Application Note 57
January 1994

Video Circuit Collection

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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
LT6553	1200 (A = 2)	T	A = 2 (Fixed), 6ns Settling Time
LT6555	1200 (A = 2)	T	2:1 MUX, A = 2 (Fixed)
LT1226	1000 ($A_V \geq 25$)	S	400V/µs SR, Good DC Specs
LT6557	1000 (A = 2)	T	A = 2 (Fixed), Automatic Bias for Single Supply
LT6554	650 (A = 1)	T	A = 1 (Fixed), 6ns Settling Time
LT6556	650 (A = 1)	T	2:1 MUX, A = 1 (Fixed)
LT1222	500 ($A_V \geq 10$)	S	12-Bit Accurate
LT1395/LT1396/LT1397	400	S, D, Q	CFA, DG = 0.02%, DP = 0.04%, 0.1dB Flat to 100MHz
LT1818/LT1819	400	S, D	900V/µs SR, DG = 0.07%, DP = 0.02%
LT1192	350 ($A_V \geq 5$)	S	Low Voltage, ±50mA Output
LT1194	350 ($A_V = 10$)	S	Differential Input, Low Voltage, Fixed Gain of 10
LT6559	300	T	CFA, Independant Enable Controls, Low Cost
LT1398/LT1399	300	D, T	CFA, Independant Enable Controls
LT1675-1/LT1675	250 (A = 2)	S, T	2:1 MUX, A = 2 (Fixed)
LT1815/LT1816/LT1817	220	S, D, Q	750V/µs SR, DG = 0.08%, DP = 0.04%
LT6210/LT6211	200	S, D	CFA, Adjustable Speed and Power
LT1809/LT1810	180	S, D	Low Voltage, Rail-to-Rail Input and Output
LT1203/LT1205	170	D, Q	MUX, 25ns Switching, DG = 0.02%, DP = 0.04°
LT1193	160 ($A_V \geq 2$)	S	Low Voltage, Differential Input, Adjustable Gain, ±50mA Output
LT1221	150 ($A_V \geq 4$)	S	250V/µs SR, 12-Bit Accurate
LT1227	140	S	CFA, 1100V/µs SR, DG = 0.01%, DP = 0.01°, Shutdown
LT1259/LT1260	130	D, T	RGB CFA, 0.1dB Flat to 30MHz, DG = 0.016%, DP = 0.075°, Shutdown
LT6550/LT6551	110 (A = 2)	T, Q	Low Voltage, Single Supply, A = 2 (Fixed)
LT1223	100	S	CFA, 12-Bit Accurate, Shutdown, 1300V/µs SR, Good DC Specs, DG = 0.02%, DP = 0.12°
LT1229/LT1230	100	D, Q	CFA, 1000V/µs SR, DG = 0.04%, DP = 0.1°
LT1252	100	S	CFA, DG = 0.01%, DP = 0.09°, Low Cost
LT1812	100	S	Low Power, 200V/µs SR
LT6205/LT6206/LT6207	100	S, D, Q	3V Single Supply
LT1191	90	S	Low Voltage, ±50mA Output
LT1253/LT1254	90	D, Q	CFA, DG = 0.03%, DP = 0.28°, Flat to 30MHz, 0.1dB
LT1813/LT1814	85	D, Q	Low Power, 200V/µs SR
LT1228	80 (gm = 0.25)	S	Transconductance Amp + CFA, Extremely Versatile
LT6552	75 ($A_V \geq 2$)	S	Differential Input, Low Power, Low Voltage
LT1204	70	S	CFA, 4-Input Video MUX Amp, 1000V/µs SR, Superior Isolation

PART	GBW (MHz)	CONFIGURATION	COMMENTS
LT1363/LT1364/LT1365	70	S, D, Q	1000V/µs SR, I_S = 7.5mA per Amp, Good DC Specs
LT1206	60	S	250mA Output Current CFA, 600V/µs SR, Shutdown
LT1187	50 ($A_V \ge 2$)	S	Differential Input, Low Power
LT1190	50	S	Low Voltage
LT1360/LT1361/LT1362	50	S, D, Q	600V/µs SR, I_S = 5mA per Amp, Good 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
LT1189	35 ($A_V \ge 10$)	S	Differential Input, Low Power, Decompensated
LT1995	32 (A = 1)	S	Internal Resistor Array
LT1213/LT1214	28	D, Q	Single Supply, Excellent DC Specs
LT1358/LT1359	25	D, Q	600V/µs SR, I_S = 2.5mA per Amp, Good DC Specs
LT1215/LT1216	23	D, Q	Single Supply, Excellent DC Specs
LT1211/LT1212	14	D, Q	Single Supply, Excellent DC Specs
LT1355/LT1356	12	D, Q	400V/µs SR, I_S = 1.25mA per Amp, Good DC Specs
LT1200/LT1201/LT1202	11	S, D, Q	I_S = 1mA per Amp, Good DC Specs
LT1217	10	S	CFA, I_S = 1mA, Shutdown

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 CABLE DRIVERS

AC-Coupled Video Drivers

When AC-coupling video, the waveform dynamics change with respect to the bias point of the amplifier according to the scene brightness of the video stream. In the worst case, $1V_{P-P}$ video (composite or Luminance + Sync in Y/C or YP_BP_R format) can exhibit a varying DC content of 0.56V, with the dynamic requirement being +0.735V/–0.825V about the nominal bias level. When this range is amplified by two to properly drive a back-terminated cable, the amplifier output must be able to swing $3.12V_{P-P}$, thus a 5V supply is generally required in such circuits, provided the amplifier output saturation voltages are sufficiently small. The following circuits show various realizations of AC-coupled video cable drivers.

Figure 1 shows the LT1995 as a single-channel driver. All the gain-setting resistors are provided on-chip to minimize part count.

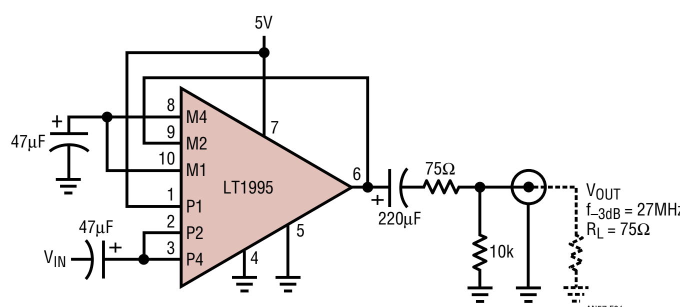


Figure 1. Single Supply Video Line Driver

Figure 2 shows an LT6551 quad amplifier driving two sets of "S-video" (Y/C format) output cables from a single Y/C source. Internal gain-setting resistors within the LT6551 reduce part-count.

Figure 3 shows the LT6553 ultra-high-speed triple video driver configured for single-supply AC-coupled operation. This part is ideal for HD or high-resolution workstation applications that demand high bandwidth and fast settling. The amplifier gains are factory-set to two by internal resistors.

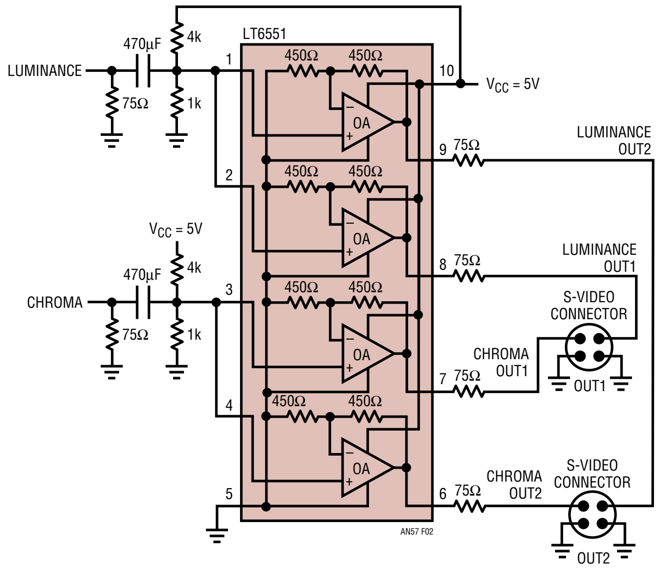


Figure 2. S-Video Splitter

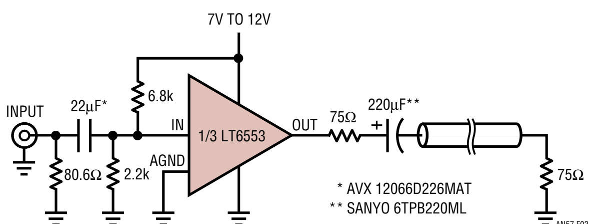


Figure 3. Single Supply Configuration, One Channel Shown

The LT6557 400MHz triple video driver is specifically designed to operate in 5V single supply AC-coupled applications as shown in Figure 4. The input biasing circuitry is contained on-chip for minimal external component count. A single resistor programs the biasing level of all three channels.

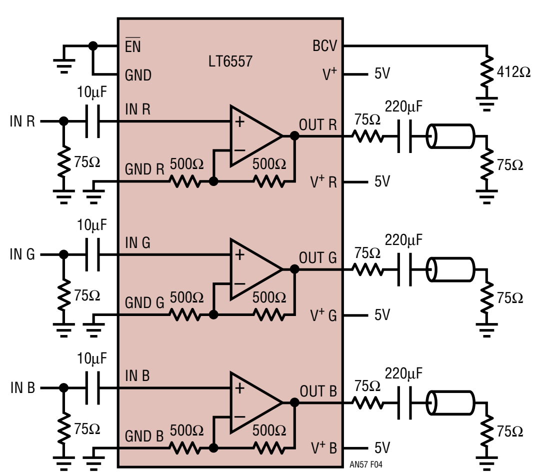


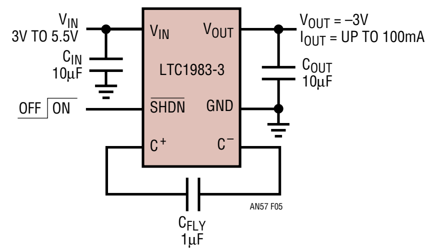
Figure 4. 400MHz, AC-Coupled, 5V Single Supply Video Driver

DC-Coupled Video Drivers

The following circuits show various DC-coupled video drivers. In DC-coupled systems, the video swings are fixed in relation to the supplies used, so back-terminated cable-drivers need only provide 2V of output range when optimally biased. In most cases, this permits operation on lower power supply potential(s) than with AC-coupling (unclamped mode). Generally DC-coupled circuits use split supply potentials since the waveforms often include or pass through zero volts. For single supply operation, the inputs need to have an appropriate offset applied to preserve linear amplifier operation over the intended signal swing.

For systems that lack an available negative supply, the LT1983-3 circuit shown in Figure 5 can be used to easily produce a local-use –3V that can simplify an overall cable-driving solution, eliminating large output electrolytics, for example.

Figure 6 shows a typical 3-channel video cable driver using an LT6553. This part includes on-chip gain-setting resistors and flow-through layout that is optimal for HD and RGB wideband video applications. This circuit is a good



C_{FLY}: TAIYO YUDEN LMK212BJ105
C_{IN}, C_{OUT}: TAIYO YUDEN JMK316BJ106ML

Figure 5. –3V at 100mA DC/DC Converter

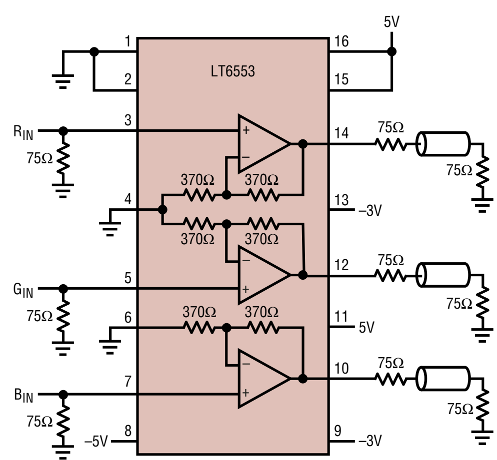


Figure 6. Triple Video Line Driver

candidate for the LT1983-3 power solution in systems that have only 5V available.

Figure 7 shows the LT6551 driving four cables and operating from just 3.3V. The inputs need to have signals centered at 0.83V for best linearity. This application would be typical of standard-definition studio-environment signal distribution equipment (RGBS format).

Figure 8 shows a simple video splitter application using an LT6206. Both amplifiers are driven by the input signal and each is configured for a gain of two, one for driving each output cable. Here again careful input biasing is required (or a negative supply as suggested previously).

Figure 9 shows a means of providing a multidrop tap amplifier using the differential input LT6552. This circuit taps the cable (loop-through configuration) at a high impedance and then amplifies the signal for transmission to a standard 75Ω video load (a display monitor for example). The looped-through signal would continue on to other locations before being terminated. The exceptional common mode rejection of the LT6552 removes any stray noise pickup on the distribution cable from corrupting the locally displayed video. This method is also useful for decoupling of ground-loop noise between equipment, such as in automotive entertainment equipment. To operate on a single supply, the input signals shown (shield and center of coax feed) should be non-negative, otherwise a small negative supply will be needed, such as the local –3V described earlier.

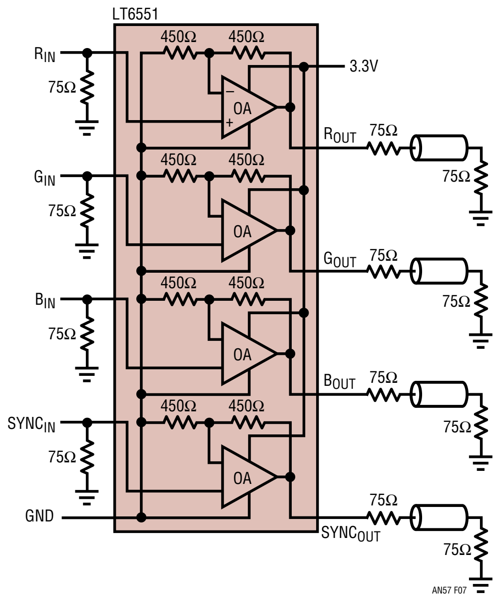


Figure 7. 3.3V Single Supply LT6551 RGB Plus SYNC Cable Driver

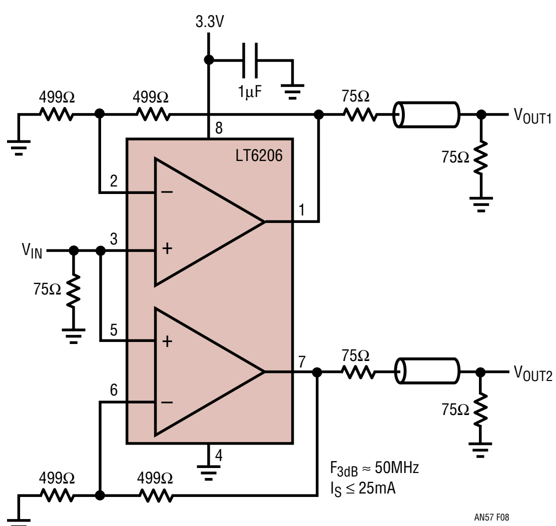


Figure 8. Baseband Video Splitter/Cable Driver

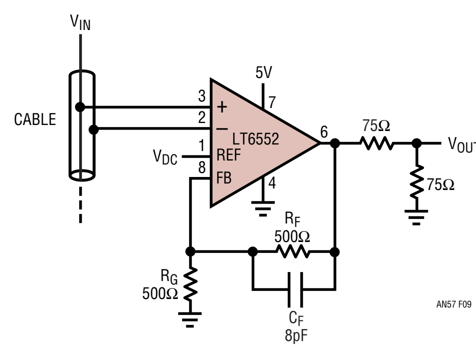


Figure 9. Cable Sense Amplifier for Loop Through Connections with DC Adjust

Clamped AC-Input Video Cable Driver

The circuit in Figure 10 shows a means of driving composite video on standard 75Ω cable with just a single 3.3V power supply. This is possible due to the low output saturation levels of the LT6205 and the use of input clamping to optimize the bias point of the amplifier for standard $1V_{P-P}$ source video. The circuit provides an active gain of two and 75Ω series termination, thus yielding a net gain of one as seen by the destination load (e.g. display device). Additional detail on this circuit and other low-voltage considerations can be found in Design Note 327.

Twisted-Pair Video Cable Driver and Receiver

With the proliferation of twisted-pair wiring practices for in-building data communication, video transmission on the

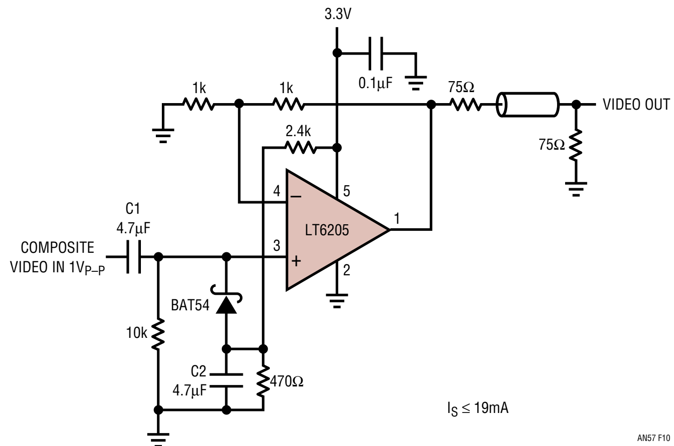


Figure 10. Clamped AC-Input Video Cable Driver

same medium offers substantial cost savings compared to conventional coaxial-cable. Launching a baseband camera signal into twisted pair is a relatively simple matter of building a differential driver such as shown in Figure 11. In this realization one LT6652 is used to create a gain of +1 and another is used to make a gain of –1. Each output is series terminated in half the line impedance to provide a balanced drive condition. An additional virtue of using the LT6552 in this application is that the incoming unbalanced signal (from a camera for example) is sensed differentially, thereby rejecting any ground noise and preventing ground loops via the coax shield.

At the receiving end of the cable, the signal is terminated and re-amplified to re-create an unbalanced output for

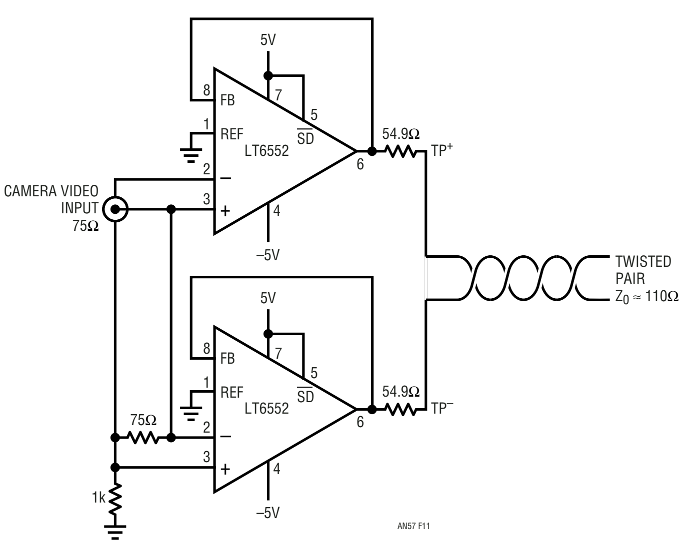


Figure 11. Super-Simple Coax to Twisted-Pair Adapter

connection to display monitors, recorders, etc. The amplifier not only has to provide the 2x gain required for the output drive, but must also make up for the losses in the cable run. Twisted pair exhibits a rolloff characteristic that requires equalization to correct for, so the circuit in Figure 12 shows a suitable feedback network that accomplishes this. Here again the outstanding common mode rejection of the LT6552 is harnessed to eliminate stray pickup that occurs in long cable runs.

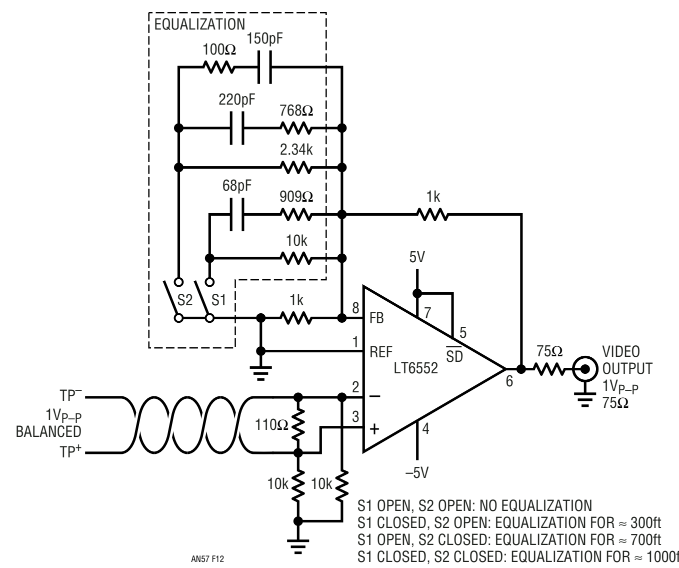


Figure 12. All-In-One Twisted-Pair Video Line Receiver, Cable Equalizer, and Display Driver

VIDEO PROCESSING CIRCUITS

ADC Driver

Figure 13 shows the LT6554 triple video buffer. This is a typical circuit used in the digitization of video within high resolution display units. The input signals (terminations not shown) are buffered to present low source impedance and fast settling behavior to ADC inputs that is generally required to preserve conversion linearity to 10 bits or better. With high resolution ADCs, it is typical that the settling-time requirement (if not distortion performance) will call for buffer bandwidth that far outstrips the baseband signals themselves in order to preserve the effective number of [conversion] bits (ENOBs). The 1kΩ loads shown are simply to represent the ADC input for characterization purposes, they are not needed in the actual use of the part.

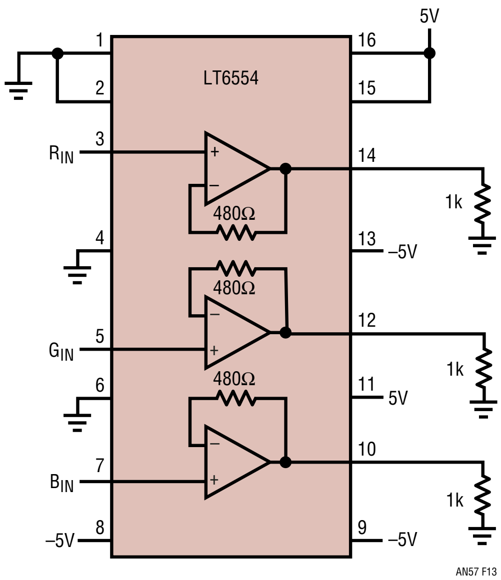


Figure 13. Triple Video Buffer and A/D Driver

Video Fader

In some cases it is desirable to adjust amplitude of a video waveform, or cross-fade between two different video sources. The circuit in Figure 14 provides a simple means of accomplishing this. The 0V to 2.5V control voltage provides a steering command to a pair of amplifier input sections; at each extreme, one section or the other takes complete control of the output. For intermediate control voltages, the inputs each contribute to the output with a weighting that follows a linear function of control voltage (e.g. at $V_{CONTROL}$ = 1.25V, both inputs contribute at 50%). The feedback network to each input sets the maximum gain in the control range (unity gain is depicted in the example), but depending on the application, other gains or even equalization functions can be voltage controlled

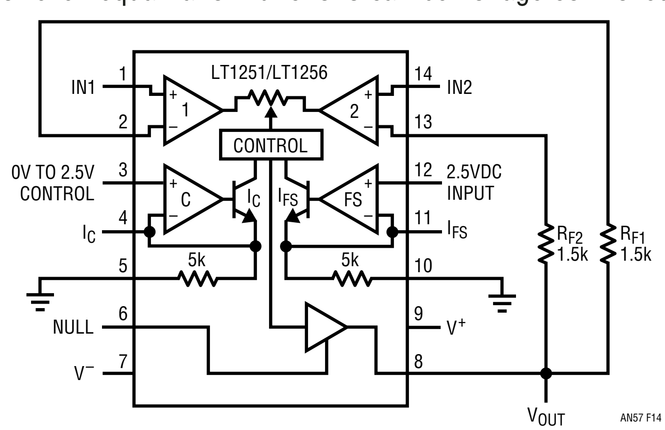


Figure 14. Two-Input Video Fader

(see datasheet and Application Note 67 for additional examples). In the fader example below, it should be noted that both input streams must be gen-locked for proper operation, including a black signal (with sync) if fading to black is intended.

Color Matrix Conversion

Depending on the conventions used by video suppliers in products targeting specific markets, various standards for color signaling have evolved. Television studios have long used RGB cameras and monitor equipment to maximize signal fidelity through the equipment chain. With computer displays requiring maximum performance to provide clear text and graphics, the VESA standards also specify an RGB format, but with separate H and V syncs sent as logic signals. Video storage and transmission systems, on the other hand, seek to minimize information content to the extent that perceptual characteristics of the eye limit any apparent degradation. This has led to utilizing color-differencing approaches that allowed reducing bandwidth on the color information channels without noticeable loss in image sharpness. The consumer 3-channel "component" video connection (YP_BP_R) has a luma + sync (Y) plus blue and red axis color-space signals (P_B and P_R, respectively) that are defined as a matrix multiplication applied to RGB raw video. The color difference signals are typically half the spatial resolution of the luma according to the compression standards defined for DVD playback and digitally broadcast source material, thus lowering "bandwidth" requirements by some 50%. The following circuits show methods of performing color-space mappings at the physical layer (analog domain).

Figure 15 shows a method of generating the standard-definition YP_BP_R signals from an RGB source using a pair of LT6550 triple amplifiers. It should be noted that to ensure Y includes a correct sync, correct syncs should be present at all three inputs or else added directly at the Y output (gated 8.5mA current sink or 350Ω switched to –3.3V). This circuit does not deliberately reduce bandwidth on the color component outputs, but most display devices will nonetheless apply a Nyquist filter at the digitizer section of the "optical engine" in the display unit. The circuit is shown as DC-coupled, so ideally black level is near ground for best operation with the low-voltage supplies shown. Adding input coupling capacitors will allow processing source video that has substantial offset.

An LT6559 and an LT1395 can also be used to map RGB

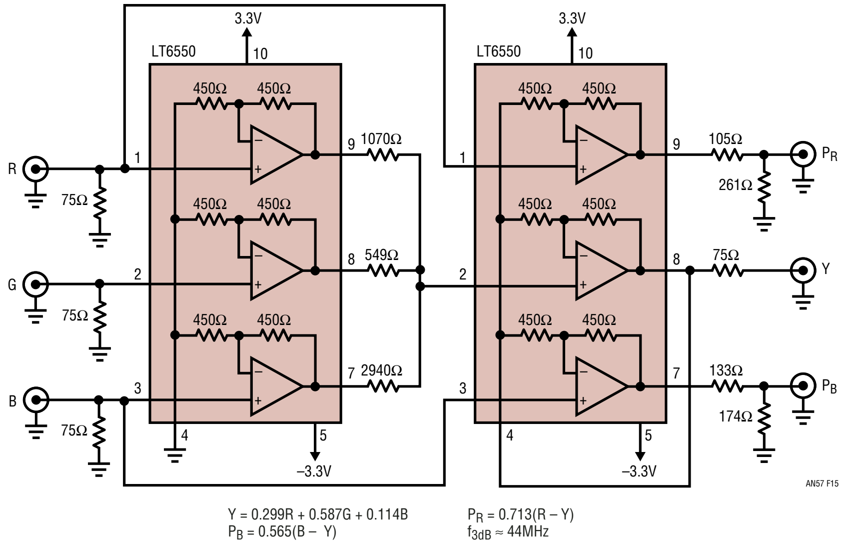


Figure 15. RGB to YP_BP_R Component-Video Conversion

signals into YP_BP_R "component" video as shown in Figure 16. The LT1395 performs a weighted inverting addition of all three inputs. The LT1395 output includes an amplification of the R input by –324/1.07k = –0.3. The amplification of the G input is by –324/549 = –0.59. Finally, the B input is amplified by –324/2.94k = – 0.11. Therefore the LT1395 output is –0.3R, –0.59G, –0.11B = –Y. This output is further scaled and inverted by –301/150 = –2 by LT6559 section A2, thus producing 2Y. With the division by two that occurs due to the termination resistors, the desired Y signal is generated at the load. The LT6559 section A1 provides a gain of 2 for the R signal, and performs a subtraction of 2Y from the section A2 output. The output resistor divider provides a scaling factor of 0.71 and forms the 75Ω back-termination resistance. Thus the signal seen at the terminated load is the desired 0.71(R – Y) = P_R. The LT6559 section A3 provides a gain of 2 for the B signal, and also performs a subtraction of 2Y from the section A2 output. The output resistor divider provides a scaling factor of 0.57 and forms the 75Ω back-termination resistance. Thus the signal seen at the terminated load is the desired 0.57(B – Y) = P_B. As with the previous circuit, to develop a normal sync on the Y signal, a normal sync must be inserted on each of the R, G, and B inputs or injected directly at the Y output with controlled current pulses.

Figure 17 shows LT6552 amplifiers connected to convert component video (YP_BP_R) to RGB. This circuit maps the sync on Y to all three outputs, so if a separate sync connection is needed by the destination device (e.g. studio monitor), any of the R, G, or B channels may be simply looped-through the sync input (i.e. set Z_{IN} for sync input to unterminated). This particular configuration takes advantage of the unique dual-differential inputs of the LT6552 to accomplish multiple arithmetic functions in each stage, thereby minimizing the amplifier count. This configuration also processes the wider-bandwidth Y signal through just a single amplification level, maximizing the available performance. Here again, operation on low supply voltages is predicated on the absence of substantial input offset, and input coupling capacitors may be used if needed (220µF/6V types for example, polarized according to the input offset condition).

Another realization of a component video (YP_BP_R) to RGB adapter is shown in Figure 18 using an LT6207. Amplifier count is minimized by performing passive arithmetic at the outputs, but this requires higher gains, thus a higher supply potential is needed for this (for at least the positive rail). One small drawback to this otherwise compact solution is that the Y channel amplifier must single-handedly drive all three outputs to produce white, so the helper current source shown is needed to increase available drive current. As with the previous circuit, the sync on Y is mapped to all outputs and input coupling-capacitors can be added if the input source has significant offset.

Two LT6559s can also be used to map YP_BP_R "component" video into RGB color space as shown in Figure 19. The Y input is properly terminated with 75Ω and buffered with a gain of 2 by amplifier A2. The P_R input is terminated and buffered with a gain of 2.8 by amplifier A1. The P_B input is terminated and buffered with a gain of 3.6 by amplifier A3. Amplifier B1

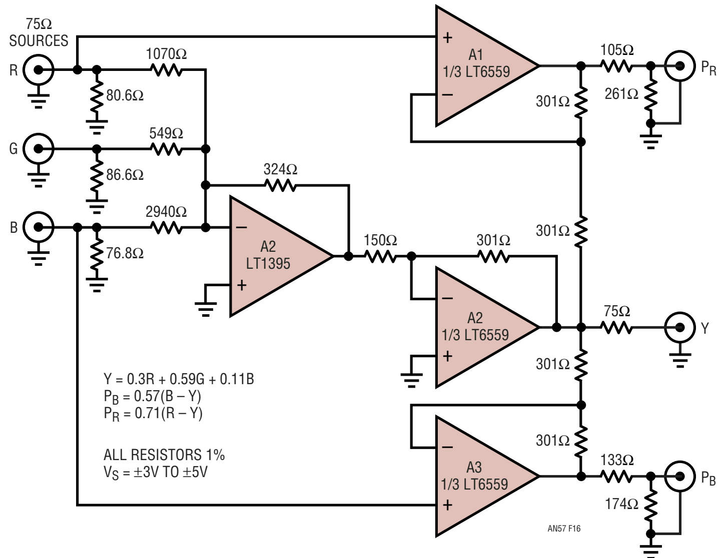


Figure 16. High Speed RGB to YP_BP_R Converter

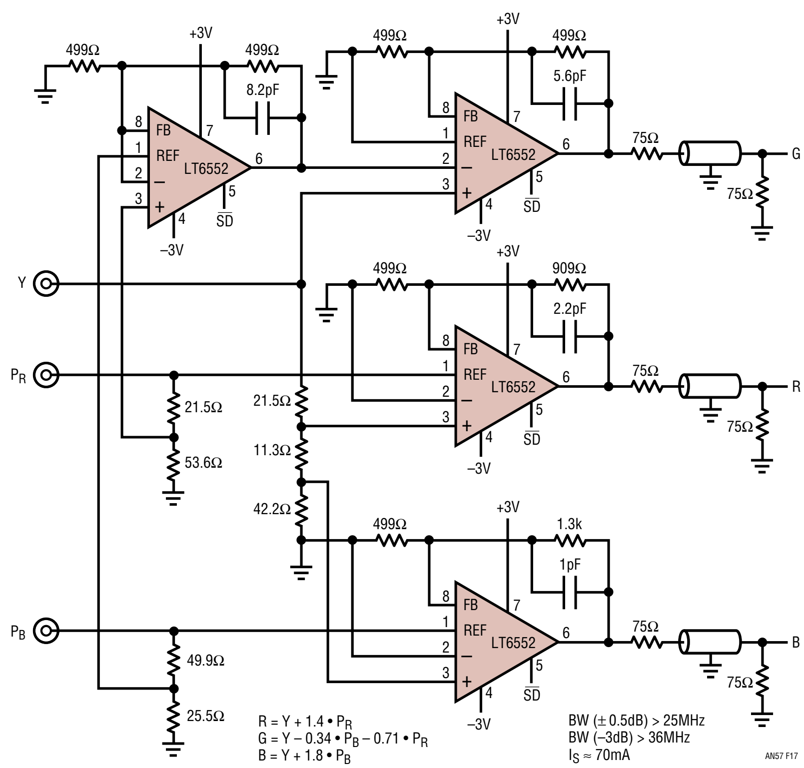


Figure 17. YP_BP_R to RGB Video Converter

performs an equally weighted addition of amplifiers A1 and A2 outputs, thereby producing $2(Y + 1.4P_R)$, which generates the desired R signal at the terminated load due to the voltage division by 2 caused by the termination resistors. Amplifier B3 forms the equally weighted addition of amplifiers A1 and A3 outputs, thereby producing $2(Y + 1.8P_B)$, which generates the desired B signal at the terminated load. Amplifier B2 performs a weighted summation of all three inputs. The P_B signal is amplified overall by $–301/1.54k \bullet 3.6 = 2(–0.34)$. The P_R signal is amplified overall by $–301/590 \bullet 2.8 = 2(–0.71)$. The Y signal is amplified overall by $1k/(1k + 698) \bullet (1 + [301/(590||1.54k)]) \bullet 2 = 2(1)$. Therefore

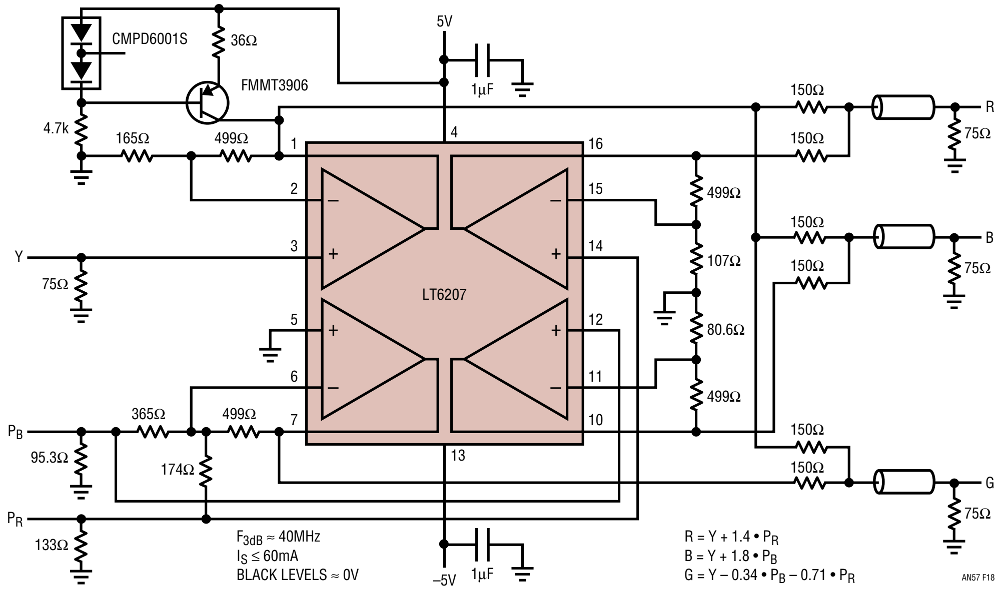


Figure 18. YP_BP_R to RGB Converter

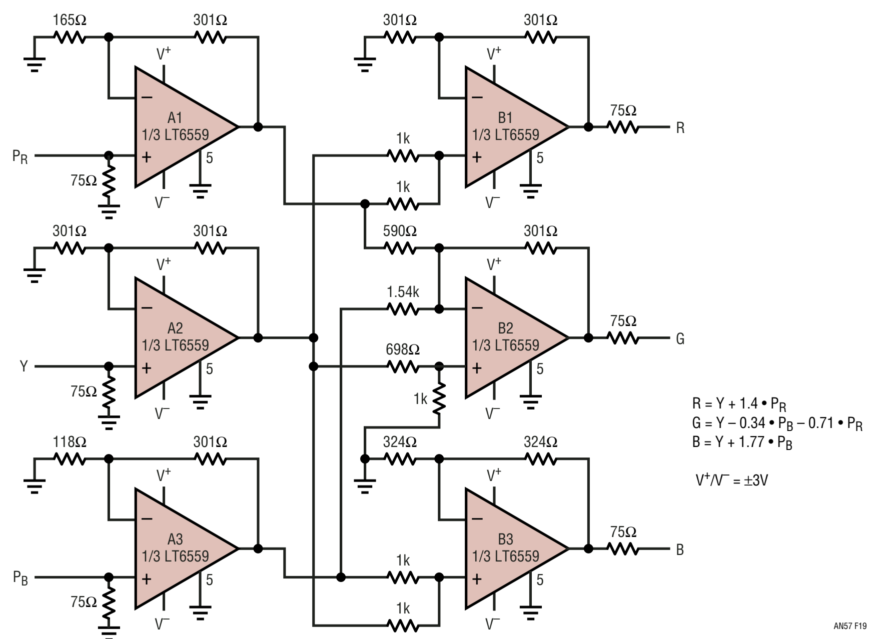


Figure 19. High Speed YP_BP_R to RGB Converter

the amplifier B2 output is $2(Y – 0.34P_B – 0.71P_R)$, which generates the desired G signal at the terminated load. Like the previous circuits shown, sync present on the Y input is reconstructed on all three R, G, and B outputs.

Video Inversion

The circuit in Figure 20 is useful for viewing photographic negatives on video. A single channel can be used for composite or monochrome video. The inverting amplifier stages are only switched in during active video so the blanking, sync and color burst (if present) are not disturbed. To prevent video from swinging negative, a voltage offset equal to the peak video signal is added to the inverted signal.

Graphics Overlay Adder

Multiplexers that provide pixel-speed switching are also useful in providing simple graphics overlay, such as time-stamps or logo “bugs”. Figure 21 shows an LT1675 pair

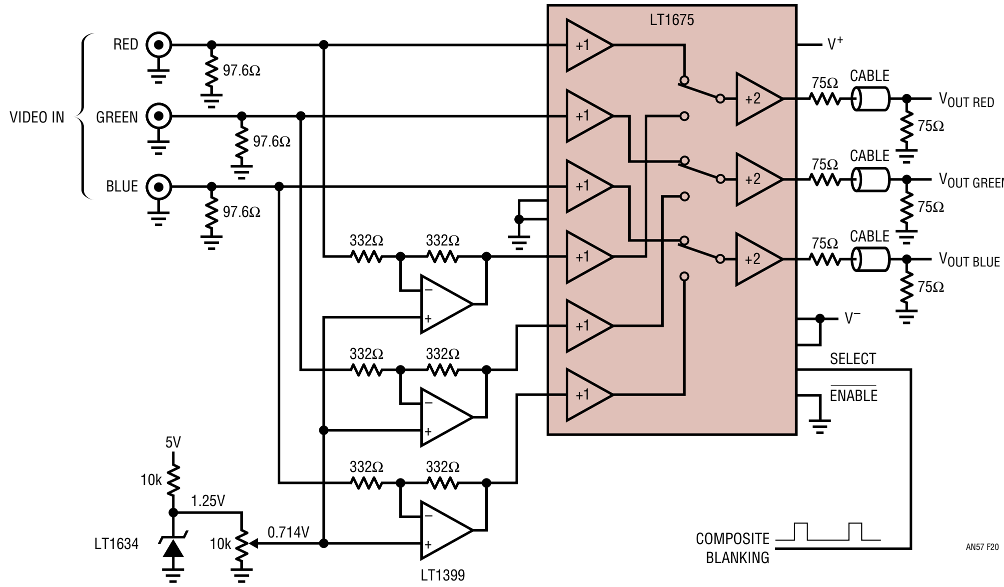
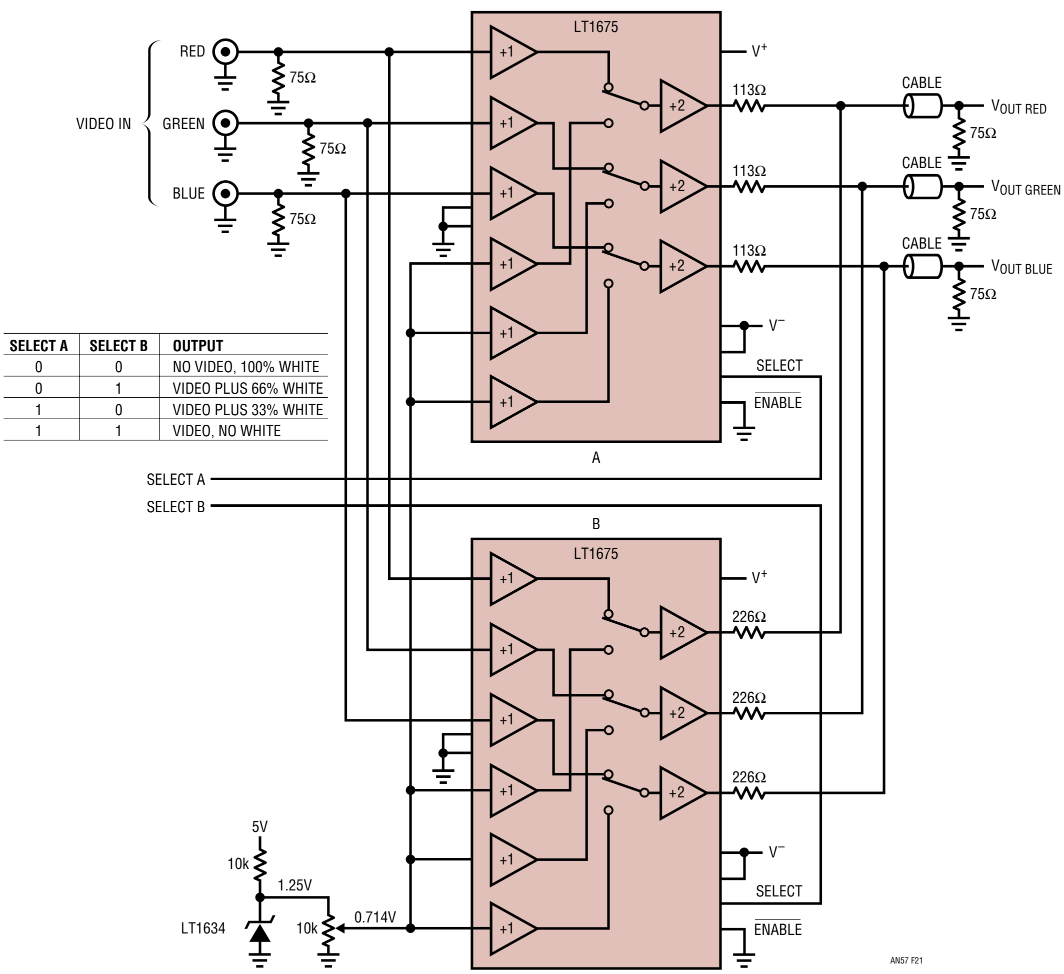


Figure 20. RGB Video Inverter



SELECT A	SELECT B	OUTPUT
0	0	NO VIDEO, 100% WHITE
0	1	VIDEO PLUS 66% WHITE
1	0	VIDEO PLUS 33% WHITE
1	1	VIDEO, NO WHITE

Figure 21. Logo or "Bug" Inserter

used to insert multilevel overlay content from a digital generator. The instantaneous state of the two input control lines selects video or white in each device and combines their outputs with the resistor-weighted summing networks at the output. With the four combinations of control line states, video, white, and two differing brightening levels are available.

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

The circuit in Figure 22 is a variable gain amp suitable for composite video use. Feedback around the transcon-duct-ance 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 23 minimizes the error by careful design.

The following techniques are critical in the design shown in Figure 23:

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.

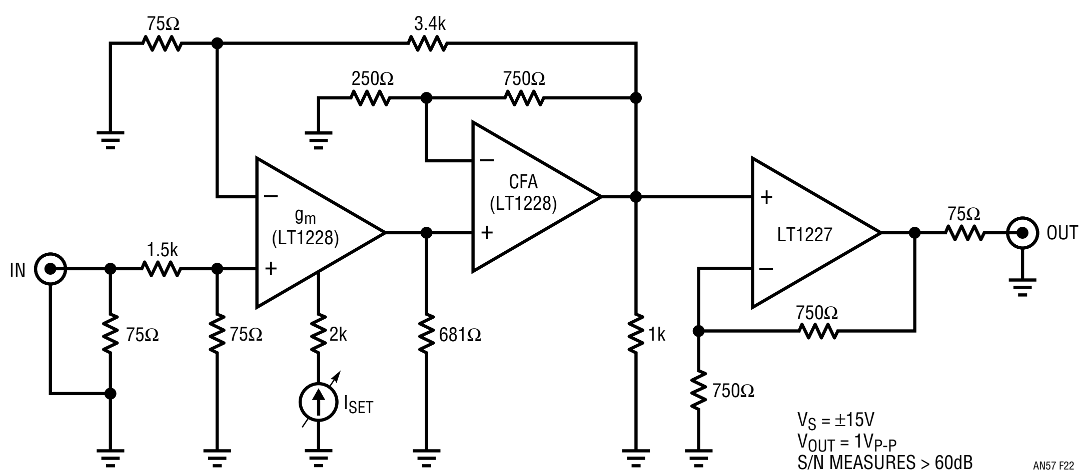


Figure 22. ±3dB Variable Gain Video Amp Optimized for Differential Gain and Phase

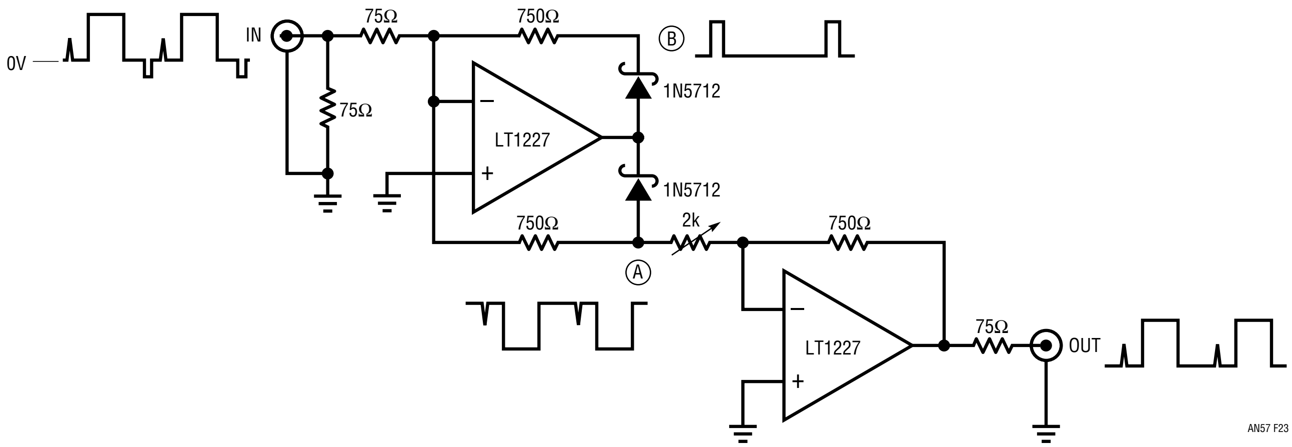


Figure 23. Black Clamp Circuit

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 24) shows the circuit's clean response to a 1T[1] pulse (some extra delay is added between the input and output for clarity).

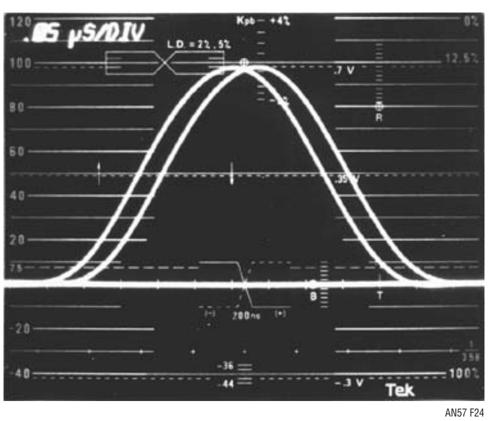


Figure 24. Black Clamp Circuit Response to a "1T" Pulse (±15V Power Supplies)

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 25 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%)].

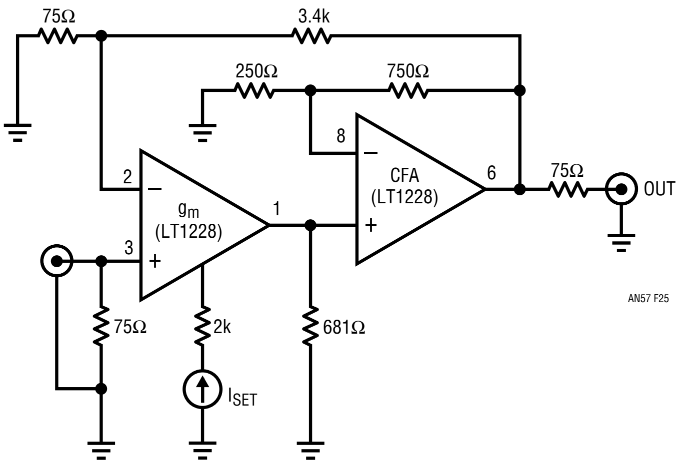


Figure 25. LT1228 Soft Limiter

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 26. Figure 27 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*.

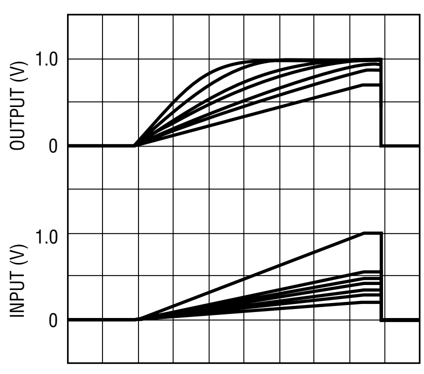


Figure 26. Output of the Limiting Amp (I_{SET} = 0.68mA), with a Ramp Input. As the Input Amplitude Increases from 0.25V to 1V, the Output is Limited to 1V

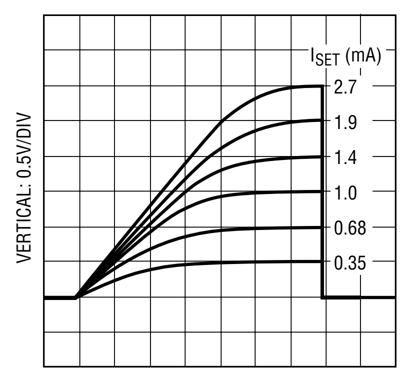


Figure 27. The Output of the Limiting Amp with Various Limiting Levels (I_{SET}). The Input is a Ramp with a Maximum Amplitude of 0.75V

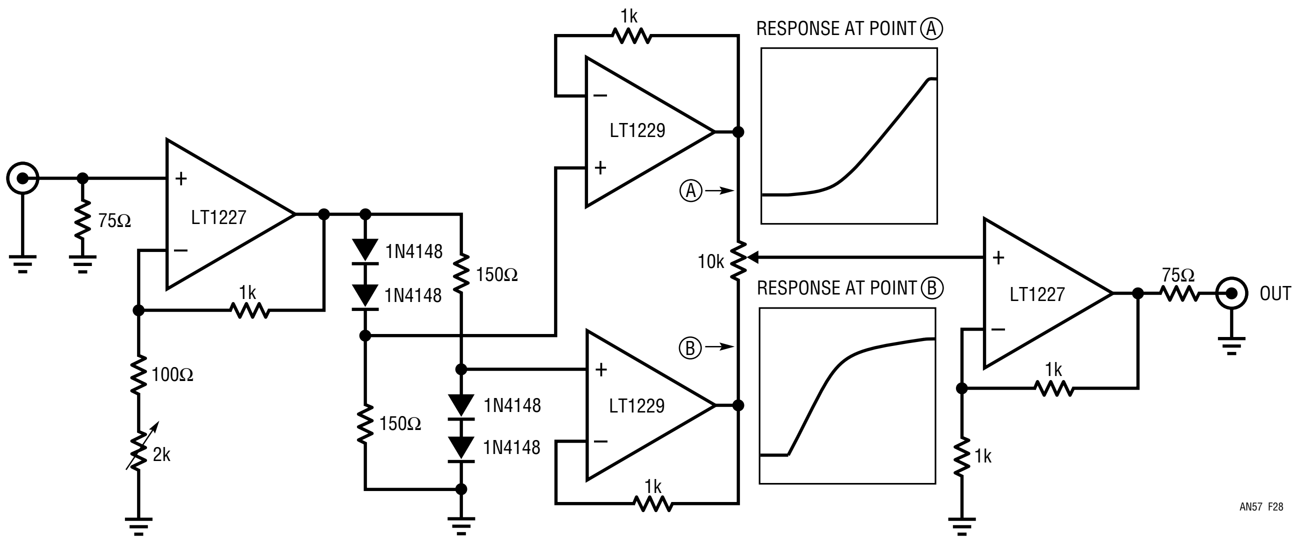


Figure 28. Gamma Amp (Input Video Should Be Clamped)

Figure 28 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 29, curve A, 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. The response of a two diode gamma circuit like that in Figure 28 is shown in Figure 29, curve B. Summing these two curves together, as in Figure 29, 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 29, curve D. Figure 30 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.

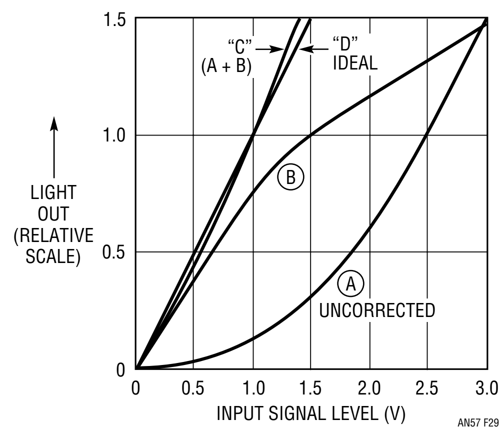


Figure 29. Uncorrected CRT Transfer Function

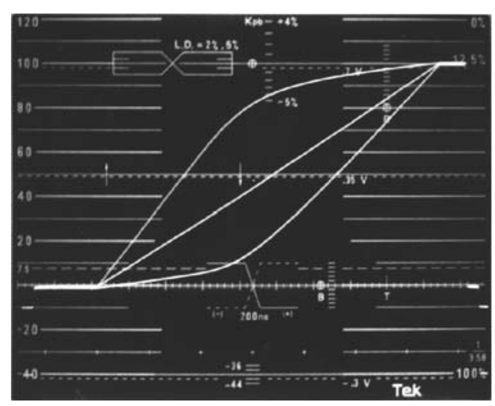


Figure 30. Gamma Corrector Circuit Adjusted for Three Gammas: –3, 1, +3 (Approximately). The Input is a Linear Ramp

LT1228 Sync Summer

The circuit shown in Figure 31a 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 31b 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 31c shows two horizontal

[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.

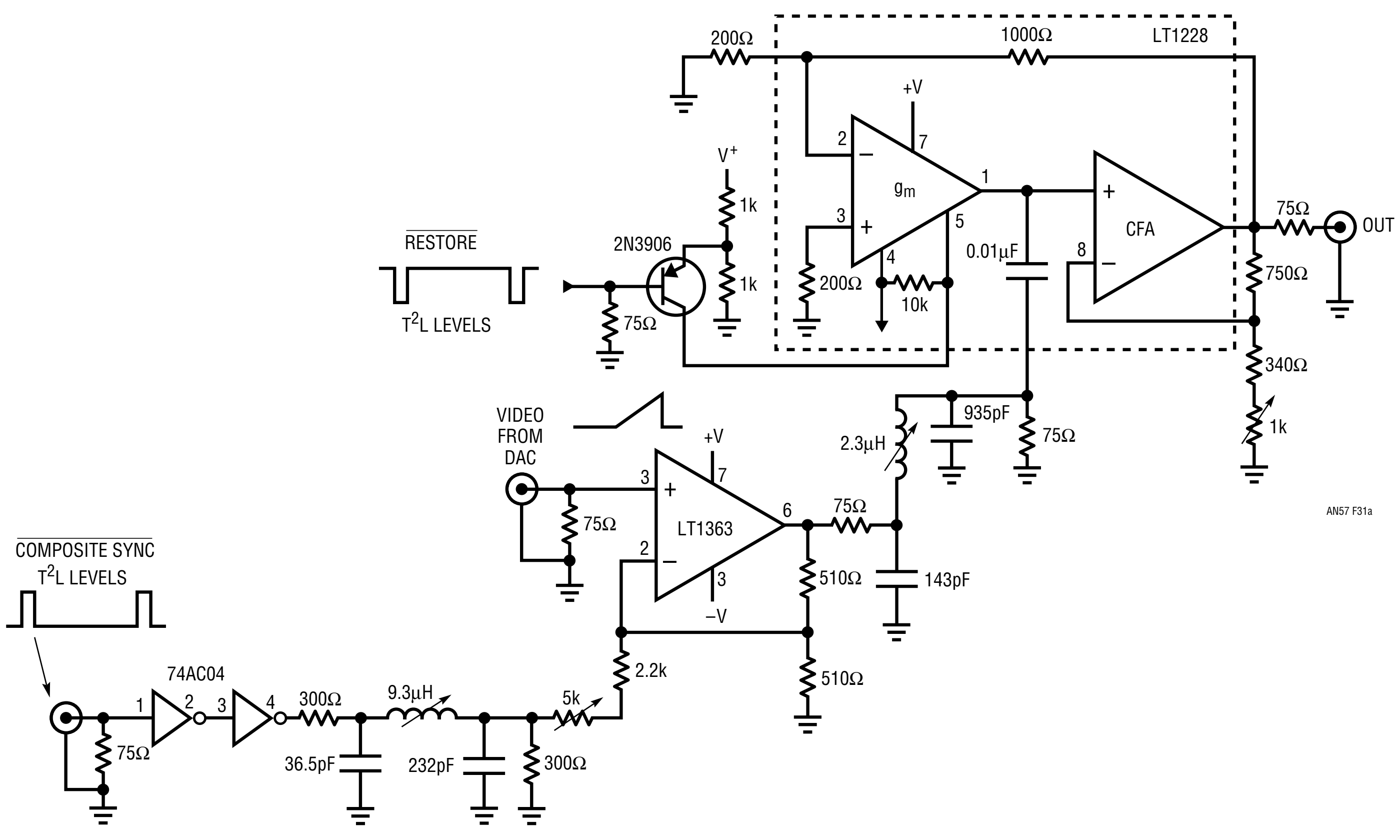
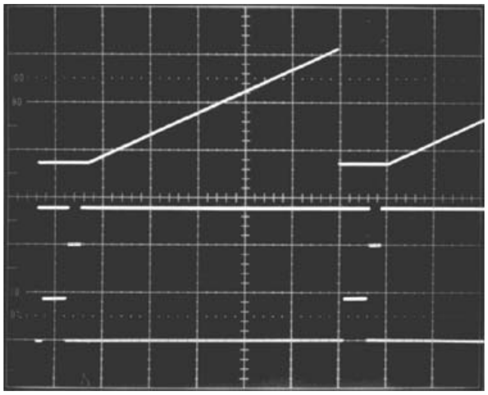
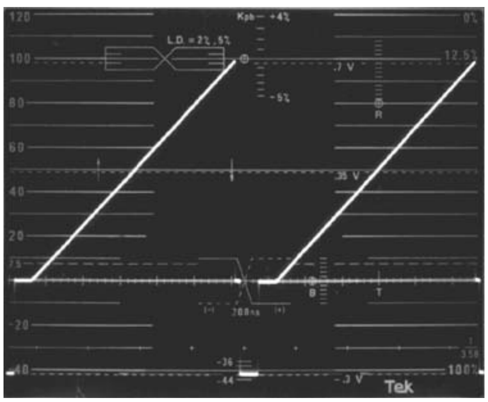


Figure 31a. Simple Sync Summer



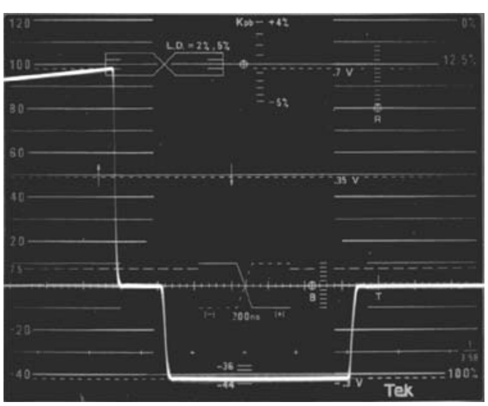
AN57 F31b

Figure 31b. Video Waveform from DAC; Clamp Pulse and Sync Pulse Used as Inputs to Sync Summer



AN57 F31c

Figure 31c. Reconstructed Video Out of Sync Summer



AN57 F31c

Figure 31d. Close-Up of Figure 31c, Showing Sync Pulse

lines of the output waveform with the DC restored and the sync added. Figure 31d is an expanded view of the banking interval showing a clean, well formed sync pulse.

MULTIPLEXER CIRCUITS

Integrated Three-Channel Output Multiplexer

The LT6555 is a complete 3-channel wideband video 2:1 multiplexer with internally set gain of two. This part is ideal as an output port driver for HD component or high-resolution RGB video products. The basic application circuit is shown in Figure 32 with terminations shown on all ports, though in many applications the input loading may not be required. One thing this diagram does not reflect is the convenient flow-through pin assignments of the part, in which no video traces need cross in the printed-circuit layout. This maximizes isolation between channels and sources for best picture quality.

Since the LT6555 includes an enable control line, it is possible to extend the selection range of the multiplexer. Figure 33 shows two LT6555 devices in a configuration that provides 4:1 selection of RGB sources to an RGB output port (these could also be YP_BP_R signals as well, depending on the source). To avoid frequency response anomalies, the

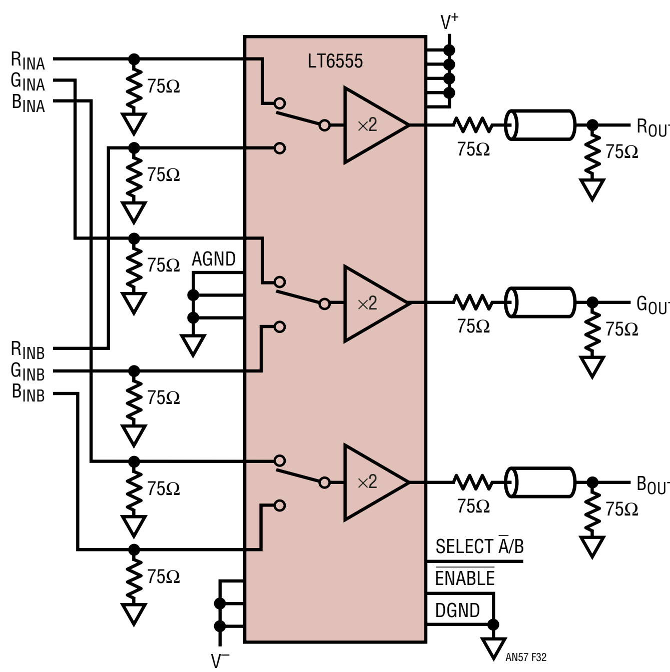


Figure 32. Multiplexer and Line Driver

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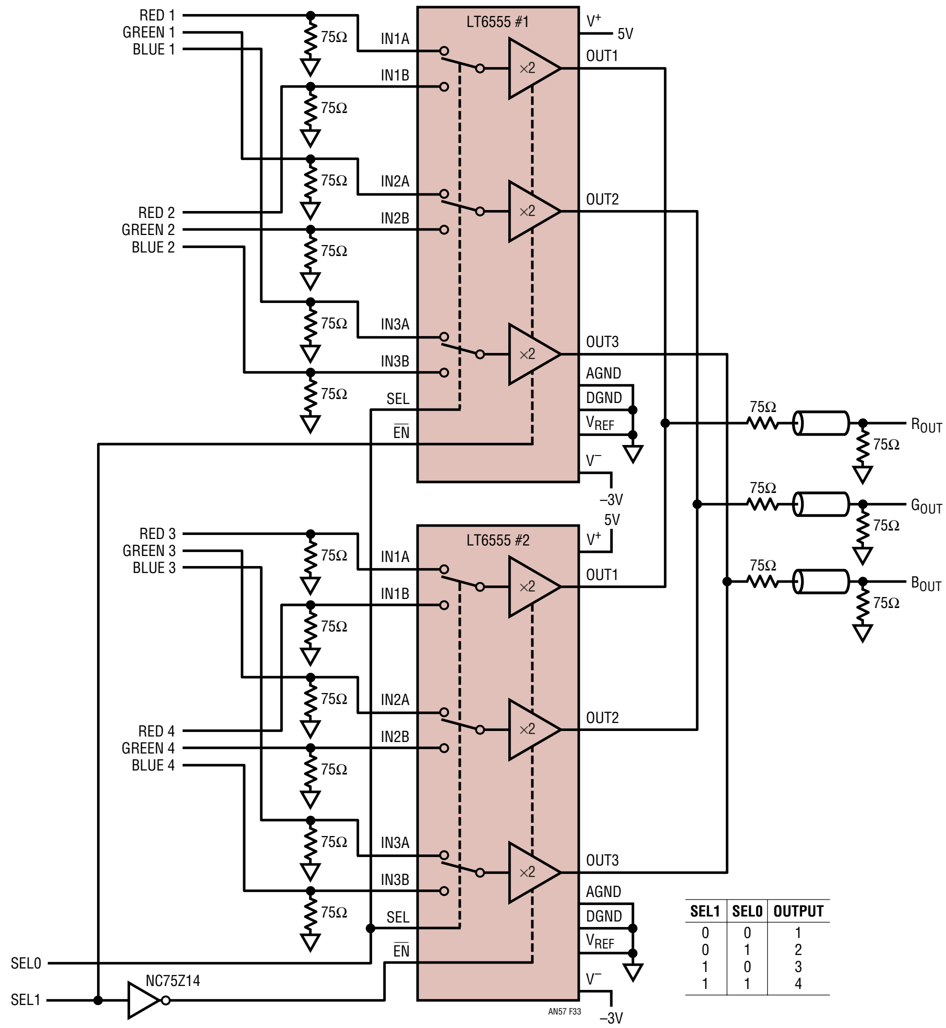


SEL1	SEL0	OUTPUT
0	0	1
0	1	2
1	0	3
1	1	4

Figure 33. 4:1 RGB Multiplexer

two devices should be closely located so that the output lines between parts are as short as possible.

The LT1675 is also an integrated 3-channel 2:1 multiplexer that includes gain of two for cable-driving applications. The basic configuration is shown in Figure 34. A single channel version for composite video applications is available as an LT1675-1.

Integrated Three-Channel Input Multiplexer

The LT6556 is a complete 3-channel wideband video 2:1 multiplexer with internally set gain of one. This part is ideal as an input port receiver for HD component or high-resolution RGB video products. The basic application circuit is shown in Figure 35, with 1kΩ output loads to represent subsequent processing circuitry (the 1kΩ resistors aren't needed, but part characterization was performed with that loading). One thing this diagram does not reflect is the convenient flow-through pin assignments of the part, in which no video traces need cross in the printed-circuit layout. This maximizes isolation between channels and sources for best picture quality.

As with the LT6555, the LT6556 includes an enable control line, so it is possible to extend the selection range of this multiplexer as well. Figure 36 shows two LT6556 devices in a configuration that provides 4:1 selection of RGB sources to an RGB signal processing function, such as a digitizer in a projection system (these could be YP_BP_R signals just as well). To avoid frequency response anomalies, the two devices should be closely located so that the output lines between parts are as short as possible.

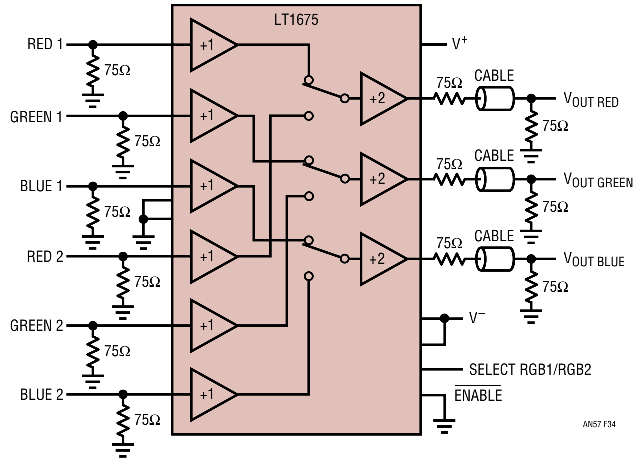


Figure 34. 2:1 RGB Multiplexer and Cable Driver

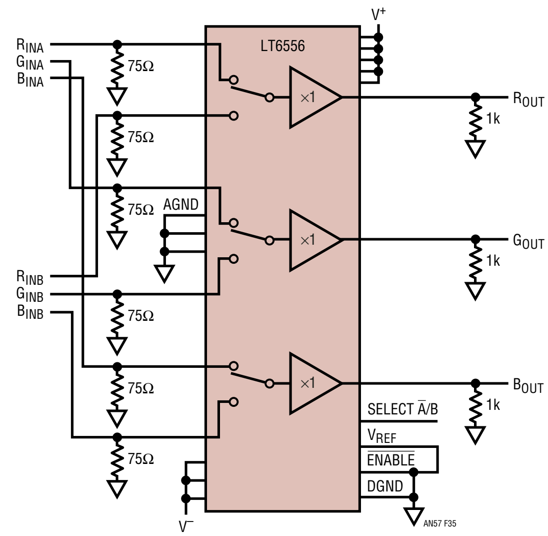
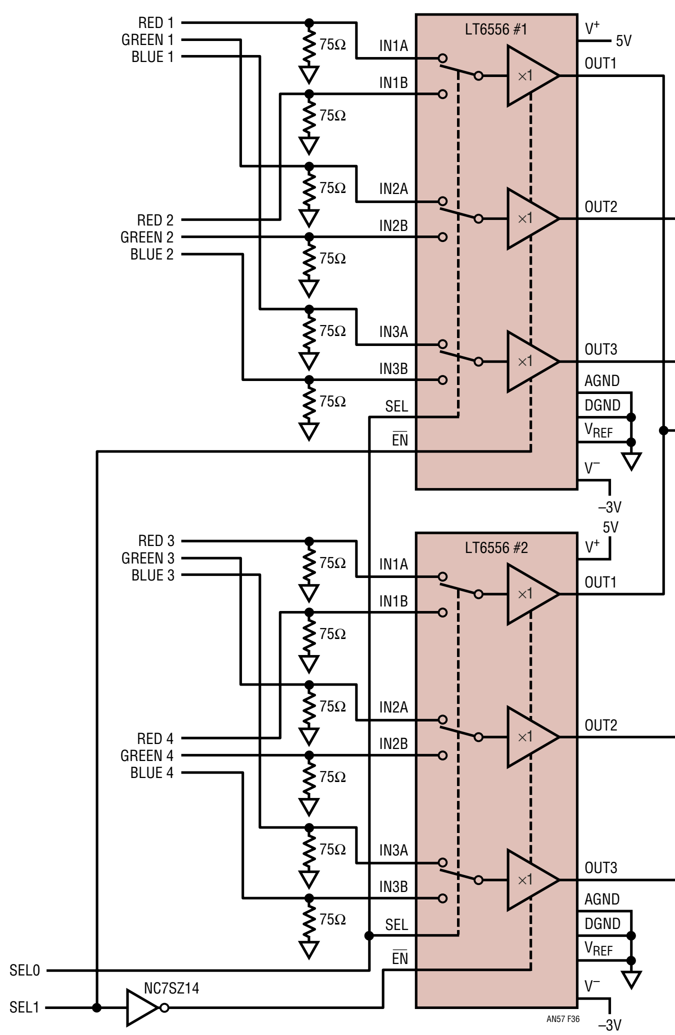


Figure 35. Buffered Input Multiplexer/ADC Driver



SEL1	SEL0	OUTPUT
0	0	1
0	1	2
1	0	3
1	1	4

Figure 36. 4:1 RGB Multiplexer

A 3:1 cable-driving multiplexer for composite video can be formed from a single LT1399 as shown in Figure 37. The LT1399 has the unusual feature of having independent enable controls for each of the three sections. The gain of the amplifiers is set to compensate for passive loss in the loading associated with the off-section feedback networks.

Forming RGB Multiplexers From Triple Amplifiers

The LT6553 triple cable driver and LT6554 triple buffer amp each provide an enable pin, so these parts can be used to implement video multiplexers. Figure 38 shows a pair of LT6553 devices configured as a 2:1 output multiplexer and cable driver. Similarly, Figure 39 shows a pair of LT6554 devices forming a 2:1 input mux, suitable as an ADC driver. These circuits are functionally similar to the LT6555 and LT6556 integrated multiplexers, but offer the flexibility of providing the mux feature as a simple stuffing option to a single printed circuit design, possibly reducing production costs when multiple product grades are being concurrently manufactured. For best results the two devices should be closely located and use minimal trace lengths between them for the shared output signals.

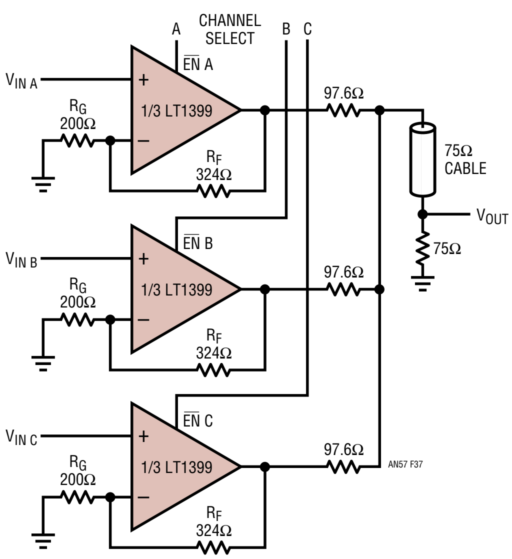


Figure 37. 3-Input Video MUX Cable Driver

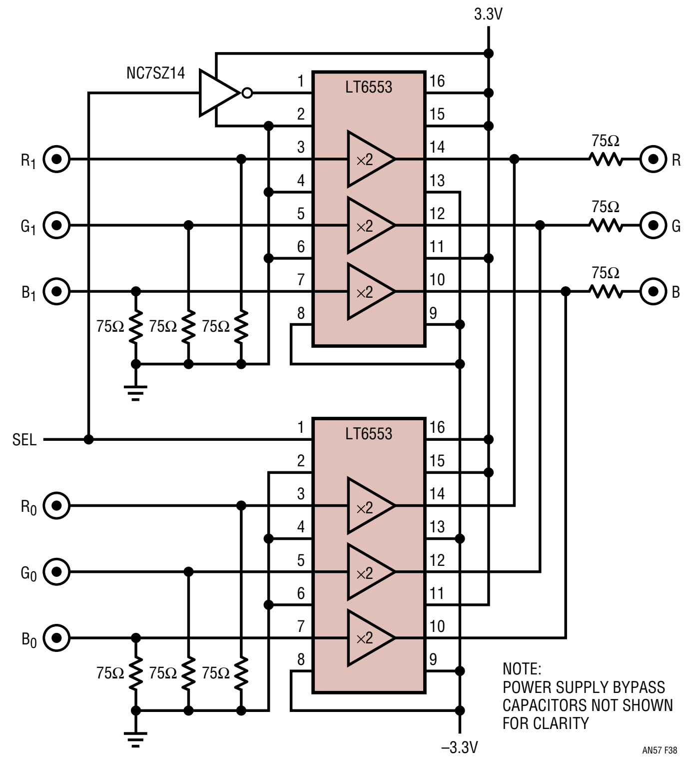


Figure 38. RGB Video Selector/Cable Driver

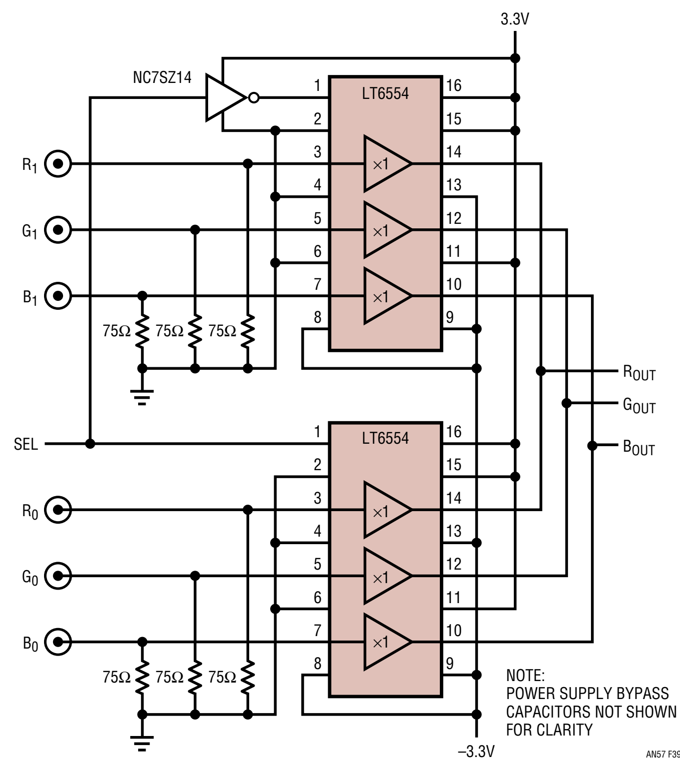


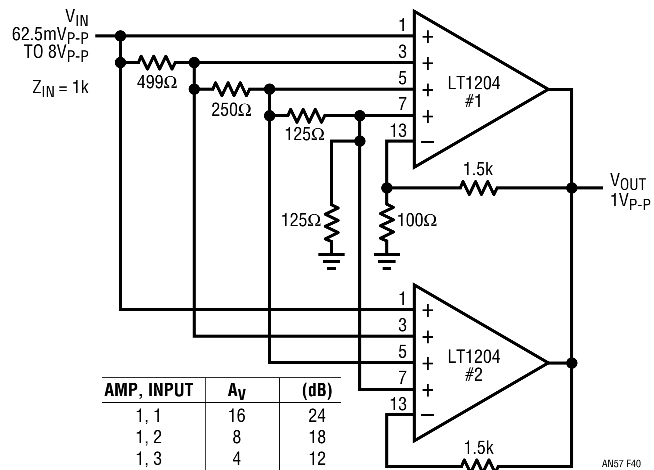
Figure 39. RGB Video Selector and A/D Driver

Stepped Gain Amp Using the LT1204

This is a straightforward approach to a switched-gain amp that features versatility. Figures 40 and 41 show circuits which implement a switched-gain amplifier; Figure 40 features an input Z of 1000Ω, while Figure 41'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 40. 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.

LT1204 Amplifier/Multiplexer Sends Video Over Long Twisted Pair

Figure 42 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, 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 40. Switchable Gain Amplifier Accepts Inputs from 62.5mV$_{P-P}$ to 8V$_{P-P}$

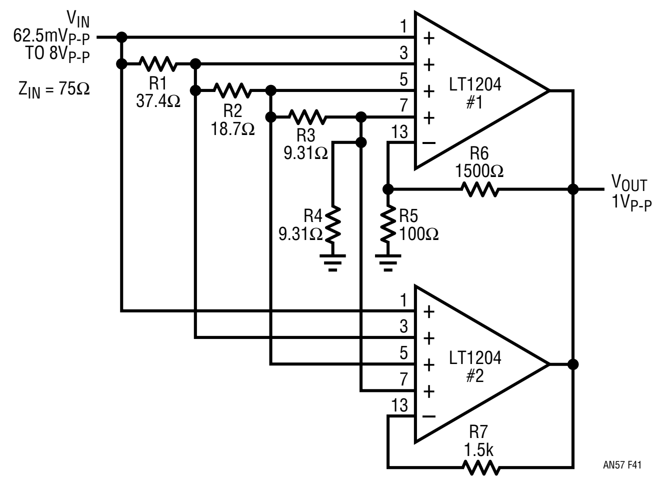


Figure 41. Switchable Gain Amplifier, Z_{IN} = 75Ω Same Gains as Figure 37

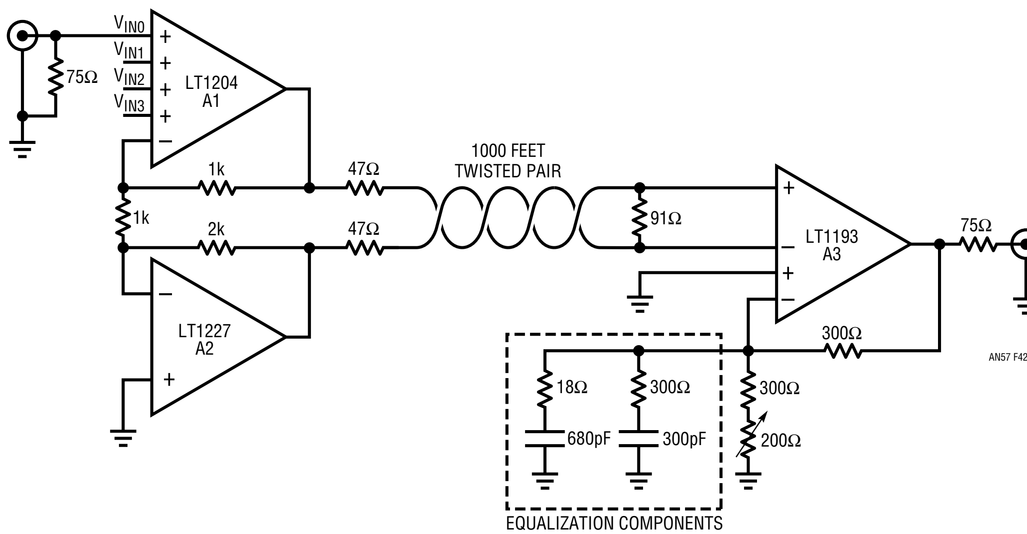


Figure 42. Twisted Pair Driver/Receiver

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 43 shows the video test signal before and after transmission but without equalization. Figure 44 shows before and after with the equalization connected. Differential gain and phase are about 1% and 1°, respectively.

Fast Differential Multiplexer

This circuit (Figure 45) 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 45's circuit also makes a robust differential to single-ended amp/MUX for high speed data acquisition.

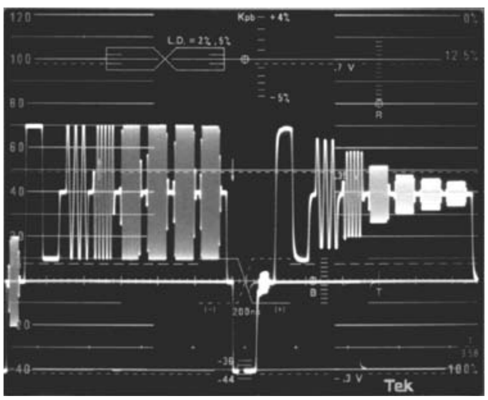


Figure 43. Multiburst Pattern Without Cable Compensation

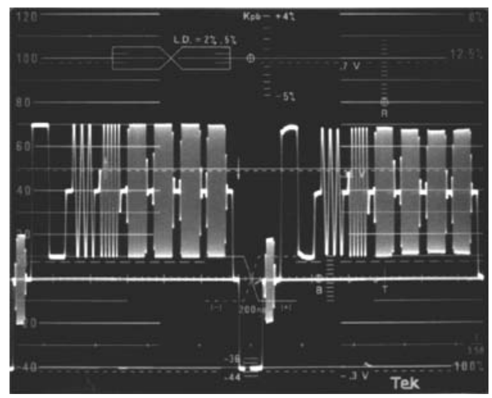


Figure 44. Multiburst Pattern with Cable Compensation

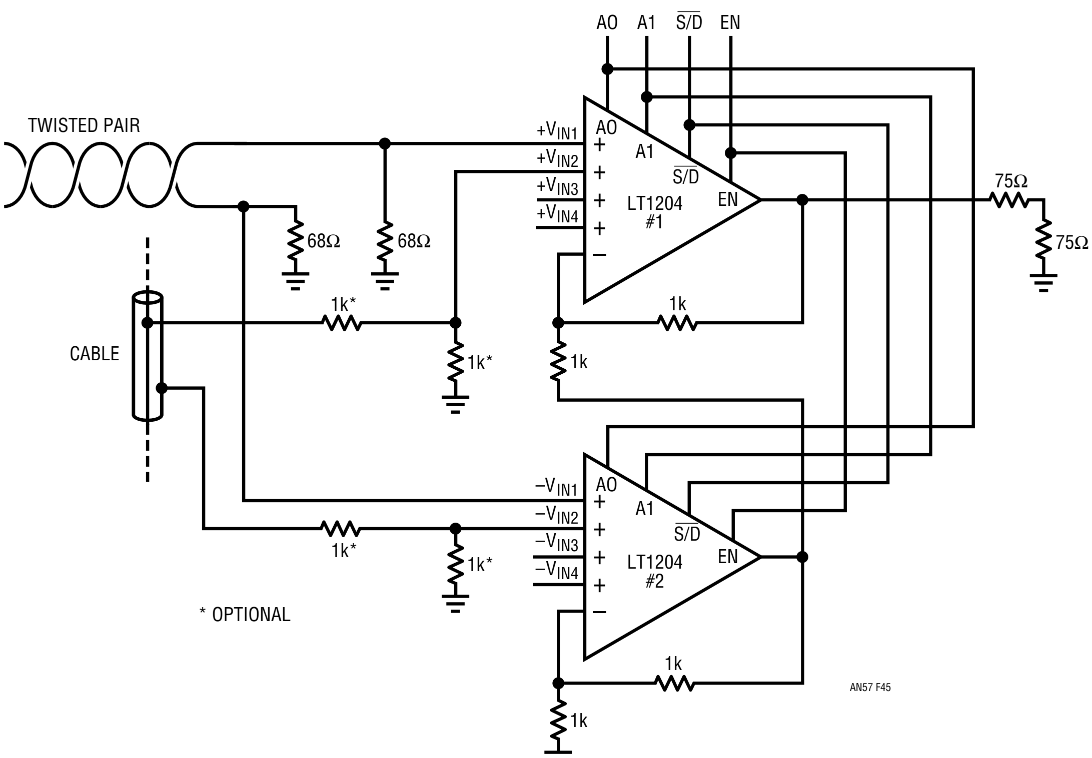


Figure 45. Fast Differential Multiplexer

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 46 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.

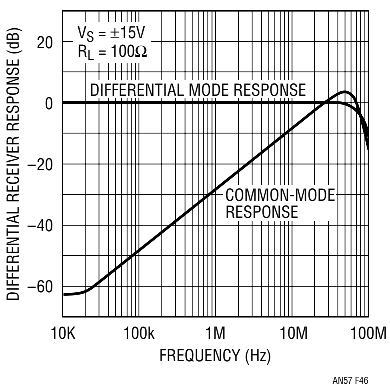


Figure 46. Differential Receiver Response vs Frequency

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.

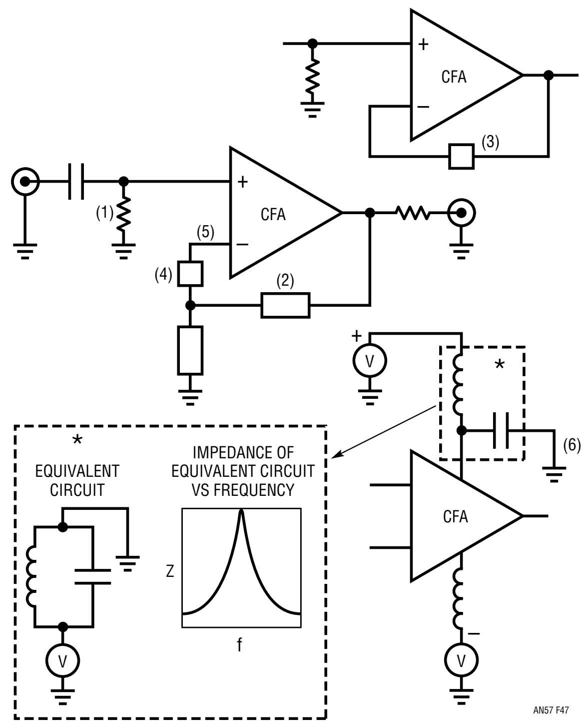


Figure 47. Examples of Misapplications

[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

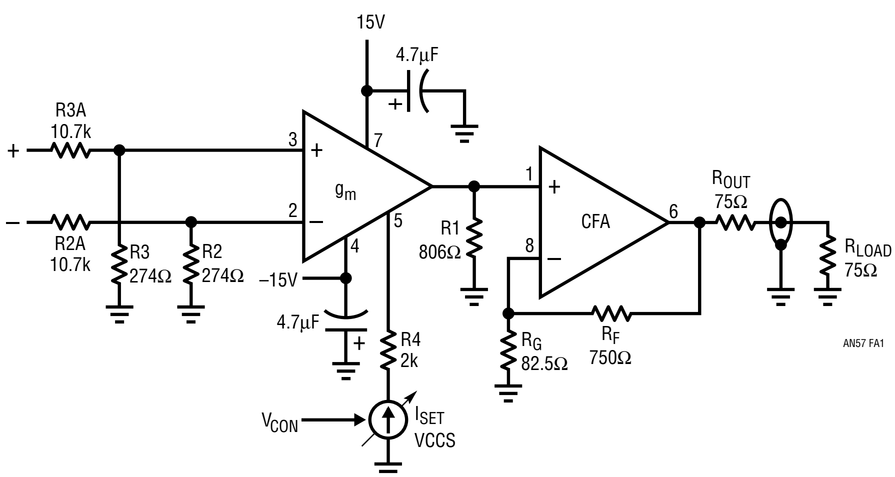


Figure A1. Differential-Input, Variable-Gain Amplifier

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.

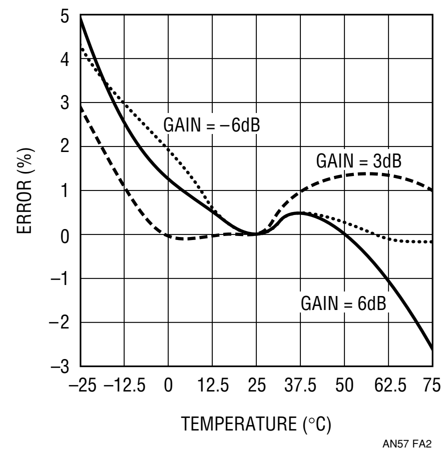


Figure A2. Gain Error for Circuit in Figure A2 Plus Temperature Compensation Circuit Shown in Figure A4 (Normalized to Gain at 25°C)

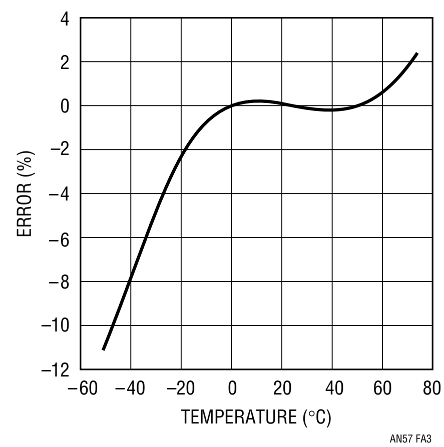


Figure A3. Thermistor Network Resistance Normalized to a Resistor with Exact –0.33%/°C Temperature Coefficient

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

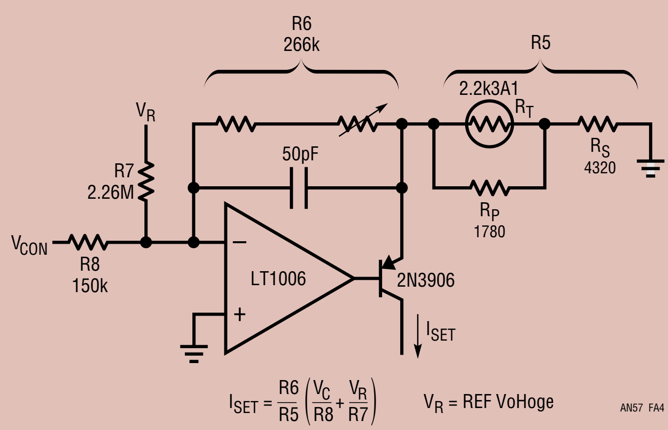


Figure A4. Voltage-Controlled Current Source (VCCS) with a Compensating Temperature Coefficient

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}.

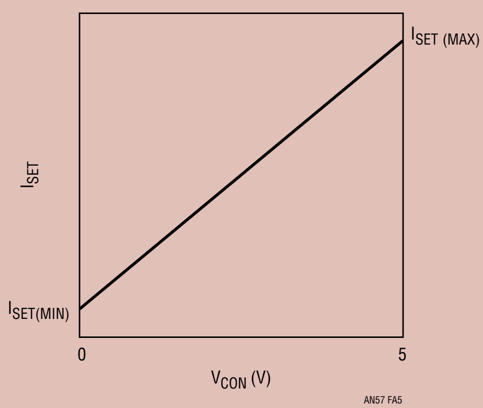


Figure A5. Voltage Control of I_{SET} with Temperature Compensation

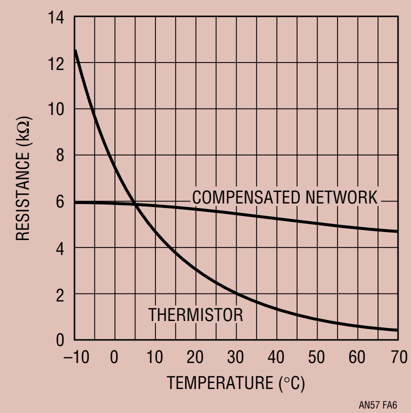


Figure A6. Thermistor and Thermistor Network Resistance vs Temperature

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).

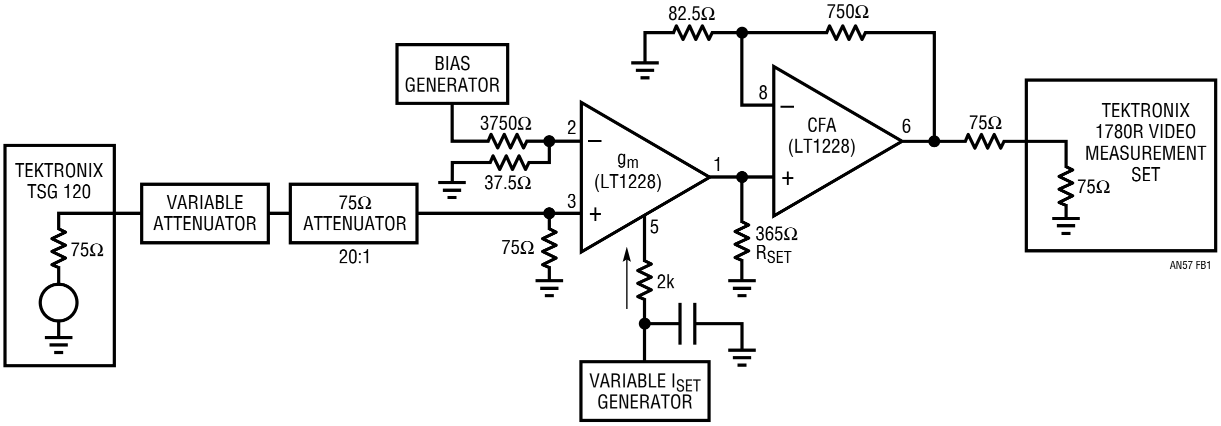


Figure B1. Schematic Diagram

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."

[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

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.

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!

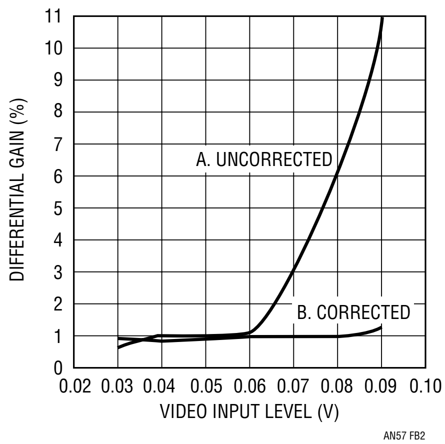


Figure B2. Differential Gain vs Input Level

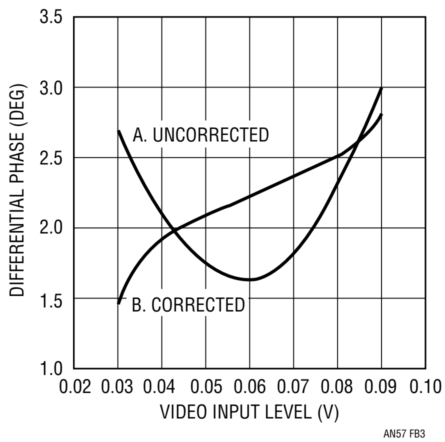


Figure B3. Differential Phase vs Input Level

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.

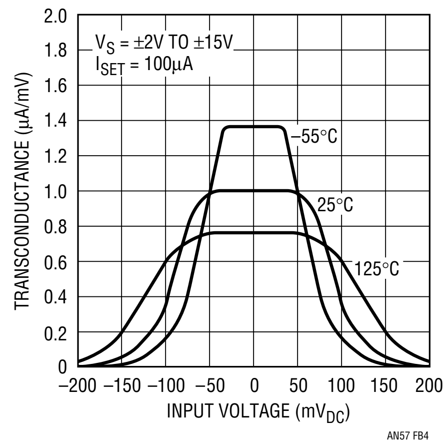


Figure B4. Small-Signal Transconductance vs DC Input Voltage

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

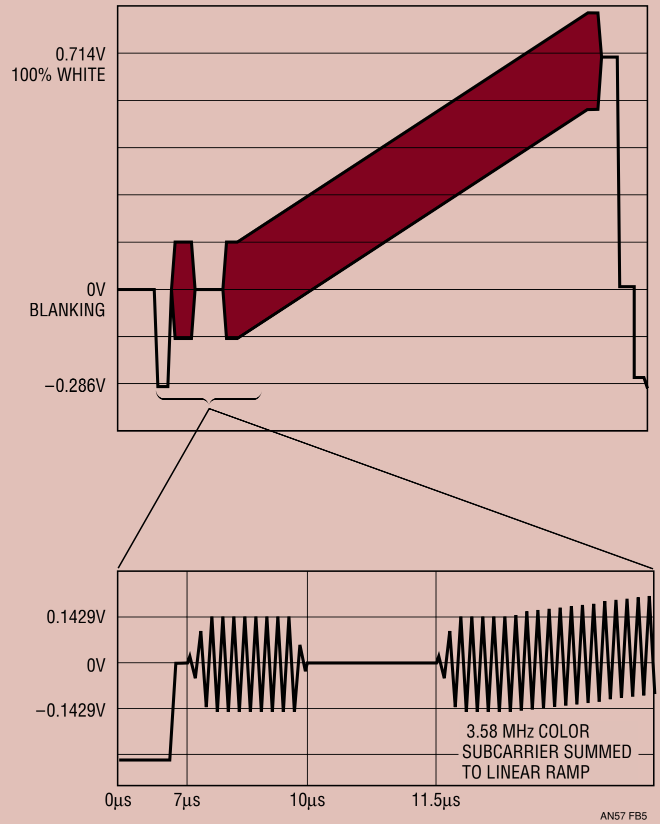


Figure B5. NTSC Test Signal

is the quality of perception which differentiates the frequency of the color, red from green, yellow-green from yellow, and so forth.

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.

[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.

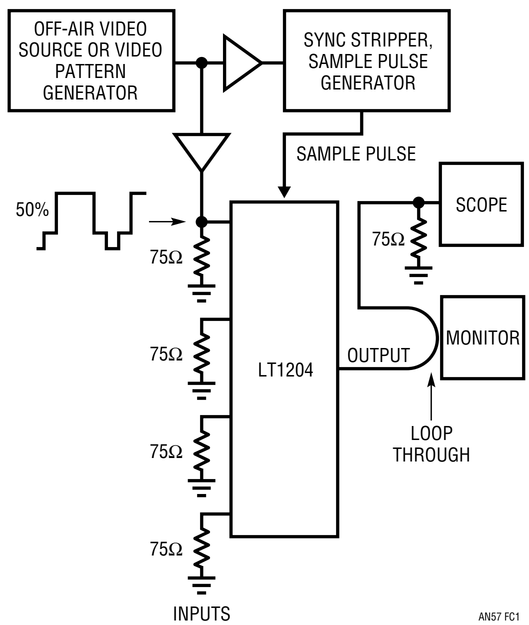
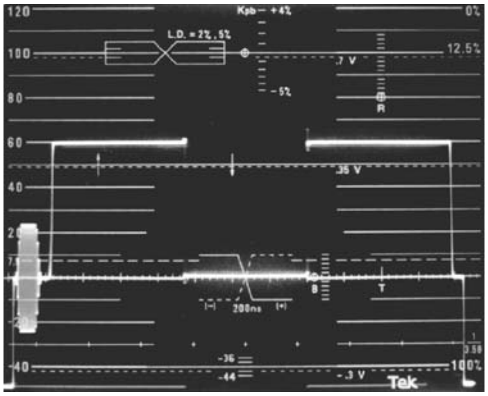
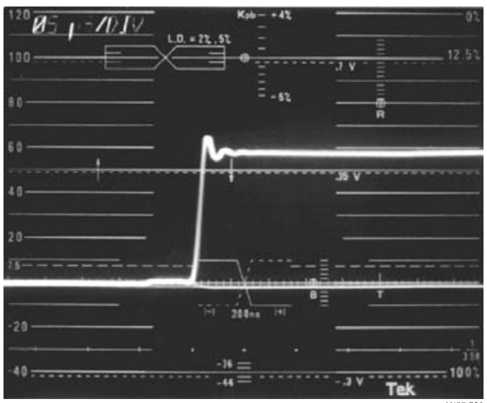


Figure C1. "Picture-in-Picture" Test Setup



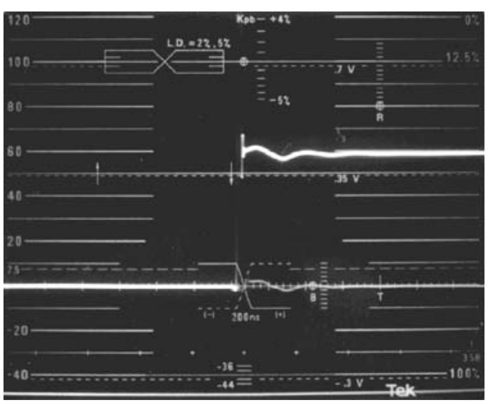
AN57 FC2

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)



AN57 FC4

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

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.

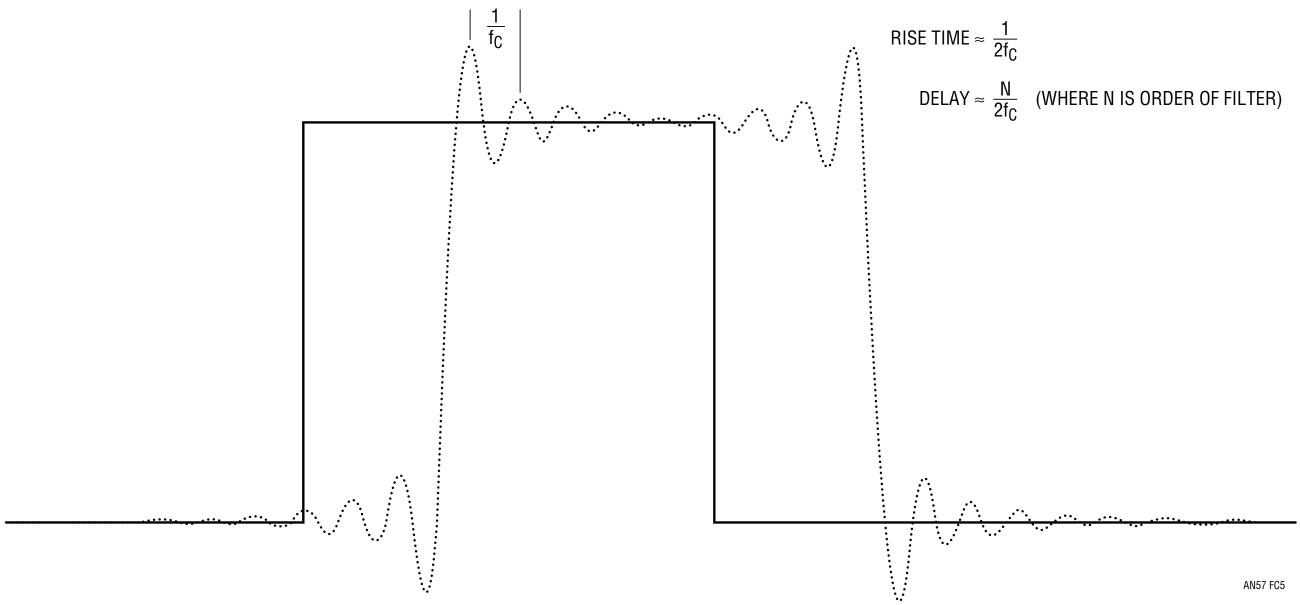


Figure C5. Pulse Response of an Ideal Sharp-Cutoff Filter at Frequency f_C

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.

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