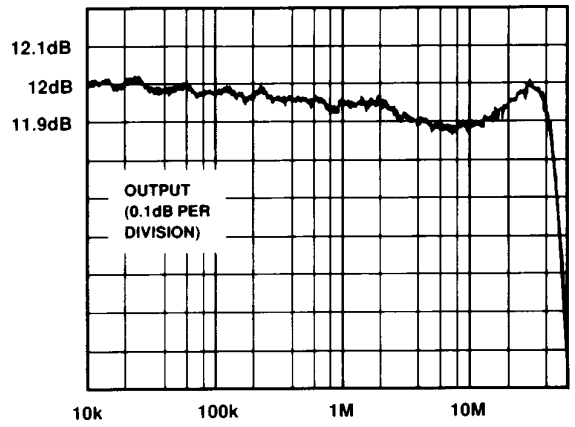


Figure 3. AC Response Can Be Held Flat to Within ±0.1 dB from DC to 40 MHz by Addition of a 0.1 pF Capacitor Across R8

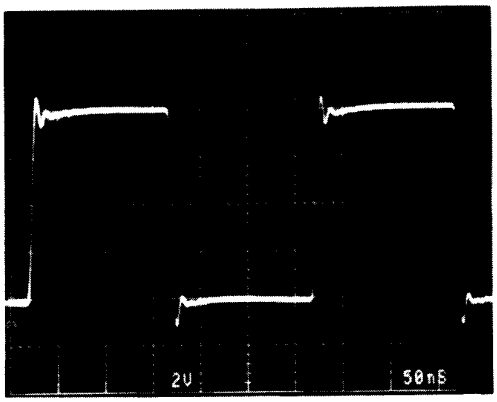


Figure 4. Full-Output Pulse Response for the 12 dB Amplifier

The gain-control input may be a positive or negative ground-referenced voltage, or fully differential, depending on the user's choice of connections at Pins 7 and 8. As shown, a positive value of V_G results in an overall noninverting response. Reversing the sign of V_G simply causes the sign of the overall response to invert. In fact, although we have called this a voltage-controlled amplifier, it can just as well be used as a general-purpose four-quadrant multiplier with good load-driving capabilities and fully symmetrical responses from X- and Y-inputs.

We have used the Y-input of the multiplier for the signal, since this port is slightly more linear than the X-input, and have shown X2 and Y2 grounded. These inputs each draw about 45 μA of bias current, so the grounded (unused) inputs should be terminated preferably in the same resistance as the source, in each case, to minimize offset voltages. The resistance of the signal source may in some cases be essentially zero (as in the case of a transformer-coupled input, or certain signal generators); note that a doubly terminated cable line of impedance Z_O will present a dc resistance of $Z_O/2$ at the input. Resistors R1 and R2 have been included in Figure 1 to minimize the likelihood of small aberrations arising in the signal path in those cases where V_G is derived from a source having poor HF characteristics; they may be omitted in the four-quadrant multiplier application.

High-frequency circuits such as those described herein are sensitive to component layout, stray capacitance, and lead lengths. Use a ground plane and make short, direct connections to ground. Bypass the power-supply connections—inductance in the power-supply leads can form resonant circuits that produce response peaking or even sustained oscillations.

Circuit Analysis

To understand the operation of the VCA, we need first to consider the scaling properties of the AD834, which is actually an accurate nonlinear (two-input) voltage-controlled current source. Figure 5 shows a simplified schematic of the whole VCA.

The exact transfer function for the AD834 would show that the differential voltage inputs at X1, X2 and Y1, Y2 are first multiplied together, divided by the scaling voltage of 1 V (determined by the on-chip bandgap reference) and the resulting voltage is then divided by an accurate 250 Ω resistor to generate the output current. A simplified form of this transfer function is

$$I_W = (X_1 - X_2)(Y_1 - Y_2) \times 4mA \quad (1)$$

where I_W is the differential current output from W1 to W2 and it is understood that the inputs X_1, X_2, Y_1, and Y_2 are expressed in volts. Thus, when both differential inputs are 1 V, I_W is 4 mA; this current is laser-calibrated to close tolerance, which simplifies the use of the AD834 in many applications. Note carefully the direction of this current in determining the correct polarity of the output connections.

It is easy to show that the output of the AD811 is

$$V_{OUT} = 2 \times I_W \times R_F \quad (2)$$

where R_F is the feedback resistor. For R_F = 500 Ω (499 Ω is the nearest standard resistor value), the overall transfer function of the VCA becomes

$$V_{OUT} = 4(X_1 - X_2)(Y_1 - Y_2) \quad (3)$$

which reduces to $V_{OUT} = 4\,V_G\,V_{IN}$ using the labeling conventions shown in Figure 1. As noted, the phase of the output reverses when V_G is negative. A slightly higher value of R_F is used to compensate for the finite gain of the AD811.

Both the AD811 and the AD834 can operate individually from power-supply voltages of ±5 V. However, to ensure proper operation of the AD811's input stage, the common-mode voltage at W1 and W2 must be within the common-mode range of these inputs. There are several ways to do this. We can use separate supplies of ±5 V for the AD834 and ≥ ±9 V for the AD811. Here, we have chosen to show how the VCA can be biased from one dual supply of nominally ±12 V. Figure 5 also helps to understand the dc biasing design.

We begin by deciding to place the AD834's outputs at about +5 V (a little higher than they operate in the other published applications of this product). Under dc conditions, the high open-loop gain of the op amp forces W1 and W2 to assume the same potential. We calculate the values of R_A to introduce the required 7 V drop, by considering the components of the total current in each of these resistors for the zero-signal condition.

First, when V_{OUT} = 0, the current in resistors R_F must be 10 mA (5 V/ 500 Ω). Second, the standing current into W1 and W2, due to the AD834's internal biasing, is 8.5 mA per side. Third, in this application we provide the positive supply voltage for the AD834 (at Pin 6) via resistors R_B which each carry one-half of the total supply current of 11 mA. Thus, the total current in resistors R_A is 24 mA (10 + 8.5 + 5.5 mA) and a value of 294 Ω is chosen (the closest standard value to 7 V/24 mA) for these resistors. Finally, we choose R_B to set the voltage at Pin 6 to +4 V, which is high enough to ensure accurate operation of the AD834 over the full signal and temperature ranges; the nearest standard resistor value is 182 Ω (1 V/ 5.5 mA).

The presence of these resistors (whose parallel sum is 112 Ω on each side) at the input of the op amp causes it to operate at a "noise gain" of 4.45 (499 Ω/112 Ω), but this neither has any significant effect on the dc scaling of the system, nor does it lower the closed-loop bandwidth (as would be the case for a conventional voltage-feedback op amp).

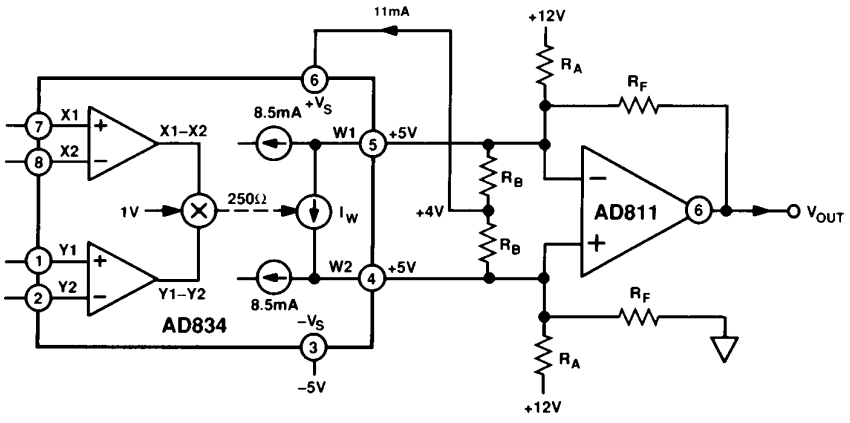


Figure 5. Simplified Schematic of the VCA for Analysis Purposes

A VIDEO KEYER BASED ON THE VCA

Using two AD834s and adding a 1 V dc source, a special form of a two-input VCA called a video keyer (Figure 6) can be assembled. Keying is the term used in reference to blending two or more video sources under the control of a further signal or signals to create such special effects as dissolves and overlays. The circuit described here is a two-input keyer, with video inputs V_A and V_B, and a control input V_G. The output at the load is given by

$$V_{OUT} = GV_A + (1 - G)V_B \quad (4)$$

where G is a dimensionless variable (actually, just the gain of the "A" signal path) that ranges from 0 when $V_G = 0$, to 1 when $V_G = +1$ V. Thus, V_{OUT} varies continuously between V_A and V_B as G varies from 0 to 1.

The operation is straightforward. Consider first the signal path through U1, which handles video input V_A. Its gain is clearly zero when $V_G = 0$ and the scaling we have chosen ensures that it is unity when $V_G = +1$ V; this takes care of the first term in Equation 4. On the other hand, the V_G input to U2 is taken to the *inverting* input X2 while X1 is biased at an accurate +1 V. Thus, when $V_G = 0$, the response to video input V_B is already at its full-scale value of unity, whereas when $V_G = +1$ V, the differential input $X_1 - X_2$ is zero. This generates the second term in Equation 4.

To generate the 1 V dc needed for the "1–G" term, an AD589 reference supplies 1.225 V ±25 mV to a voltage divider consisting of resistors R2 through R4. Potentiometer R3 should be adjusted to provide exactly +1 V at the X1 input.

In this case, we have shown an alternative arrangement using dual supplies of ±5 V for the AD834 and ±12 V for the AD811. Also, the overall gain in this case is arranged to be unity *at the load*, when it is driven from a reverse-terminated 75 Ω line. This means that the "dual VCA" has to operate at a maximum gain of ×2, rather than ×4 as in Figure 1. However, this cannot be achieved by lowering the feedback resistor, since below a critical value (not much less that 500 Ω) the AD811 will become unstable. This is because the dominant pole in the closed-loop ac response of a current-feedback amplifier is controlled by this feedback resistor. It would be possible to operate at a gain of ×4 and then attenuate the signal at the output. Instead, we have chosen to attenuate the signals by 6 dB at the input to the AD811; this is the function of R8 through R11.

The −3dB bandwidth is about 85 MHz and the gain is flat within ±0.1 dB to 30 MHz (Figure 7). Output noise and signal isolation with either channel fully off and the other fully on is about −60 dB to 20 MHz. The feedthrough at 100 MHz is limited primarily by board layout.

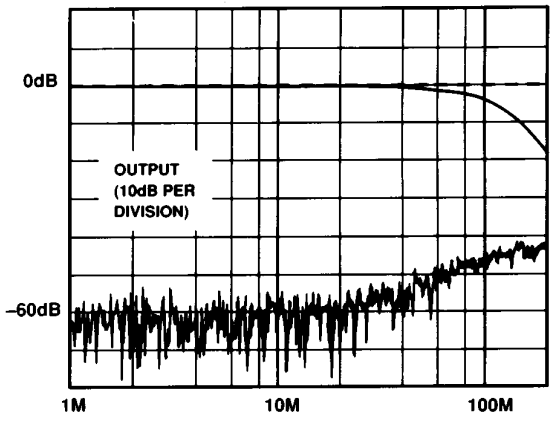


Figure 7. AC Response of the Video Keyer, at V_G = Zero and +1 V; Feedthrough Is About −60 dB

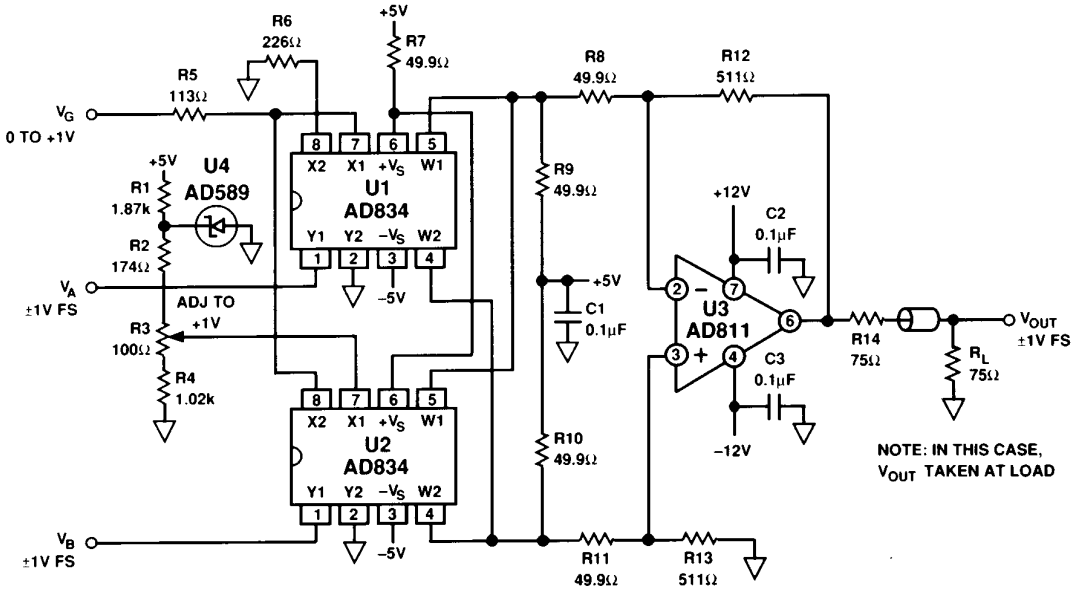


Figure 6. A Two-Input Video Keyer Based on the VCA

A FEEDBACK KEYER

The gain accuracy of the "VCA-based" keyer is dependent on the feedback resistor, R_F. Also, any nonlinearity in the multipliers will show up as a differential gain error. Using an alternative technique, in which the feedback is routed back to unused signal inputs on the AD834s, we can eliminate the feedback resistor and achieve higher accuracy. In the design shown here, we have also used a level-shifting network between the AD834 and the AD811 that eliminates the need for separate power supplies for the two ICs. (In fact, this technique can also be used in the VCAs.)

The basic idea is shown in Figure 8. Note first that V_{OUT} is returned to the inverting inputs Y2 of the multipliers and that their outputs are added. The sum is forced to zero by the assumed high open-loop gain of the op amp. Multiplier M1 produces an output $G(V_A-V_{OUT})$, while M2 produces an output $(1-G)(V_B-V_{OUT})$, where G is $V_G/(1\ V)$ and ranges from 0 to 1. Therefore, the complete system is described by the limiting condition

$$G(V_A-V_{OUT}) + (1-G)(V_B-V_{OUT}) \rightarrow 0 \qquad (5)$$

which requires that

$$V_{OUT} = GV_A + (1-G)V_B \qquad (6)$$

exactly as required for a two-input keyer. The summation of the differential current-mode outputs of the two AD834s is simply achieved by connecting together their respective W1 and W2 nodes. The resulting signal—essentially the loop error represented by the left-hand side of Equation 5—is forced to zero by the high gain of an AD811 op amp.

Figure 9 provides a practical embodiment of these ideas. The gain-control details to provide G and (1–G) terms

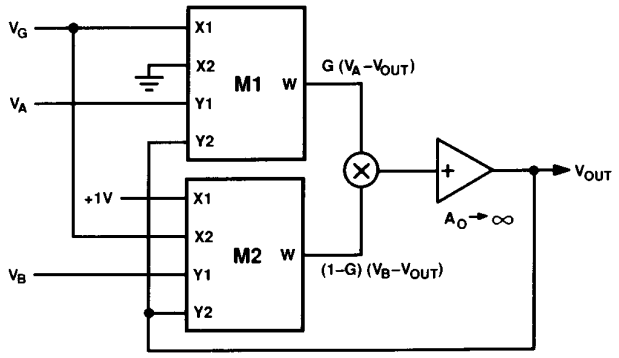


Figure 8. Elements of a Feedback Keyer

are identical to those used previously. The bias currents required at the output of the multipliers are provided by R8 and R9. A dc-level-shifting network comprising R10/R12 and R11/R13 ensures that the input nodes of the AD811 are positioned within an acceptable common-mode range for this IC. At high frequencies, C1 and C2 bypass R10 and R11, respectively.

R14 is included to lower the HF loop gain, and is needed because the voltage-to-current conversion in the AD834s, via the Y2 inputs, results in an effective value of the feedback resistance of 250 Ω (see Figure 5); this is only half the minimum value of 500 Ω required for HF stability of the AD811. (Note that this resistance is unaffected by G: when G = 1, all the feedback is via U1, while when G = 0 it is all via U2.) Resistor R14 reduces the fractional amount of output current from the multipliers into the current-summing inverting input of the AD811, by sharing it with R8. This resistor can be used to adjust the bandwidth and damping factor to best suit the application.

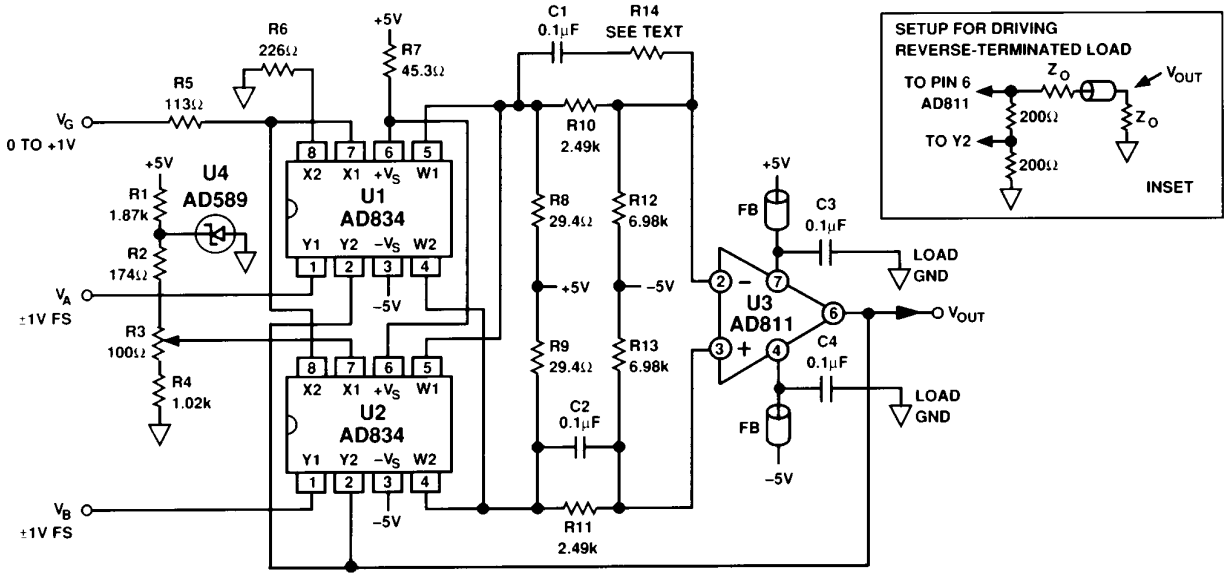


Figure 9. A Practical Embodiment of a Feedback Keyer. The Inset Shows the Feedback Configuration (Gain of ×2) for Driving a Reverse-Terminated Load.

Figure 10 shows the small-signal ac response of this system of the "A" channel at unity gain and zero gain; as is inevitably the case, there is a small amount of feedthrough at the highest frequencies. Two representative values of R14 are shown; using 402 Ω, the pulse response is considerably overdamped, resulting in a −3 dB bandwidth of 15 MHz, while a value of 107 Ω provides a maximally flat response with a −3 dB bandwidth of 70 MHz.

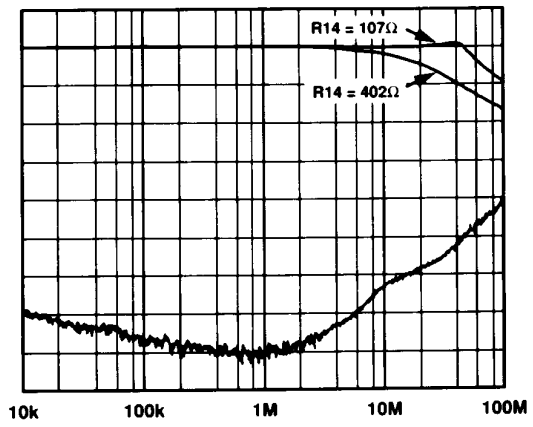
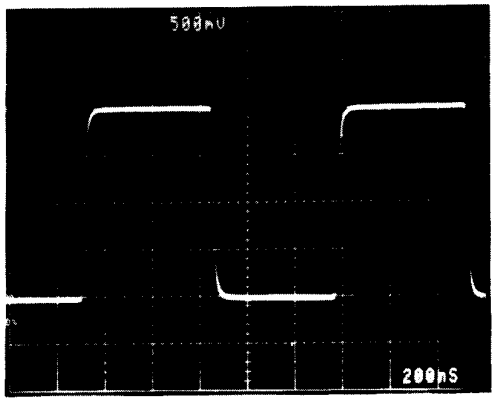
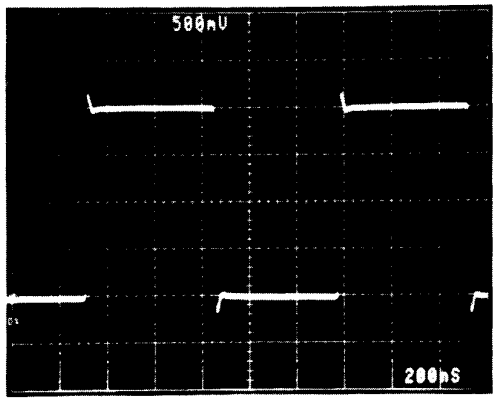


Figure 10. AC Response of the Feedback Keyer. For V_G = +1 V, the −3 dB Bandwidth Is 15 MHz Using R14 = 402 Ω and 70 MHz with R14 = 107 Ω. For These Measurements, R_L = 50 Ω

Figure 11 shows the pulse response at unity gain: in (a) R14 = 402 Ω, while in (b) R14 = 107 Ω. The frequency and pulse responses of the "B" channel, and of the gain-control input are the same, being limited by the output amplifier rather than the AD834s. Likewise, the differential gain and phase behavior will be determined primarily by the AD811; the data sheet should be consulted for more information. The feedthrough at 1 MHz is about −80 dB and −64 dB at 10 MHz and, as before, is eventually limited by board layout. All of these results used a 50 Ω load at the output.



a.



b.

Figure 11. Pulse Response of the Feedback Keyer. In (a), R14 = 402 Ω While in (b), R14 = 107 Ω. For These Measurements, R_L = 50 Ω

Unlike Figure 6's circuit, this keyer provides unity-gain operation. In applications where a reverse-terminated line (50+50 Ω or 75+75 Ω) is to be driven, the gain can be doubled by the inclusion of a resistive divider between V_{OUT} and the Y2 pins; equal resistors of 200 Ω can be used (see the inset in Figure 9). This halving of the feedback voltage also lowers the bandwidth, which can now be restored by reducing, or even eliminating, R14. Figures 12 and 13 show the modified circuit's performance when driving a 50 Ω reverse-terminated line.

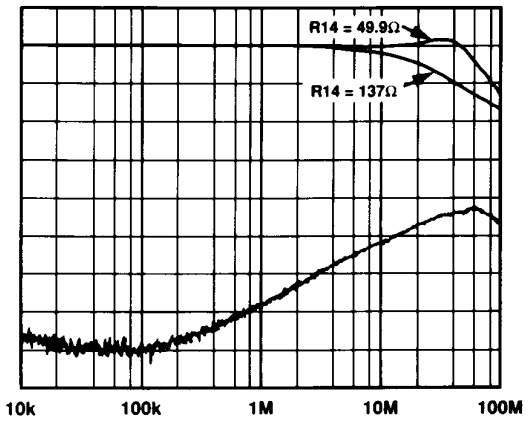
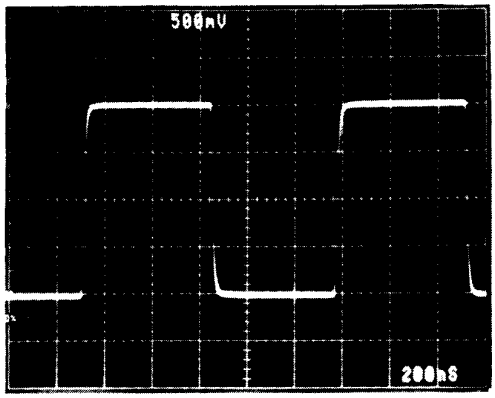
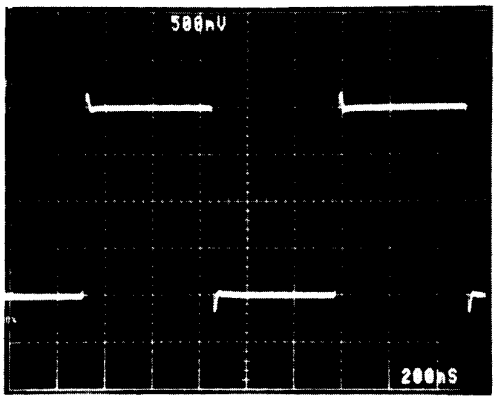


Figure 12. AC Response of the Feedback Keyer, Now Configured for a Gain of ×2. For V_G = + 1 V, the −3 dB Bandwidth Is 15 MHz Using R14 = 137 Ω and 70 MHz with R14 = 49.9 Ω. For These Measurements, R_L = 50 Ω



a.



b.

Figure 13. Pulse Response of the Feedback Keyer Now Configured for a Gain of ×2. In (a), R14 = 137 Ω While in (b), R14 = 49.9 Ω. For These Measurements, R_L = 50 Ω

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