Application Note 57

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Video Circuit Collection

Frank Cox

INTRODUCTION

Even in a time of rapidly advancing digital image processing, analog video signal processing still remains eminently viable. The video A/D converters need a supply of properly amplified, limited, DC restored, clamped, clipped, contoured, multiplexed, faded and filtered analog video before they can accomplish anything. After the digital magic is performed, there is usually more amplifying and filtering to do as an adjunct to the D/A conversion process, not to mention all those pesky cables to drive. The analog way is often the most expedient and efficient, and you don't have to write all that code.

The foregoing is only partly in jest. The experienced engineer will use whatever method will properly get the job done; analog, digital or magic (more realistically, a combination of all three). Presented here is a collection of analog video circuits that have proven themselves useful.

CIRCUIT INDEX

VIDEO AMPLIFIER SELECTION GUIDE

PART	GBW (MHz)	CONFIGURATION	COMMENTS
LT1217	10	S	CFA, I_S = 1mA, Shutdown
LT1200/LT1201/ LT1202	11	S, D, Q	I_S = 1mA per Amp, Good DC Specs
LT1355/LT1356	12	D, Q	400V/µs SR, I_S = 1.25mA per Amp, Good DC Specs
LT1211/LT1212	14	D, Q	Single Supply, Excellent DC Specs
LT1215/LT1216	23	D, Q	Single Supply, Excellent DC Specs
LT1358/LT1359	25	D, Q	600V/µs SR, I_S = 2.5mA per Amp, Good DC Specs
LT1213/LT1214	28	D, Q	Single Supply, Excellent DC Specs
LT1208/LT1209	45	D, Q	400V/µs SR
LT1220	45	S	250V/µs, Good DC Specs, 12-Bit Accurate
LT1224	45	S	400V/µs SR
LT1190	50	S	Low Voltage
LT1360	50	S	800V/µs SR, I_S = 5mA per Amp, Good DC Specs
LT1206	60	S	250mA Output Current CFA, 600V/µs SR, Shutdown
LT1363	70	S	1000V/µs SR, I_S = 7.5mA per Amp, Good DC Specs
LT1204	70	Q	CFA, 4-Input Video MUX Amp, 1000V/µs SR, Superior Isolation
LT1228	80 (g_m = 0.25)	S	Transconductance Amp + CFA, Extremely Versatile
LT1191	90	S	Low Voltage, ±50mA Output
LT1229/LT1230	100	D, Q	CFA, 1000V/µs SR, DG = 0.04%, DP = 0.1°
LT1223	100	S	CFA, 12-Bit Accurate, Shutdown, 1300V/µs SR, Good DC Specs, DG = 0.02%, DP = 0.12°
LT1252	100	S	CFA, DG = 0.01%, DP = 0.09°, Low Cost
LT1253	117	D, Q	CFA, DG = 0.03%, DP = 0.28°, Flat to 30MHz, 0.1dB
LT1259/LT1260	130	D, T	RGB CFA, 0.1dB Flat to 30MHz, DG = 0.016%, DP = 0.075°, Shutdown
LT1227	140	S	CFA, 1100V/µs SR, DG = 0.01%, DP = 0.01°, Shutdown
LT1221	150 (A_V = 4)	S	250V/µs SR, 12-Bit Accurate
LT1193	160 (A_V = 2)	S	Low Voltage, Differential Input, Adjustable Gain, ±50mA Output
LT1203/LT1205	170	D, Q	MUX, 25ns Switching, DG = 0.02%, DP = 0.04°
LT1192	350 (A_V = 5)	S	Low Voltage, ±50mA Output
LT1194	350 (A_V = 10)	S	Differential Input, Low Voltage, Fixed Gain of 10
LT1222	500 (A_V = 10)	S	12-Bit Accurate
LT1226	1000 (A_V = 25)	S	400V/µs SR, Good DC Specs

Key to Abbreviations:

CFA = Current Feedback Amplifier
DG = Differential Gain
DP = Differential Phase
MUX = Multiplexer
SR = Slew Rate

S = Single
D = Dual
Q = Quad
T = Triple

Note:

Differential gain and phase is measured with a 150Ω load, except for the LT1203/LT1205 in which case the load is 1000Ω.

VIDEO PROCESSING CIRCUITS

Variable Gain Amplifier Has ±3dB Range While Maintaining Good Differential Gain and Phase

The circuit in Figure 1 is a variable gain amp suitable for composite video use. Feedback around the transconductance amp (LT1228) acts to reduce the differential input voltage at the amplifier's input, and this reduces the differential gain and phase errors. Table 1 shows the differential phase and gain for three gains. Signal-to-noise ratio is better than 60dB for all gains.

Table 1.

INPUT (V)	I_{SET} (mA)	DIFFERENTIAL GAIN	DIFFERENTIAL PHASE
0.707	4.05	0.4%	0.15°
1.0	1.51	0.4%	0.1°
1.414	0.81	0.7%	0.5°

Black Clamp

Here is a circuit that removes the sync component of the video signal with minimal disturbance to the luminance (picture information) component. It is based on the classic op amp half-wave-rectifier with the addition of a few refinements.

The classic "diode-in-the-feedback-loop" half-wave-rectifier circuit generally does not work well with video frequency signals. As the input signal swings through zero volts, one of the diodes turns on while the other is turned off, hence the op amp must slew through two diode drops. During this time the amplifier is in slew limit and the output signal is distorted. It is not possible to entirely prevent this source of error because there will always be some time when the amp will be open-loop (slewing) as the diodes are switched, but the circuit shown here in Figure 2 minimizes the error by careful design.

The following techniques are critical in the design shown in Figure 2:

1. The use of diodes with a low forward voltage drop reduces the voltage that the amp must slew.

2. Diodes with a low junction capacitance reduce the capacitive load on the op amp. Schottky diodes are a good choice here as they have both low forward voltage and low junction capacitance.

3. A fast slewing op amp with good output drive is essential. An excellent CFA like the LT1227 is mandatory for good results.

4. Take some gain. The error contribution of the diode switch tends to be constant, so a larger signal means a smaller percentage error.

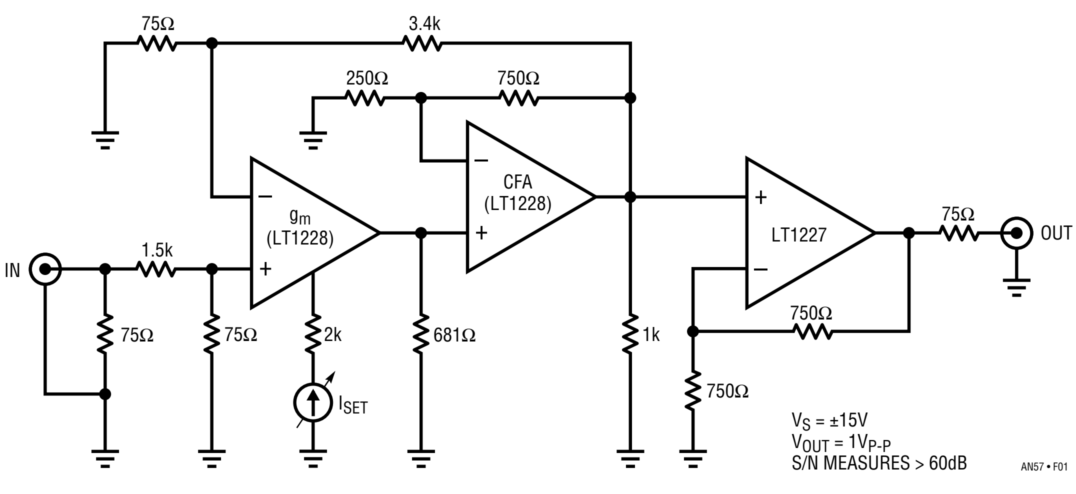


Figure 1. ±3dB Variable Gain Video Amp Optimized for Differential Gain and Phase

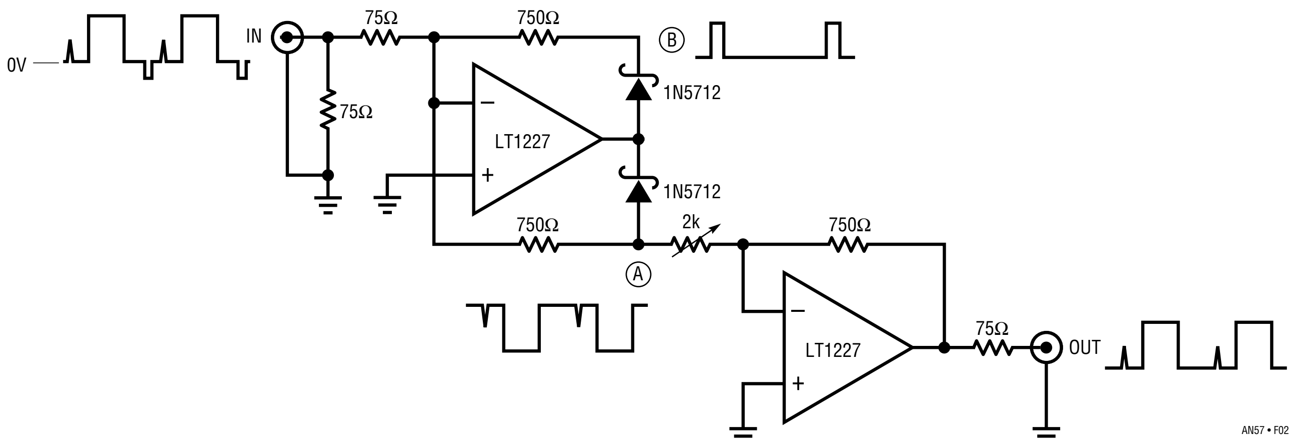


Figure 2. Black Clamp Circuit

Since this circuit discriminates between the sync and video on the basis of polarity, it is necessary to have an input video signal that has been DC restored (the average DC level is automatically adjusted to bring the blanking level to zero volts). Notice that not only is the positive polarity information (luminance: point A in the schematic) available, but that the negative polarity information (sync: point B in the schematic) is also. Circuits that perform this function are called "black clamps." The photograph (Figure 3) shows the circuit's clean response to a 1T[1] pulse (some extra delay is added between the input and output for clarity).

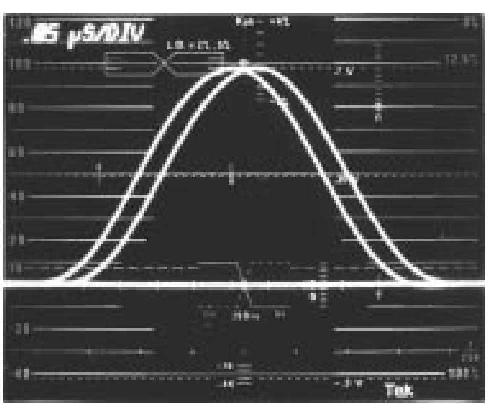


Figure 3. Black Clamp Circuit Response to a "1T" Pulse (±15V Power Supplies)

Video Limiter

Often there is a need to limit the amplitude excursions of the video signal. This is done to avoid exceeding luminance reference levels of the video standard being used, or to avoid exceeding the input range of another processing stage such as an A/D converter. The signal can be hard limited in the positive direction, a process called "white peak clipping," but this destroys any amplitude information and hence any scene detail in this region. A more gradual limiting ("soft limiter") or compression of the peak white excursion is performed by elements called "knee" circuits, after the shape of the amplifier transfer curve.

A soft limiter circuit is shown in Figure 4 which uses the LT1228 transconductance amp. The level at which the limiting action begins is adjusted by varying the set current into pin 5 of the transconductance amplifier. The LT1228 is used here in a slightly unusual, closed-loop configuration. The closed-loop gain is set by the feedback and gain resistors (R_F and R_G) and the open-loop gain by the transconductance of the first stage times the gain of the CFA.

[1] A 1T pulse is a specialized video waveform whose salient characteristic is a carefully controlled bandwidth which is used to quickly quantify gain and phase flatness in video systems. Phase shift and/or gain variations in the video system's passband result in transient distortions which are very noticeable on this waveform (not to mention the picture). [For you video experts out there, the K factor was 0.4% (the TEK TSG120 video signal generator has a K factor of 0.3%)].

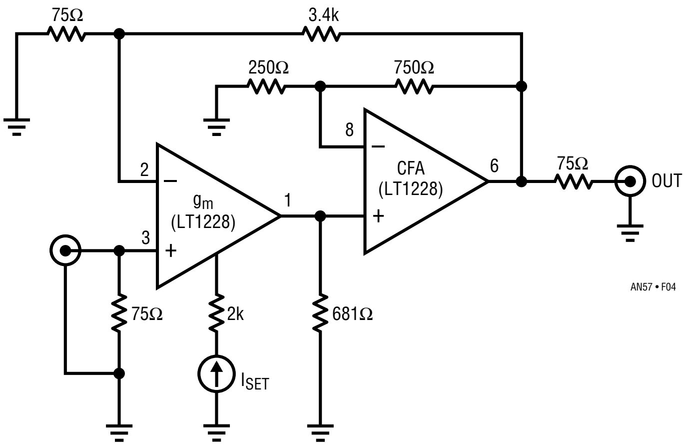


Figure 4. LT1228 Soft Limiter

As the transconductance is reduced (by reducing the set current), the open-loop gain is reduced below that which can support the closed-loop gain and the amp limits. A family of curves which show the response of the limiting amplifier subject to different values of set current with a ramp input is shown in Figure 5. Figure 6 shows the change in limiting level as I_{SET} is varied.

Circuit for Gamma Correction

Video systems use transducers to convert light to an electric signal. This conversion occurs, for example, when a camera scans an image. Video systems also use transducers to convert the video signal back to light when the signal is sent to a display, a CRT monitor for example. Transducers often have a transfer function (the ratio of *signal in to light out*) that is unacceptably nonlinear.

The newer generation of camera transducers (CCDs and the improved versions of vidicon-like tubes) are adequately linear, however, picture monitor CRTs are not. The transfer functions of most CRTs follow a power law. The following equation shows this relation:

$$\text{Light Out} = k \bullet V_{SIG}^{\gamma}$$

where k is a constant of proportionality and gamma (γ) is the exponent of the power law (gamma ranges from 2.0 to 2.4).

This deviation from nonlinearity is usually called just *gamma* and is reported as the exponent of the power law. For instance, “the gamma of this vidicon is 0.43.” The correction of this effect is *gamma correction*.

In the equation above, notice that a gamma value of 1 results in a linear transfer function. The typical CRT will have a transfer function with a gamma from about 2.0 to 2.4. Such values of gamma give a nonlinear response which compresses the blacks and stretches the whites. Cameras usually contain a circuit to correct this nonlinearity. Such a circuit is a *gamma corrector* or simply a *gamma circuit*.

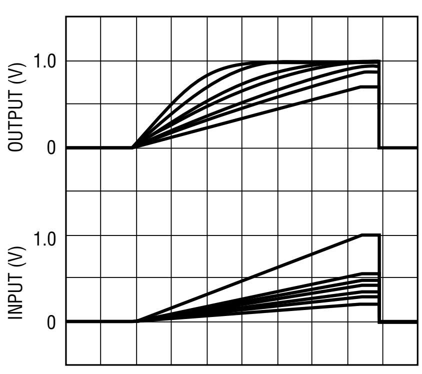


Figure 5. Output of the Limiting Amp (I_{SET} = 0.68mA), with a Ramp Input. As the Input Amplitude Increases from 0.25V to 1V, the Output is Limited to 1V

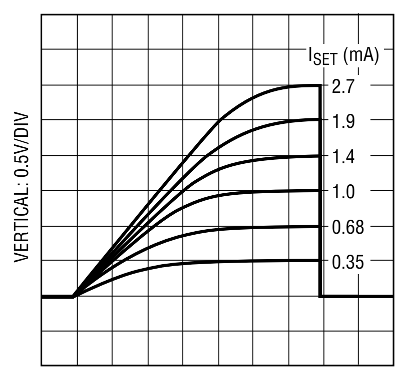


Figure 6. The Output of the Limiting Amp with Various Limiting Levels (I_{SET}). The Input is a Ramp with a Maximum Amplitude of 0.75V

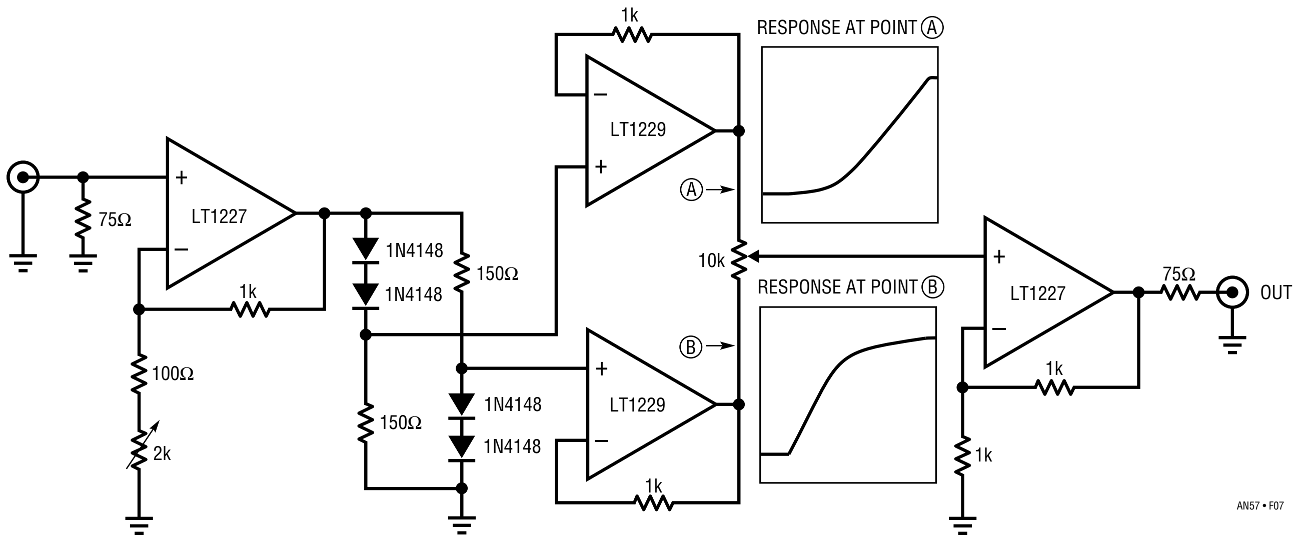


Figure 7. Gamma Amp (Input Video Should Be Clamped)

Figure 7 shows a schematic of a typical circuit which can correct for positive or negative gamma. This is an upgrade of a classic circuit which uses diodes as the nonlinear elements. The temperature variation of the diode junction voltages is compensated to the first order by the balanced arrangement. LT1227s and LT1229s were used in the prototype, but a quad (LT1230) could save some space and work as well.

Figure 8 shows a response curve (transfer function) for an uncorrected CRT. To make such a response linear, the gamma corrector must have a gamma that is the reciprocal of the gamma of the device being linearized. Figure 8, curve A shows a response curve (transfer function) for an uncorrected CRT. The response of a two diode gamma circuit like that in Figure 7 is shown in Figure 8, curve B. Summing these two curves together, as in Figure 8, curve C, demonstrates the action of the gamma corrector. A straight line of appropriate slope, which would be an ideal response, is shown for comparison in Figure 8, curve D. Figure 9 is a triple exposure photograph of the gamma corrector circuit adjusted for gammas of –3, 1 and +3 (approximately). The input is a linear ramp of duration 52µs which is the period of an active horizontal line in NTSC video.

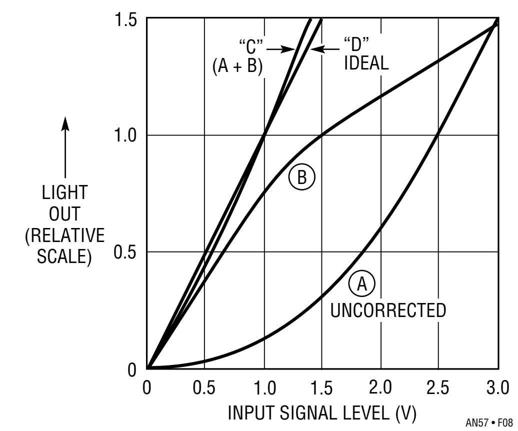


Figure 8. Uncorrected CRT Transfer Function

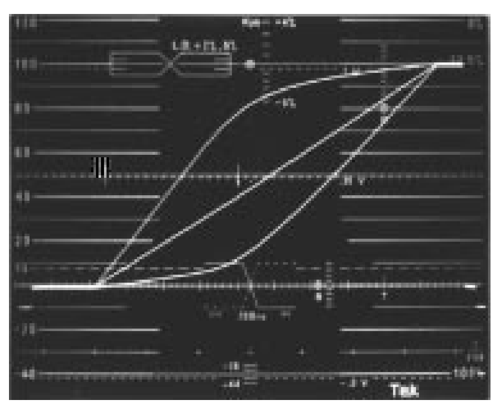


Figure 9. Gamma Corrector Circuit Adjusted for Three Gammas: –3, 1, +3 (Approximately). The Input is a Linear Ramp

LT1228 Sync Summer

The circuit shown in Figure 10a restores the DC level and adds sync to a video waveform. For this example the video source is a high speed DAC with an output which is referenced to –1.2V. The LT1228 circuit (see the LT1228 data sheet for more details) forms a DC restore[2] that maintains a zero volt DC reference for the video. Figure 10b shows the waveform from the DAC, the DC restore pulse, and composite sync. The LT1363 circuit sums the video and composite sync signals. The 74AC04 CMOS inverters are used to buffer the TTL composite sync signal. In addition they drive the shaping network and, as they are mounted on the same ground plane as the analog circuitry, they isolate the ground noise from the digital system used to generate the video timing signals. Since the sync is directly summed to the video, any ground bounce or noise gets added in too. The shaping network is simply a third order Bessel lowpass filter with a bandwidth of 5MHz and an impedance of 300Ω. This circuitry slows the edge rate of the digital composite sync signal and also attenuates the noise. The same network, rescaled to an impedance of 75Ω, is used on the output of the summing amp to attenuate the switching noise from the DAC and to remove some of the high frequency components of the waveform. A more selective filter is not used here as the DAC has low glitch energy to start with and the signal does not have to meet stringent bandwidth requirements. The LT1363 used for the summing amp has excellent transient characteristics with no overshoot or ringing. Figure 10c shows two

[2] This is also referred to as "DC clamp" (or just clamp) but, there is a distinction. Both clamps and DC restore circuits act to maintain the proper DC level in a video signal by forcing the blanking level to be either zero volts or some other appropriate value. This is necessary because the video signal is often AC coupled as in a tape recorder or a transmitter. The DC level of an AC coupled video signal will vary with scene content and therefore the black referenced level must be "restored" in order for the picture to look right. A clamp is differentiated from a DC restore by its speed of response. A clamp is faster, generally correcting the DC error in one horizontal line (63.5µs for NTSC). A DC restore responds slower, more on the order of the frame time (16.7ms for NTSC). If there is any noise on the video signal the DC restore is the preferred method. A clamp can respond to noise pulses that occur during the blanking period and as a result give an erroneous black level for the line. Enough noise causes the picture to have an objectionable distortion called "piano keying." The black reference level and hence the luminance level change from line to line.

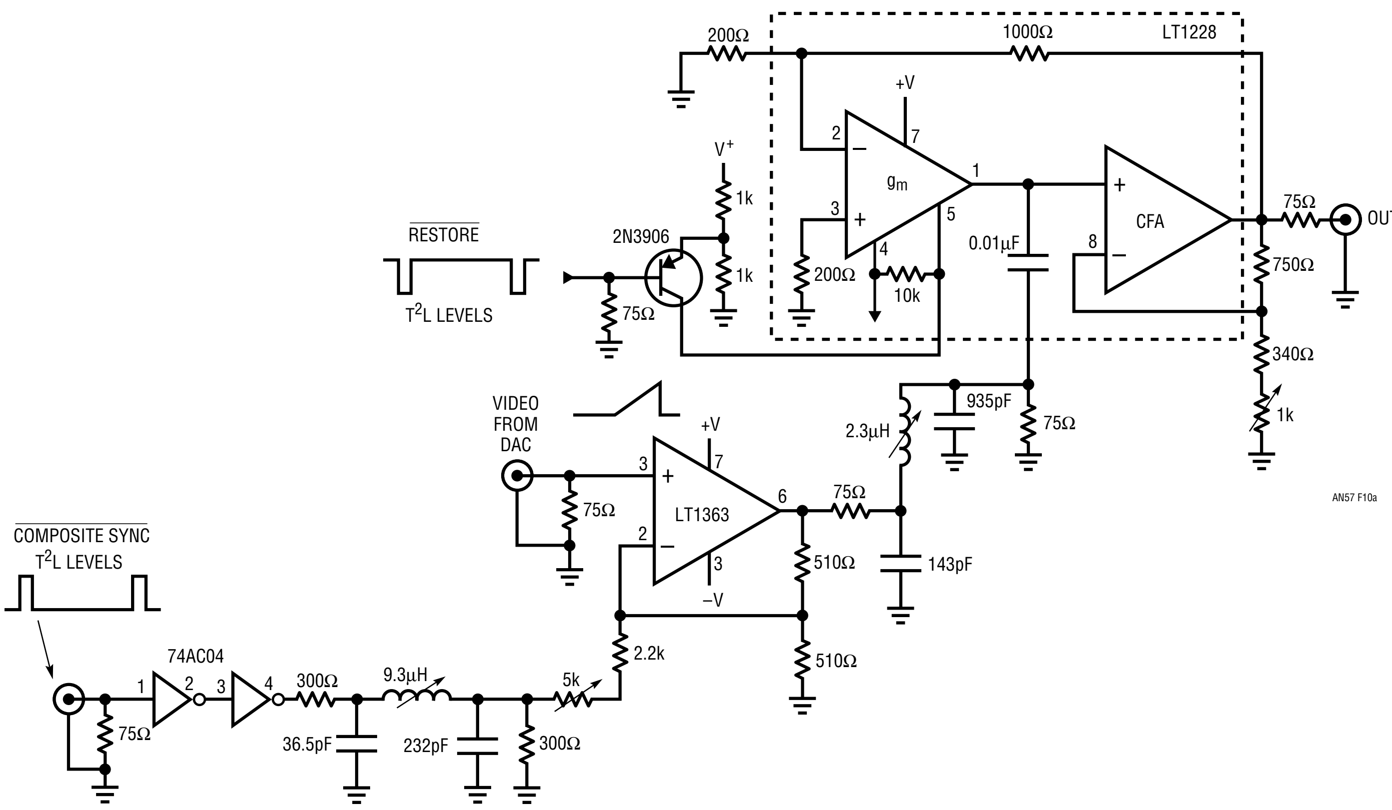


Figure 10a. Simple Sync Summer

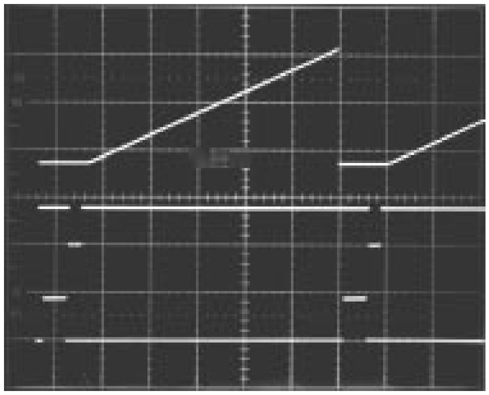


Figure 10b. Video Waveform from DAC; Clamp Pulse and Sync Pulse Used as Inputs to Sync Summer

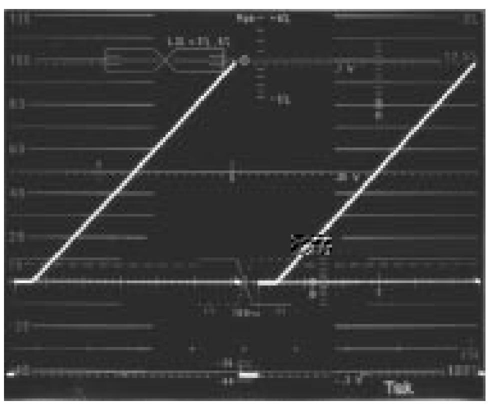


Figure 10c. Reconstructed Video Out of Sync Summer

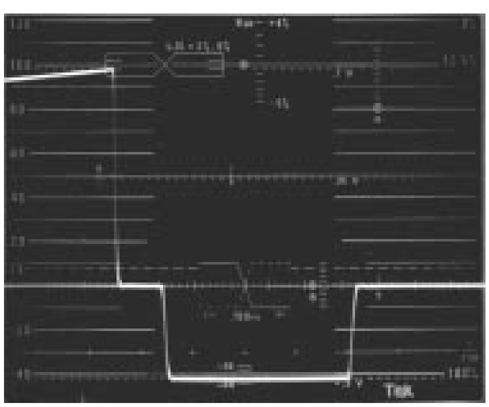


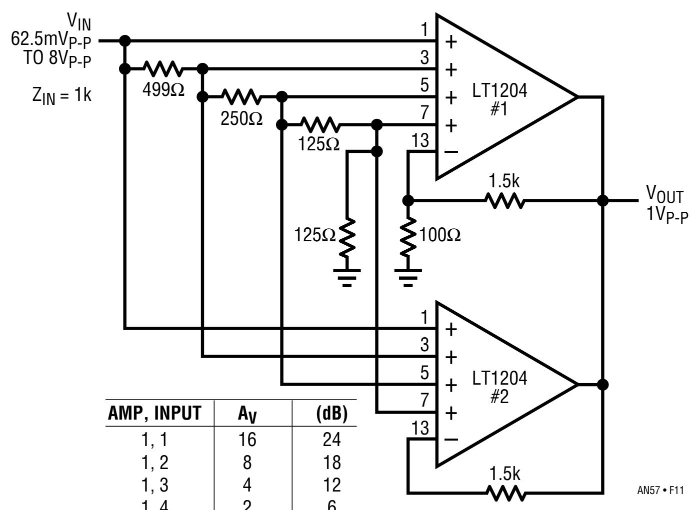
Figure 10d. Close-Up of Figure 10c, Showing Sync Pulse

horizontal lines of the output waveform with the DC restored and the sync added. Figure 10d is an expanded view of the banking interval showing a clean, well formed sync pulse.

LT1204 70MHz MULTIPLEXER CIRCUITS

Stepped Gain Amp Using the LT1204

This is a straightforward approach to a switched-gain amp that features versatility. Figures 11 and 12 show circuits which implement a switched-gain amplifier; Figure 11 features an input Z of 1000Ω, while Figure 12's input Z is 75Ω. In either circuit, when LT1204 amp/MUX #2 is selected the signal is gained by one, or is attenuated by the resistor divider string depending on the input selected. When LT1204 amp/MUX #1 is selected there is an additional gain of sixteen. Consult the table in Figure 11. The gain steps can be either larger or smaller than shown here. The input impedance (the sum of the divider resistors) is also arbitrary. Exercise caution in taking large gains however, because the bandwidth will change as the output is switched from one amp to another. Taking more gain in the amp/MUX #1 will lower its bandwidth even though it is a current feedback amplifier (CFA). This is less true for a CFA than for a voltage feedback amp.



AMP, INPUT	A_V	(dB)
1, 1	16	24
1, 2	8	18
1, 3	4	12
1, 4	2	6
2, 1	1	0
2, 2	0.5	–6
2, 3	0.25	–12
2, 4	0.125	–18

Figure 11. Switchable Gain Amplifier Accepts Inputs from $62.5mV_{P-P}$ to $8V_{P-P}$

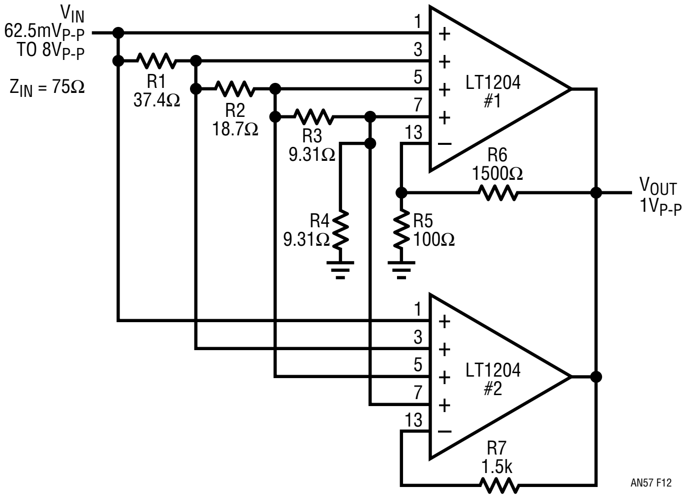


Figure 12. Switchable Gain Amplifier, Z_{IN} = 75Ω Same Gains as Figure 11

LT1204 Amplifier/Multiplexer Sends Video Over Long Twisted Pair

Figure 13 is a circuit which can transmit baseband video over more than 1000 feet of very inexpensive twisted-pair wire and allow the selection of one-of-four inputs. Amp/MUX A1 (LT1204) and A2 (LT1227) form a single differential driver. A3 is a variable gain differential receiver built using the LT1193. The rather elaborate equalization (highlighted on the schematic) is necessary here as the twisted pair goes self-resonant at about 3.8MHz.

Figure 14 shows the video test signal before and after transmission but without equalization. Figure 15 shows before and after with the equalization connected. Differential gain and phase are about 1% and 1°, respectively.

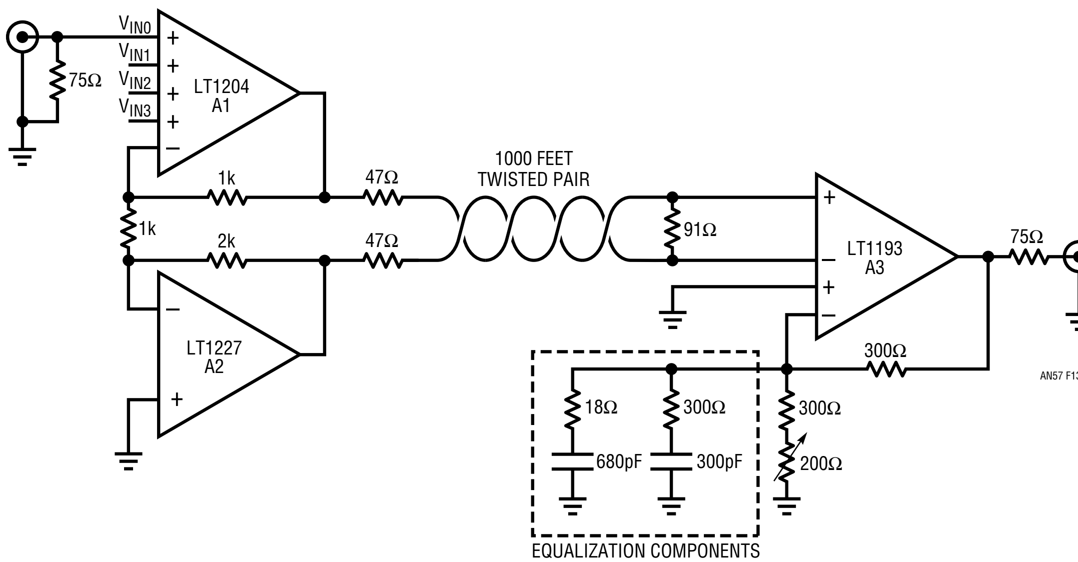


Figure 13. Twisted Pair Driver/Receiver

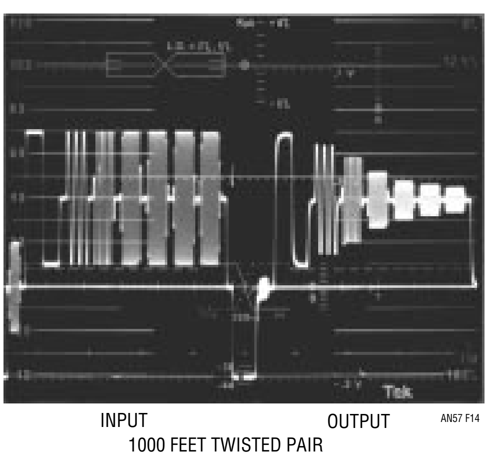


Figure 14. Multiburst Pattern Without Cable Compensation

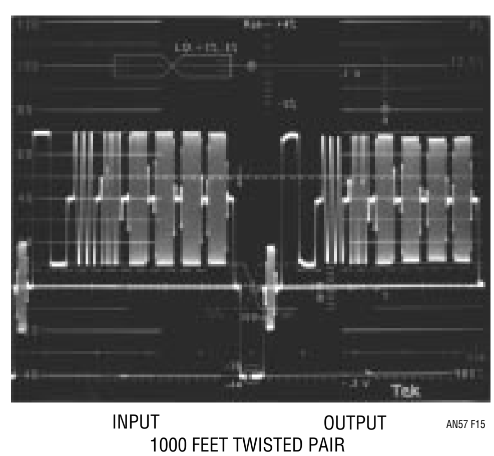


Figure 15. Multiburst Pattern With Cable Compensation

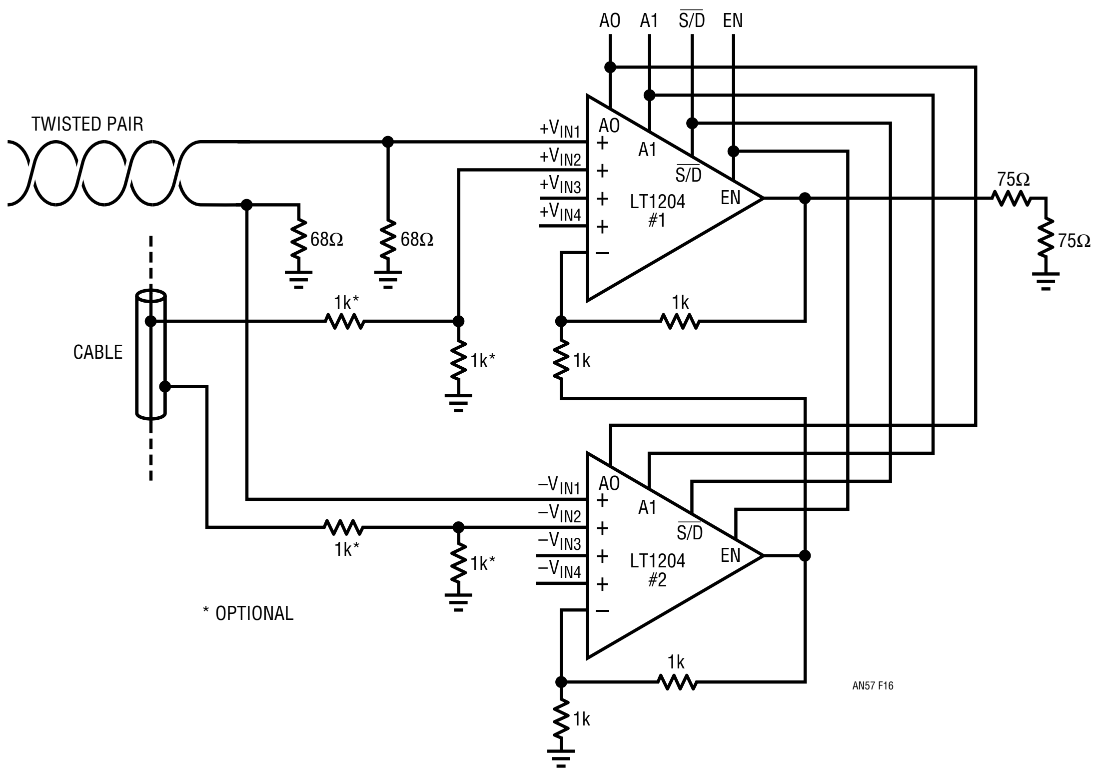


Figure 16. Fast Differential Multiplexer

Fast Differential Multiplexer

This circuit (Figure 16) takes advantage of the gain node on the LT1204 to make a high speed differential MUX for receiving analog signals over twisted pair. Common-mode noise on loop-through connections is reduced because of the unique differential input. Figure 15's circuit also makes a robust differential to single-ended amp/MUX for high speed data acquisition.

Signals passing through LT1204 #1 see a noninverting gain of two. Signals passing through LT1204 #2 also see a noninverting gain of two and then an inverting gain of one (for a resultant gain of minus two) because this amp drives the gain resistor on amp #1. The result is differential amplification of the input signal.

The optional resistors on the second input are for input protection. Figure 17 shows the differential mode response versus frequency. The limit to the response (at low frequency) is the matching of the gain resistors. One percent resistors will match to about 0.1% (60dB) if they are from the same batch.

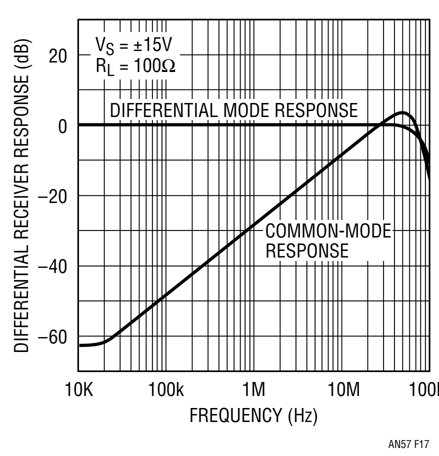


Figure 17. Differential Receiver Response vs Frequency

MISAPPLICATIONS OF CFAs

In general the current feedback amplifier (CFA) is remarkably docile and easy to use. These amplifiers feature "real," usable gain to 100MHz and beyond, low power consumption and an amazingly low price. However, CFAs are still new enough so that there is room for breadboard adventure. Consult the diagrams and the following list for some of the pitfalls that have come to my attention[3].

1. Be sure there is a DC path to ground on the noninverting input pin. There is a transistor in the input that needs some bias current.

2. Don't use pure reactances for a feedback element. This is one sure way to get the CFA to oscillate. Consult the amplifier data sheet for guidance on feedback resistor values. Remember that these values have a direct effect on the bandwidth. If you wish to tailor frequency response with reactive networks, put them in place of R_G, the gain setting resistor.

3. Need a noninverting buffer? Use a feedback resistor!

4. Any resistance between the inverting terminal and the feedback node causes loss of bandwidth.

5. For good dynamic response, avoid parasitic capacitance on the inverting input.

6. Don't use a high Q inductor for power supply decoupling (or even a middling Q inductor for that matter). The inductor and the bypass capacitors form a tank circuit, which can be excited by the AC power supply currents, causing just the opposite of the desired effect. A lossy ferrite choke can be a very effective way to decouple power supply leads without the voltage drop of a series resistor. For more information on ferrites call Fair Rite Products Corp. (914) 895-2055.

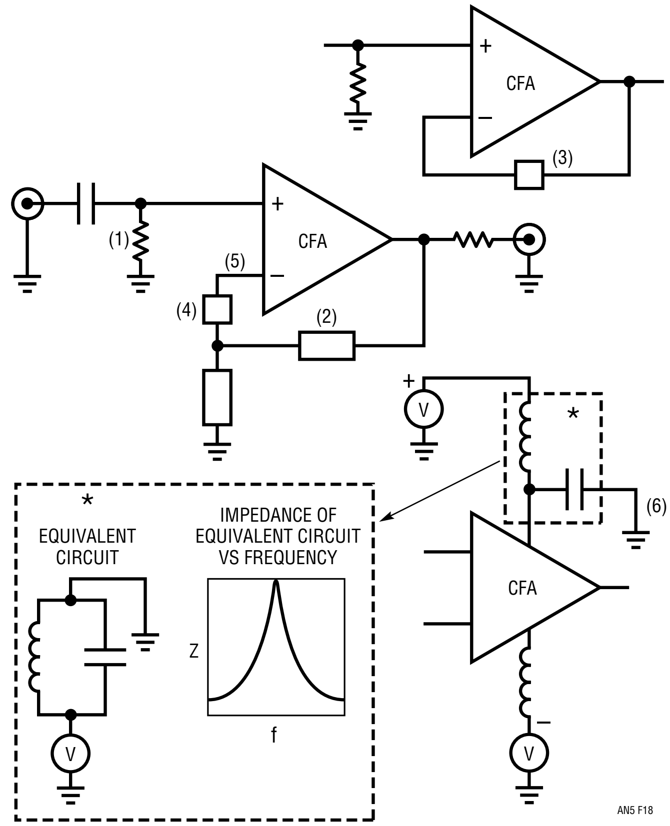


Figure 18. Examples of Misapplications

[3] All the usual rules for any high speed circuit still apply, of course.
A partial list:
a. Use a ground plane.
b. Use good RF bypass techniques. Capacitors used should have short leads, high self-resonant frequency, and be placed close to the pin.
c. Keep values of resistors low to minimize the effects of parasitics. Make sure the amplifier can drive the chosen low impedance.
d. Use transmission lines (coax, twisted pair) to run signals more than a few inches.
e. Terminate the transmission lines (back terminate the lines if you can).
f. Use resistors that are still resistors at 100MHz.
Refer to AN47 for a discussion of these topics.

APPENDICES — VIDEO CIRCUITS FROM LINEAR TECHNOLOGY MAGAZINE

APPENDIX A

A Temperature-Compensated, Voltage-Controlled Gain Amplifier Using the LT1228

It is often convenient to control the gain of a video or intermediate frequency (IF) circuit with a voltage. The LT1228, along with a suitable voltage-to-current converter circuit, forms a versatile gain-control building block ideal for many of these applications.

In addition to gain control over video bandwidths, this circuit can add a differential input and has sufficient output drive for 50Ω systems.

The transconductance of the LT1228 is inversely proportional to absolute temperature at a rate of –0.33%/°C. For circuits using closed-loop gain control (i.e., IF or video automatic gain control) this temperature coefficient does not present a problem. However, open-loop gain-control circuits that require accurate gains may require some compensation. The circuit described here uses a simple thermistor network in the voltage-to-current converter to achieve this compensation. Table A1 summarizes the circuit's performance.

Figure A1 shows the complete schematic of the gain-control amplifier. Please note that these component choices are not the only ones that will work nor are they necessarily the best. This circuit is intended to demonstrate one approach out of many for this very versatile part and, as always, the designer's engineering judgment must be fully engaged. Selection of the values for the input attenuator, gain-set resistor, and current feedback amplifier resistors is relatively straightforward, although some iteration is usually necessary. For the best bandwidth, remember to keep the gain-set resistor R1 as small as possible and the set current as large as possible with due regard for gain compression. See the "Voltage-Controlled Current Source" (I_{SET}) box for details.

Table A1. Characteristics of Example

Input Signal Range	0.5V to 3.0V_{PK}
Desired Output Voltage	1.0V_{PK}
Frequency Range	0Hz to 5MHz
Operating Temperature Range	0°C to 50°C
Supply Voltages	±15V
Output Load	150Ω (75Ω + 75Ω)
Control Voltage vs Gain Relationship	0V to 5V Min to Max Gain
Gain Variation Over Temperature	±3% from Gain at 25°C

Several of these circuits have been built and tested using various gain options and different thermistor values. Test results for one of these circuits are shown in Figure A2. The gain error versus temperature for this circuit is well within the limit of ±3%. Compensation over a much wider

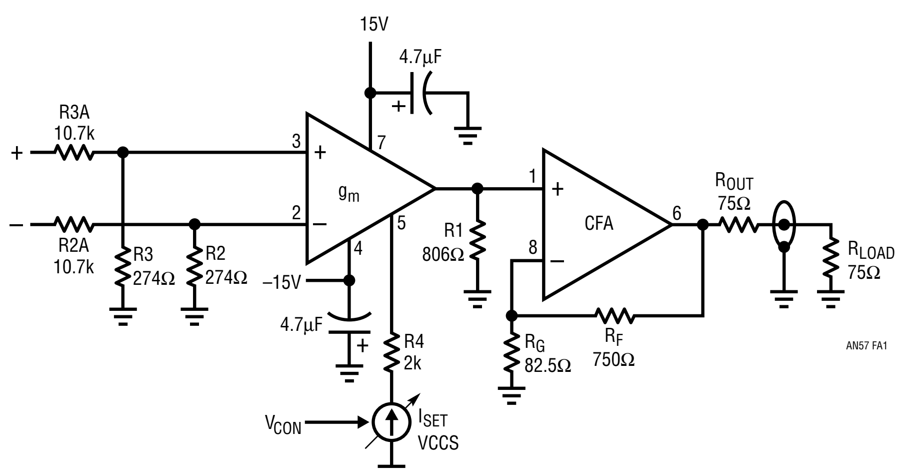


Figure A1. Differential-Input, Variable-Gain Amplifier

range of temperatures or to tighter tolerances is possible, but would generally require more sophisticated methods, such as multiple thermistor networks.

The VCCS is a standard circuit with the exception of the current-set resistor R5, which is made to have a temperature coefficient of –0.33%/°C. R6 sets the overall gain and is made adjustable to trim out the initial tolerance in the LT1228 gain characteristic. A resistor (R_P) in parallel with the thermistor will tend, over a relatively small range, to linearize the change in resistance of the combination with temperature. R_S trims the temperature coefficient of the network to the desired value.

This procedure was performed using a variety of thermistors. BetaTHERM Corporation is one possible source, phone 508-842-0516. Figure A3 shows typical results reported as errors normalized to a resistance with a –0.33%/°C temperature coefficient. As a practical matter, the thermistor need only have about a 10% tolerance for this gain accuracy. The sensitivity of the gain accuracy to the thermistor tolerance is decreased by the linearization network in the same ratio as is the temperature coefficient. The room temperature gain may be trimmed with R6. Of course, particular applications require analysis of aging stability, interchangeability, package style, cost, and the contributions of the tolerances of the other components in the circuit.

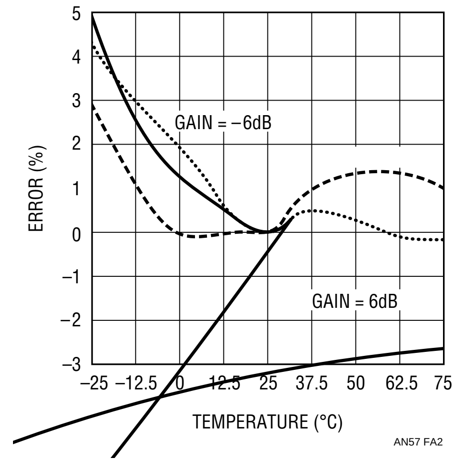


Figure A2. Gain Error for Circuit in Figure 18 Plus Temperature Compensation Circuit Shown in Figure 20 (Normalized to Gain at 25°C)

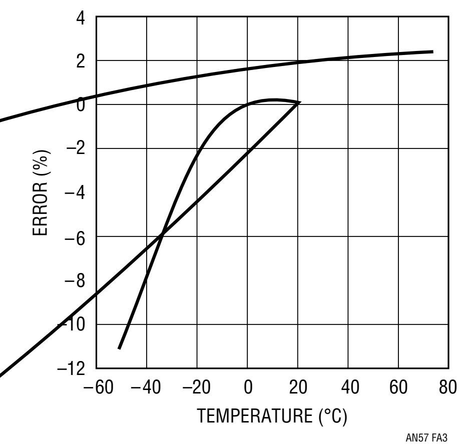


Figure A3. Thermistor Network Resistance Normalized to a Resistor with Exact –0.33%/°C Temperature Coefficient

Voltage-Controlled Current Source (VCCS) with a Compensating Temperature Coefficient

VCCS Design Steps

1. Measure, or obtain from the data sheet, the thermistor resistance at three equally spaced temperatures (in this case 0°C, 25°C, and 50°C). Find R_P from:

$$R_P = \frac{\left(R0 \times R25 + R25 \times R50 - 2 \times R0 \times R50\right)}{\left(R0 + R50 - 2 \times R25\right)}$$

where R0 = thermistor resistance at 0°C
R25 = thermistor resistance at 25°C
R50 = thermistor resistance at 50°C

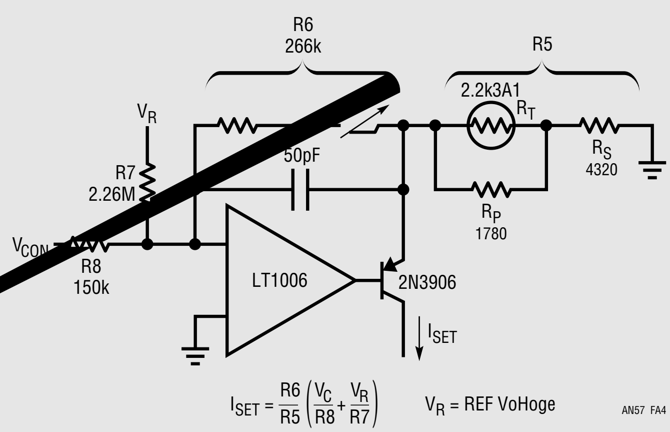


Figure A4. Voltage-Controlled Current Source (VCCS) with a Compensating Temperature Coefficient

2. Resistor R_P is placed in parallel with the thermistor. This network has a temperature dependence that is approximately linear over the range given (0°C to 50°C).

3. The parallel combination of the thermistor and R_P ($R_P||R_T$) has a temperature coefficient (TC) of resistance given by:

$$\text{TC of } R_P \,||\, R_T = \left(\frac{R0 \,||\, R_P - R50 \,||\, R_P}{R25 \,||\, R_P}\right)\left(\frac{100}{T_{HIGH} - T_{LOW}}\right)$$

4. The desired tempco to compensate the LT1228 gain temperature dependence is –0.33%/°C. A series resistance (R_S) is added to the parallel network to trim its tempco to the proper value. R_S is given by:

$$\frac{\left(\text{TC of } R_P \,||\, R_T\right)}{-0.33} \times \left(R_P \,||\, R_{25}\right) - \left(R_P \,||\, R_{25}\right)$$

5. R6 contributes to the resultant TC and so is made large with respect to R5.

6. The other resistors are calculated to give the desired range of I_{SET}.

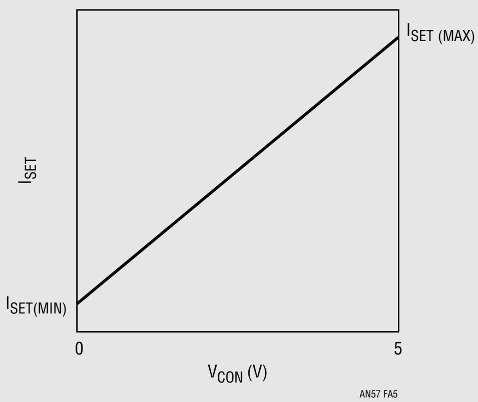


Figure A5. Voltage Control of I_{SET} with Temperature Compensation

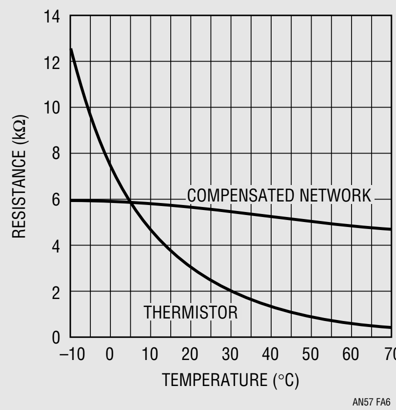


Figure A6. Thermistor and Thermistor Network Resistance vs Temperature

APPENDIX B

Optimizing a Video Gain-Control Stage Using the LT1228

Video automatic-gain-control (AGC) systems require a voltage- or current-controlled gain element. The performance of this gain-control element is often a limiting factor in the overall performance of the AGC loop. The gain element is subject to several, often conflicting restraints. This is especially true of AGC for composite color video systems, such as NTSC, which have exacting phase- and gain-distortion requirements. To preserve the best possible signal-to-noise ratio (S/N),[1] it is desirable for the input signal level to be as large as practical. Obviously, the larger the input signal the less the S/N will be degraded by the noise contribution of the gain-control stage. On the other hand, the gain-control element is subject to dynamic range constraints; exceeding these will result in rising levels of distortion.

Linear Technology makes a high speed transconductance (g_m) amplifier, the LT1228, which can be used as a quality, inexpensive gain-control element in color video and some lower frequency R_F applications. Extracting the optimum performance from video AGC systems takes careful attention to circuit details.

[1] Signal-to-noise ratio, S/N = 20 × log(RMS signal/RMS noise).

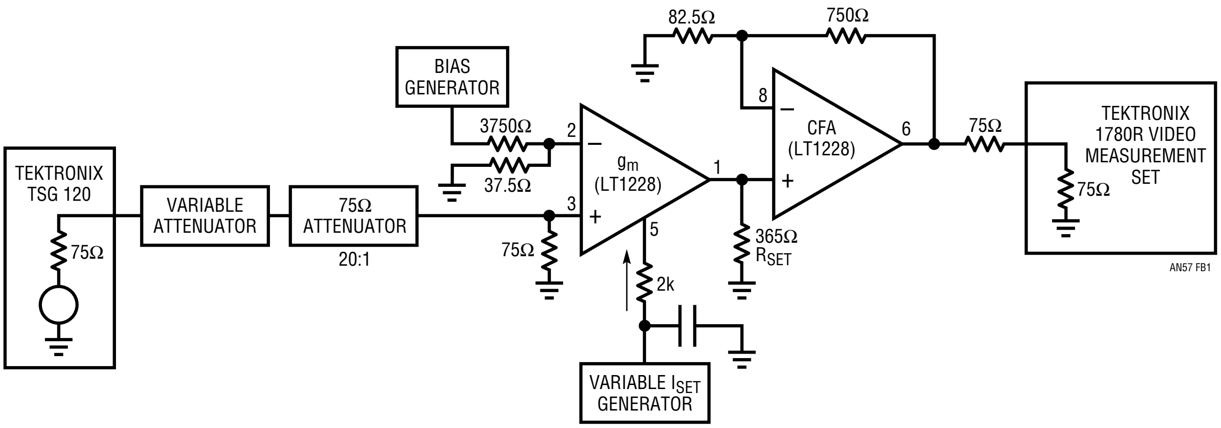


Figure B1. Schematic Diagram

As an example of this optimization, consider the typical gain-control circuit using the LT1228 shown in Figure B1. The input is NTSC composite video, which can cover a 10dB range from 0.56V to 1.8V. The output is to be $1V_{P-P}$ into 75Ω. Amplitudes were measured from peak negative chroma to peak positive chroma on an NTSC modulated ramp test signal. See "Differential Gain and Phase" box.

Notice that the signal is attenuated 20:1 by the 75Ω attenuator at the input of the LT1228, so the voltage on the input (pin 3) ranges from 0.028V to 0.090V. This is done to limit distortion in the transconductance stage. The gain of this circuit is controlled by the current into the I_{SET} terminal, pin 5 of the IC. In a closed-loop AGC system, the loop-control circuitry generates this current by comparing the output of a detector[2] to a reference voltage, integrating the difference and then converting to a suitable current. The measured performance for this circuit is presented in tables B1 and B2. Table B1 has the uncorrected data and Table B2 shows the results of the correction.

All video measurements were taken with a Tektronix 1780R video-measurement set, using test signals generated by a Tektronix TSG 120. The standard criteria for characterizing NTSC video color distortion are the differential gain and the differential phase. For a brief explanation of these tests see the box "Differential Gain and Phase." For this design exercise the distortion limits were set at a somewhat arbitrary 3% for differential gain and 3° for differential phase. Depending on conditions, this should be barely visible on a video monitor.

[2] One way to do this is to sample the colorburst amplitude with a sample-and-hold and peak detector. The nominal peak-to-peak amplitude of the colorburst for NTSC is 40% of the peak luminance.

Table B1. Measured Performance Data (Uncorrected)

INPUT (V)	I_{SET}(mA)	DIFFERENTIAL GAIN	DIFFERENTIAL PHASE	S/N
0.03	1.93	0.5%	2.7°	55dB
0.06	0.90	1.2%	1.2°	56dB
0.09	0.584	10.8%	3.0°	57dB

Table B2. Measured Performance Data (Corrected)

INPUT (V)	BIAS VOLTAGE	I_{SET}(mA)	DIFFERENTIAL GAIN	DIFFERENTIAL PHASE	S/N
0.03	0.03	1.935	0.9%	1.45°	55dB
0.06	0.03	0.889	1.0%	2.25°	56dB
0.09	0.03	0.584	1.4%	2.85°	57dB

Figures B2 and B3 plot the measured differential gain and phase, respectively, against the input signal level (the curves labeled "A" show the uncorrected data from Table B1). The plots show that increasing the input signal level beyond 0.06V results in a rapid increase in the gain distortion, but comparatively little change in the phase distortion. Further attenuating the input signal (and consequently increasing the set current) would improve the differential gain performance but degrade the S/N. What this circuit needs is a good tweak!

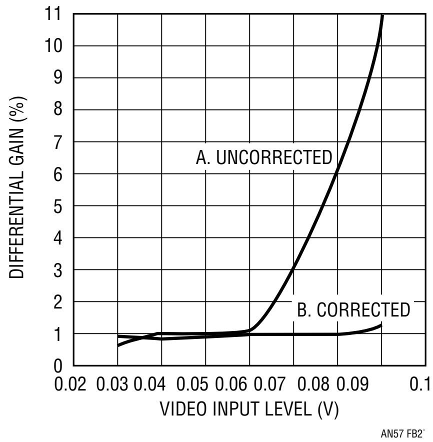


Figure B2. Differential Gain vs Input Level

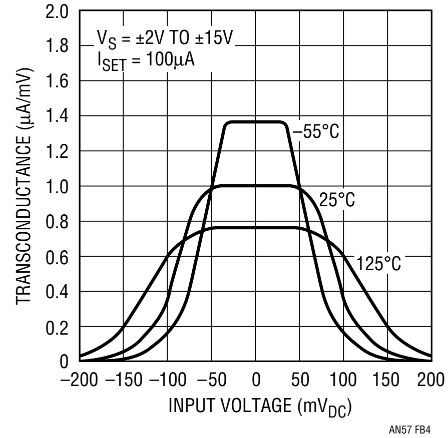


Figure B4. Small-Signal Transconductance vs DC Input Voltage

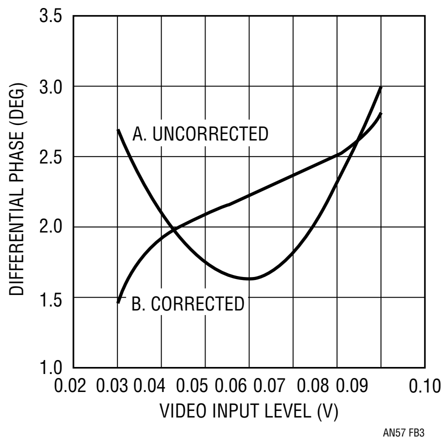


Figure B3. Differential Phase vs Input Level

Optimizing for Differential Gain

Referring to the small signal transconductance versus DC input voltage graph (Figure B4), observe that the transconductance of the amplifier is linear over a region centered around zero volts.[3] The 25°C g_m curve starts to become quite nonlinear above 0.050V. This explains why the differential gain (see Figure B2, curve A) degrades so quickly with signals above this level. Most RF signals do not have DC bias levels, but the composite video signal is mostly unipolar.

[3] Notice also that the linear region expands with higher temperature. Heating the chip has been suggested.

Video is usually clamped at some DC level to allow easy processing of sync information. The sync tip, the chroma reference burst, and some chroma signal information swing negative, but 80% of the signal that carries the critical color information (chroma) swings positive. Efficient use of the dynamic range of the LT1228 requires that the input signal have little or no offset. Offsetting the video signal so that the critical part of the chroma waveform is centered in the linear region of the transconductance amplifier allows a larger signal to be input before the onset of severe distortion. A simple way to do this is to bias the unused input (in this circuit the inverting input, pin 2) with a DC level.

In a video system it might be convenient to clamp the sync tip at a more negative voltage than usual. Clamping the signal prior to the gain-control stage is good practice because a stable DC reference level must be maintained.

The optimum value of the bias level on pin 2 used for this evaluation was determined experimentally to be about 0.03V. The distortion tests were repeated with this bias voltage added. The results are reported in Table B2 and Figures B2 and B3. The improvement to the differential phase is inconclusive, but the improvement in the differential gain is substantial.

Differential Gain and Phase

Differential gain and phase are sensitive indications of chroma signal distortion. The NTSC system encodes color information on a separate subcarrier at 3.579545MHz. The color subcarrier is directly summed to the black and white video signal. The black and white information is a voltage proportional to image intensity and is called luminance or luma. Each line of video has a burst of 9 to 11 cycles of the subcarrier (so timed that it is not visible) that is used as a phase reference for demodulation of the color information of that line. The color signal is relatively immune to distortions, except for those that cause a phase shift or an amplitude error to the subcarrier during the period of the video line.

Differential gain is a measure of the gain error of a linear amplifier at the frequency of the color subcarrier. This distortion is measured with a test signal called a modulated ramp (shown in Figure B5). The modulated ramp consists of the color subcarrier frequency superimposed on a linear ramp or sometimes on a stair step. The ramp has the duration of the active portion of a horizontal line of video. The amplitude of the ramp varies from zero to the maximum level of the luminance, which, in this case, is 0.714V. The gain error corresponds to compression or expansion by the amplifier (sometimes called "incremental gain") and is expressed as a percentage of the full amplitude range. An appreciable amount of differential gain will cause the luminance to modulate the chroma causing visual chroma distortion. The effect of differential gain errors is to change the saturation of the color being displayed. Saturation is the relative degree of dilution of a pure color with white. A 100% saturated color has 0% white, a 75% saturated color has 25% white, and so on. Pure red is 100% saturated whereas, pink is red with some percentage of white and is therefore less than 100% saturated.

Differential phase is a measure of the phase shift in a linear amplifier at the color subcarrier frequency when the modulated ramp signal is used as an input.

The phase shift is measured relative to the colorburst on the test waveform and is expressed in degrees. The visual effect of the distortion is a change in hue. Hue is the quality of perception which differentiates the frequency of the color, red from green, yellow-green from yellow, and so forth.

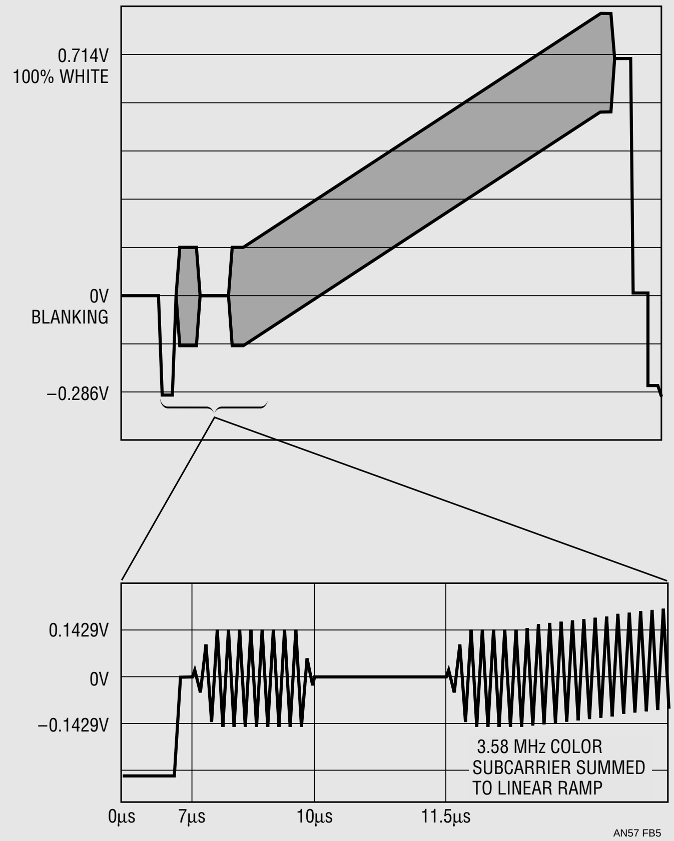


Figure B5. NTSC Test Signal

Three degrees of differential phase is about the lower limit that can unambiguously be detected by observers. This level of differential phase is just detectable on a video monitor as a shift in hue, mostly in the yellow-green region. Saturation errors are somewhat harder to see at these levels of distortion—3% of differential gain is very difficult to detect on a monitor. The test is performed by switching between a reference signal, SMPTE (Society of Motion Picture and Television Engineers) 75% color bars, and a distorted version of the same signal with matched signal levels. An observer is then asked to note any difference.

In professional video systems (studios, for instance) cascades of processing and gain blocks can reach hundreds of units. In order to maintain a quality video signal, the distortion contribution of each processing block must be a small fraction of the total allowed distortion budget[4] because the errors are cumulative. For this reason, high-quality video amplifiers will have distortion specifications as low as a few thousandths of a degree for differential phase and a few thousandths of a percent for differential gain.

[4] From the preceding discussion, the limits on visibility are about 3° differential phase, 3% differential gain. Please note that these are not hard and fast limits. Tests of perception can be very subjective.

APPENDIX C

Using a Fast Analog Multiplexer to Switch Video Signals for NTSC "Picture-in-Picture" Displays

The majority of production[1] video switching consists of selecting one video source out of many for signal routing or scene editing. For these purposes the video signal is switched during the vertical interval in order to reduce visual switching transients. The image is blanked during this time, so if the horizontal and vertical synchronization and subcarrier lock are maintained, there will be no visible artifacts. Although vertical-interval switching is adequate for most routing functions, there are times when it is desirable to switch two synchronous video signals during the active (visible) portion of the line to obtain picture-in-picture, key, or overlay effects. Picture-in-picture or active video switching requires signal-to-signal transitions that are both clean and fast. A clean transition should have a minimum of pre-shoot, overshoot, ringing, or other aberrations commonly lumped under the term "glitching."

Using the LT1204

A quality, high speed multiplexer amplifier can be used with good results for active video switching. The important specifications for this application are a small, controlled switching glitch, good switching speed, low distortion, good dynamic range, wide bandwidth, low path loss, low channel-to-channel crosstalk, and good channel-to-channel offset matching. The LT1204 specifications match these requirements quite well, especially in the areas of bandwidth, distortion, and channel-to-channel crosstalk which is an outstanding –90dB at 10MHz. The LT1204 was evaluated for use in active video switching with the test setup shown in Figure C1. Figure C2 shows the video waveform of a switch between a 50% white level and a 0% white level about 30% into the active interval and back again at about 60% of the active interval. The switch artifact is brief and well controlled. Figure C3 is an expanded view of the same waveform. When viewed on a monitor, the switch artifact is just visible as a very fine line. The lower trace is a switch between two black level (0V) video signals showing a very slight channel-to-channel offset, which is not visible on the monitor. Switching between two DC levels is a worst-case test as almost any active video will have enough variation to totally obscure this small switch artifact.

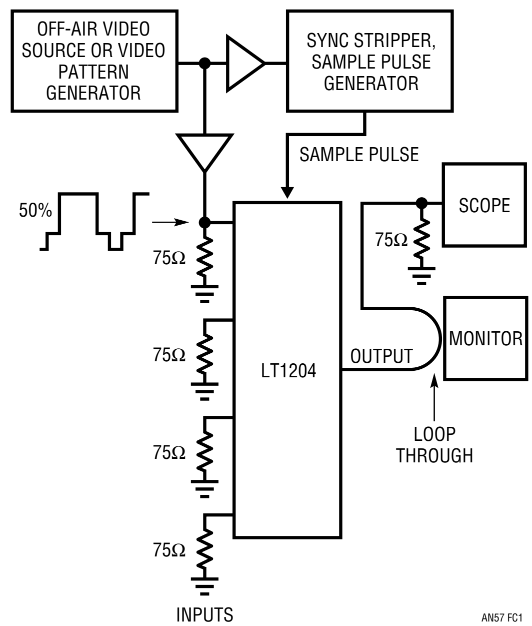
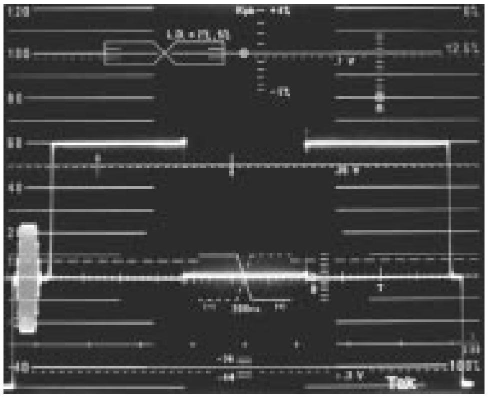


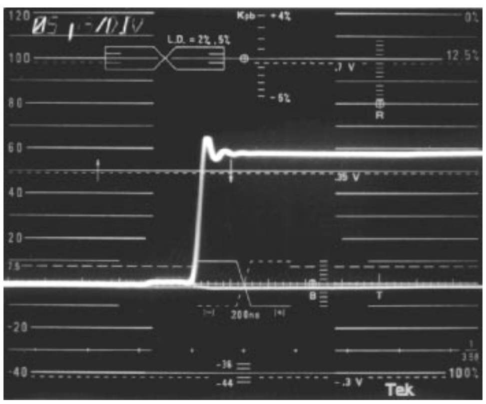
Figure C1. "Picture-in-Picture" Test Setup

[1] Video production, in the most general sense, means any purposeful manipulation of the video signal, whether in a television studio or on a desktop PC.



AN57 FC2

Figure C2. Video Waveform Switched from 50% White Level to 0% White Level and Back



AN57 FC3

Figure C3. Expanded View of Rising Edge of LT1204 Switching from 0% to 50% (50ns Horizontal Division)

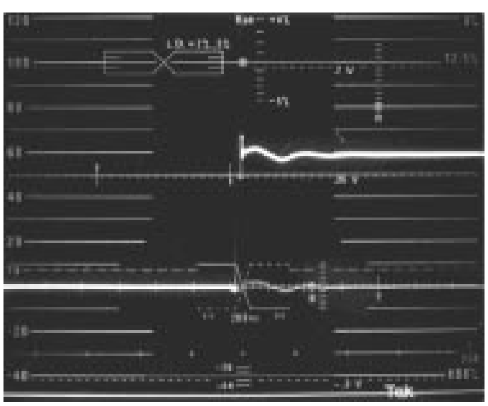


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Figure C4. Expanded View of "Brand-X" Switch 0% to 50% Transistion

Video-Switching Caveats

In a video processing system that has a large bandwidth compared to the bandwidth of the video signal, a fast transition from one video level to another with a low-amplitude glitch will cause minimal visual disturbance. This situation is analogous to the proper use of an analog oscilloscope. In order to make accurate measurement of pulse waveforms, the instrument must have much more bandwidth than the signal in question (usually five times the highest frequency). Not only should the glitch be small, it should be otherwise well controlled. A switching glitch that has a long settling "tail" can be more troublesome (that is, more visible) than one that has more amplitude but decays quickly. The LT1204 has a switching glitch that is not only low in amplitude but well controlled and quickly damped. Refer to Figure C4 which shows a video multiplexer that has a long, slow-settling tail. This sort of distortion is highly visible on a video monitor.

Composite video systems, such as NTSC, are inherently band-limited and thus edge-rate limited. In a sharply band-limited system, the introduction of signals that contain significant energy higher in frequency than the filter cutoff will cause distortion of transient waveforms (see Figure C5). Filters used to control the bandwidth of these video systems should be group-delay equalized to minimize this pulse distortion. Additionally, in a band-limited system, the edge rates of switching glitches or level-to-level transitions should be controlled to prevent ringing and other pulse aberrations that could be visible. In practice, this is usually accomplished with pulse-shaping networks. Bessel filters are one example. Pulse-shaping networks and delay-equalized filters add cost and complexity to video systems and are usually found only on expensive equipment. Where cost is a determining factor in system design, the exceptionally low amplitude and brief duration of the LT1204's switching artifact make it an excellent choice for active video switching.

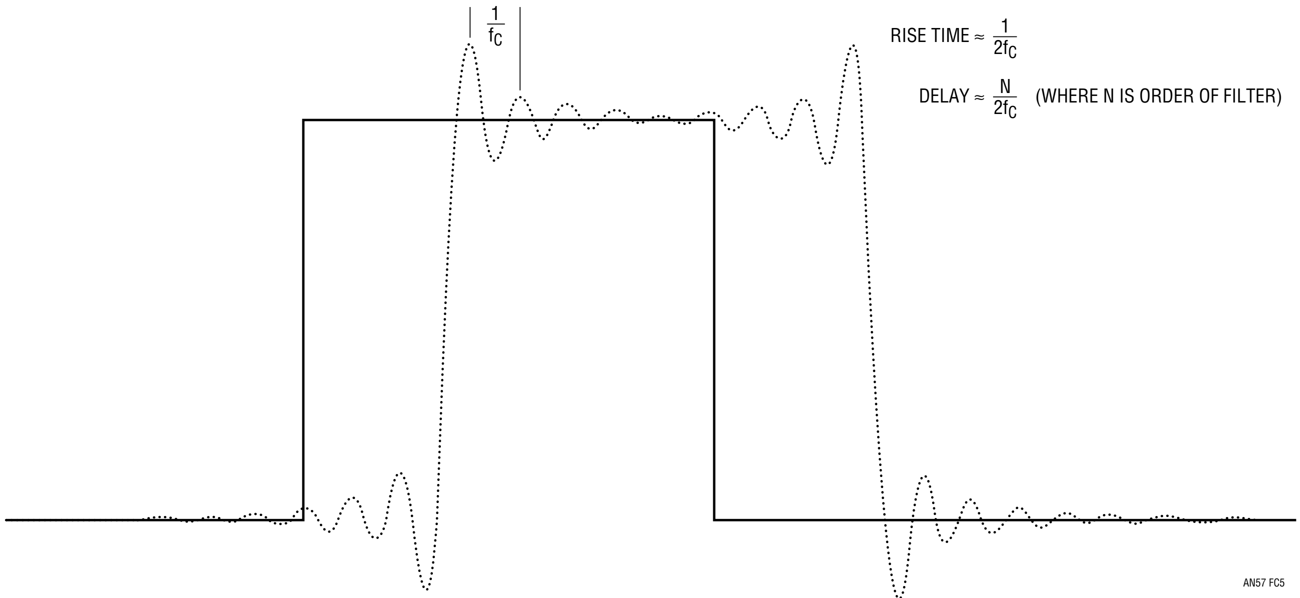

Figure C5. Pulse Response of an Ideal Sharp-Cutoff Filter at Frequency f_C

Conclusion

Active video switching can be accomplished inexpensively and with excellent results when care is taken with both the selection and application of the high speed multiplexer. Both fast switching and small, well-controlled switching glitches are important. When the LT1204 is used for active video switching between two flat-field video signals (a very critical test) the switching artifacts are nearly invisible. When the LT1204 is used to switch between two live video signals, the switching artifacts are invisible.

Some Definitions—

"Picture-in-picture" refers to the production effect in which one video image is inserted within the boundaries of another. The process may be as simple as splitting the screen down the middle or it may involve switching the two images along a complicated geometric boundary. In order to make the composite picture stable and viewable, both video signals must be in horizontal and vertical sync. For composite color signals, the signals must also be in subcarrier lock.

"**Keying**" is the process of switching among two or more video signals triggering on some characteristic of one of the signals. For instance, a chroma keyer will switch on the presence of a particular color. Chroma keyers are used to insert a portion of one scene into another. In a commonly used effect, the TV weather person (the "talent") appears to be standing in front of a computer generated weather map. Actually, the talent is standing in front of a specially colored background; the weather map is a separate video signal, which has been carefully prepared to contain none of that particular color. When the chroma keyer senses the keying color, it switches to the weather map background. Where there is no keying color, the keyer switches to the talent's image.

Linear Technology Corporation
1630 McCarthy Blvd., Milpitas, CA 95035-7487
(408) 432-1900 • FAX: (408) 434-0507 • TELEX: 499-3977

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