TEXAS INSTRUMENTS

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Measuring Differential Gain and Phase

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ABSTRACT

Standard video signals are based on a system developed in the 1950's. The colors and brightness we see on a television screen are encoded within an analog signal. How well an amplifier reproduces this video signal is the foundation for the measurement of differential gain and phase.

The topic of this discussion is the basic concept of differential gain and phase. This discussion will then lead to the analysis of the method used by Texas Instruments to measure differential gain and phase for the high speed amplifier product line—the THS family.

Contents

List of Figures

List of Tables

1 Introduction

Composite video contains the two parts of the video signal—the subcarrier portion and a broadband portion. The broadband portion of the composite video signal controls the brightness, or luminance, of the picture to be displayed. The subcarrier portion of the composite video signal controls the color, or chrominance, of the picture to be displayed. In turn, this color information has two parts to it: the amplitude and the phase. Together, all of these constitute what we see on our television or monitor screens. If any part of the signal is not reproduced correctly, the picture will not be displayed correctly. It could show sunny images as night scenes, or it could show the bright yellow sun as a blue sphere.

One of the most critical factors within a composite video system is how well an amplifier reproduces the composite video signal. A series of tests were conducted to verify if an operational amplifier is useful for video signals. These are called differential gain and phase tests. Even though this is a standardized specification for all manufacturers data sheets, not everyone does it the same way.

Because video information can cover a range from 0 Hz to 6 MHz or greater, the amplifier must have a frequency response of at least 50 MHz to 60 MHz. This precludes the vast majority of amplifiers, and typically leaves the high-speed amplifier product group as the only viable solution for amplifiers within a video system. It may be best to start with the definitions for differential gain and phase.

2 Differential Gain and Phase Definitions

Differential gain is the error in the amplitude of the color signal due to a change in luminance (brightness) level. Basically, the subcarrier reference signal (3.58 MHz or 4.43 MHz) is being offset by a low-frequency ramp going from 0 IRE (0 V) to 100 IRE (714 mV). This signal is then fed through an amplifier configured for a gain of +2 with a fixed load resistance (typically 150 Ω). The change in the amplitude of the subcarrier reference signal being offset by the ramp is called differential gain error. It is usually expressed as a percent error.

Differential gain is important to video signals because if the amplitude of the subcarrier signal changes with a change in luminance level, then the picture being displayed will have color saturation problems. A color with 100% saturation is commonly called a *deep* color. A deep red ball being displayed with bright lights shining on it should not turn dull red when the lights start to dim. It may turn dull red if the saturation level of the base color (red) decreases as the luminance level decreases.

Differential phase is the error in the phase of the color signal due to a change in luminance (brightness) level. The phase of the color signal is responsible for displaying the proper hue, or base color. This measurement follows the exact same procedure as the differential gain measurement, but only looks for the change in phase of the subcarrier reference signal as the offset level changes. Differential phase is usually expressed in degrees.

Differential phase is important to video signals because if the phase of the subcarrier signal changes with a change in luminance level, then the picture being displayed will have color selection problems. For example, a red ball being displayed with bright lights shining on it should not turn orange when the lights start to dim. It may turn orange if the base color information starts to add some green to the image.

Typically, the human eye cannot discern the difference of color saturation if the differential gain error is 1% or lower. Additionally, the human eye cannot discern a hue error of less than 1%. If this is true, then why would anyone want an amplifier with less than 0.5% differential gain or phase error? The reasoning is that there can be several amplifiers along the video signal path. Cascading multiple amplifiers adds both positive and negative differential gain and differential phase errors together. A typical video system will consist of the recording equipment, the transmitter, the receiver, and the display device. So, if there are 20 amplifiers in the entire video signal chain, with each one having a 0.1% differential gain and phase error, then it is probable that the total system will have a differential gain and phase error of $20 \times 0.1\%$ or 2%. This is way too high for a good system, not to mention a studio quality system.

3 Differential Gain and Phase Test Setup

Testing differential gain and phase is typically done at the same time. We are interested in the amplitude and phase changes as the offset level is shifted. Because operational amplifiers are commonly used as inverting amplifiers to sum signals together, the offset level will then become inverted. To account for this, Texas Instruments uses a modulated saw-tooth ramp, as shown in Figure 1. It consists of a subcarrier reference signal at either 3.58 MHz (NTSC) or 4.43 MHz (PAL/SECAM) with an amplitude of 40 IRE. A ±100-IRE low-frequency saw-tooth waveform is then used to perform the offset function. This is a standard test signal for both NTSC and PAL. The only slight difference is that PAL uses a 43-IRE subcarrier level instead of the 40-IRE level used by NTSC; but this difference is negligible. Additionally, 100-IRE levels were used instead of 80-IRE levels to fully test the entire operating range of the amplifier under test (AUT).

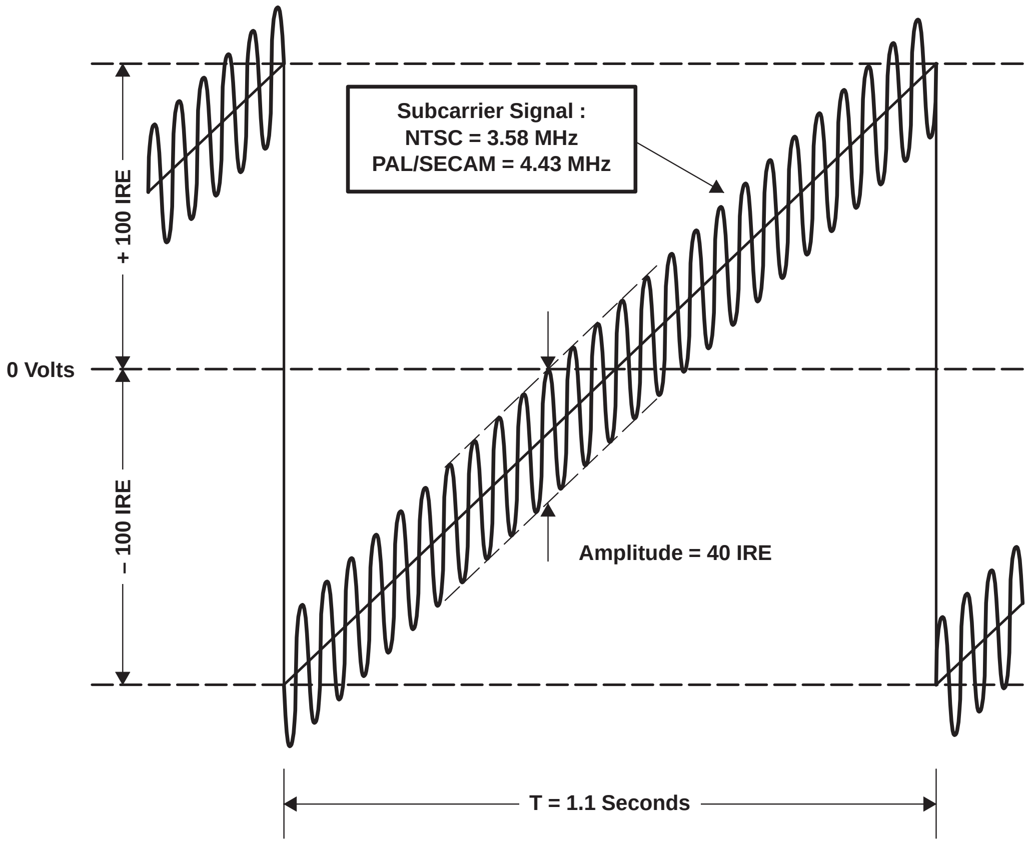

Figure 1. Differential Gain and Phase Test Signal

Some test setups use a staircase ramp instead of the linear ramp. The problem with a staircase ramp is that it does not measure every point. Differential gain and phase errors are not usually proportional to the offset level. In fact, it has been observed on numerous occasions that the maximum errors occurred somewhere between the two end points and not at the ends of the ramp. Considering the unpredictability of the amplifier, discrete steps may not give a true representation of the real differential gain and phase errors.

To accomplish the test using this waveform and to perform the analysis of the operational amplifier, an HP8753D (or E) network analyzer was chosen. An HP3325A function generator is then connected into the bias input port of the network analyzer (see Figure 2). To keep everything correctly timed, the sync output of the function generator is connected to the sync input of the network analyzer. The HP3325A generates the low-frequency saw-tooth waveform, while the HP8753D (or E) applies the high-frequency reference signal (3.58 MHz or 4.43 MHz) on top of the saw tooth. The HP3325A function generator was chosen because it is not influenced by the high-frequency signal being generated by the network analyzer. It was found that other function-generator outputs would be influenced by the subcarrier frequency and would cause the saw-tooth waveform to behave unpredictably.

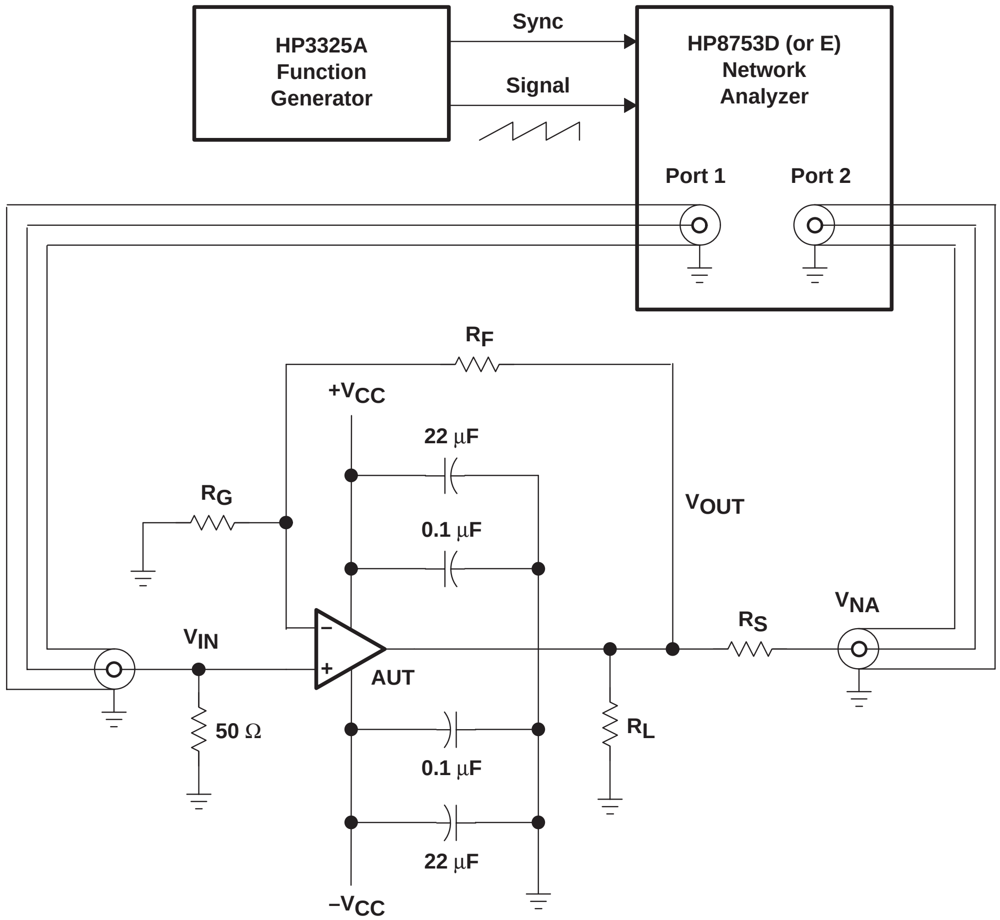

Figure 2. Differential Gain and Phase Test Setup

The AUT is usually set up with a gain of +2 ($R_F = R_G$). This is done because most video systems, not to mention most transmission-line systems, use double termination on the amplifier output. When a double termination is performed, the signal amplitude at the receiving side is reduced by a factor of two. To compensate for this, the driving amplifier is set to a gain of +2.

Although it is realistic to see an amplifier in the inverting configuration, the differential gain and phase tests are done in the noninverting configuration. There are two reasons for this: first, the noise gain of an amplifier is always referred to the noninverting input; so an amplifier set to a gain of –1 is equivalent to a gain of +2 configuration when it comes to amplifier noise gain; the second reason is that a noninverting configuration should theoretically have worse errors than an inverting configuration because the input terminals of the amplifier are held at a fixed reference point (a virtual ground for the inverting terminal). But, in the noninverting gain of +2 configuration, the amplifier’s input-circuitry voltage fluctuation is directly proportional to the applied input-signal voltage. If the voltage of the amplifier’s input circuitry changes, there is more room for errors to occur due to the reduced common-mode rejection ratio of the op-amp. For these reasons, the AUT is tested with a gain of +2 to provide the worst-case conditions for errors to occur.

Experimentation has shown that the load placed on the amplifier will affect its differential gain and phase measurements. This makes sense, since loading also affects harmonic distortion measurements. Because this is a video specification, a doubly-terminated 75-Ω load, which is the standard video termination, results in an equivalent 150-Ω load to the amplifier. A load higher than 150 Ω should never be used in testing of standard differential gain and phase. The network analyzer has a built-in 50-Ω termination to ground that is always present. So, to accomplish the 150-Ω load, a 100-Ω resistor is used for R_S. A question often asked is will the resistor divider on the output of the AUT affect the differential measurements? The answer is no. It will not affect the measurements because we are dealing with ratios of AUT signals. Since the differential gain (in dB) is found by dividing two ratios, the influence of the resistor divider cancels out.

It is common to use amplifiers to drive multiple video lines. The differential gain and phase measurements are taken for each load to evaluate the effects of driving lower-impedance loads. The equivalent resistance load presented to the AUT is simply 150 Ω divided by the number of lines being driven. To simulate this, the values for R_S and R_L are chosen as follows:

Table 1. AUT Loads for Measuring Differential Gain and Phase

NUMBER OF LINES	EQUIVALENT RESISTANCE (Ω)	R_S (Ω)	R_L (Ω)
1	150	100	–
2	75	50	300
3	50	50	100
4	37.5	50	62
6	25	50	33
8	18.75	50	23

Experience shows that resolution is the most important factor when selecting the measuring instrument to perform these tests. The most common piece of video test equipment found was the Tektronix VM700. But its resolution is only around 0.03% for differential gain, and 0.03° for differential phase. This may be acceptable for many video systems, but not for very high-speed, ultra-low distortion amplifiers.

The HP8753D (or E) was chosen for its great versatility and high resolution. The minimum resolution is 0.001 dB/division for amplitude measurements, and 0.01°/division for phase measurements. A simple mathematical calculation is used to convert from dB to percent error. The details are left to the reader.

$$\%\ Error = \left[10^{\left(\left|Gain\ Error\ (dB)\right|/20\right)} - 1\right] \times 100 \qquad (1)$$

Where

Gain Error = maximum gain reading—reading at 0 IRE (reference)

When this equation is used, the calculated gain error resolution is approximately 0.0115%/division. This is a respectable figure, but it still has some limitations. Especially when it comes to the high-quality amplifiers produce by Texas Instruments.

4 Testing Differential Gain and Phase

To properly measure differential gain and phase, the test setup should be calibrated first. This is accomplished by first adjusting both channels of the HP8753D (or E) network analyzer to the following settings:

- Sweep type = CW time
- CW frequency = 3.58 MHz (or 4.43 MHz)
- Power = –6.9 dBm (40 IRE)
- IF = 300 Hz
- Number of data points = 401
- Averaging factor = 20 (sometimes higher, depending on signal)
- Smoothing factor = 1% to 5% (depending on amplitude)
- External trigger on sweep

The HP3325A function generator is then adjusted to the following settings:

- Frequency = 0.905 Hz
- Amplitude = 1.428 V_{PP} (± 100 IRE)
- Offset = 0 V
- Function = saw tooth

Once this setup is complete, the output cable on port 1 is connected to the input cable on port 2 with a double female SMA adapter. The network analyzer is set to calibrate mode and allowed to calibrate itself. Then the AUT circuit is place in the test path as shown in Figure 2.

The tests are conducted for both the NTSC subcarrier at 3.58 MHz, and the PAL subcarrier at 4.43 MHz. Additionally, the AUT's power-supply voltages are run at ±15 V and ±5 V at each subcarrier frequency. All of these tests are then repeated using the equivalent loads specified in Table 1. The network analyzer is recalibrated whenever the subcarrier frequency is changed to ensure reliable measurement results.

Once all of the data is taken, the differential gain and phase numbers are extracted. Because differential gain and phase are measurements of error caused by a change in offset voltage, Texas Instruments looks only for the largest change along each of the ±100 IRE signals with the reference signal level at 0 IRE (0 V). The largest difference on gain may be on the +100-IRE magnitude, and on the –100-IRE magnitude for phase. It does not matter where it is, we are simply looking for the worst-case condition. The differential gain error is calculated by using formula (1) for the absolute-maximum difference in gain signal. The differential-phase error is calculated by simply subtracting the absolute-maximum difference from the reference value.

Depending on the signal coming through the setup, the number of averages may need to be increased. This is done specifically when the network analyzer is at its highest resolution settings (0.001 dB and 0.01°). The signal may be very choppy, and increasing the averages will tend to average out random noise from the system.

For very low distortion amplifiers driving a 150-Ω load, the error signal may still be too hard to see. This is where the smoothing function plays a role. Smoothing computes a displayed data point based on a moving average of several adjacent points. Its function is to reduce relatively-small peak-to-peak noise values in the measured data. The minimum smoothing value is used as much as possible in order to provide accurate results. The drawback to using the smooth function is that the values at the very beginning and the very end of the time-base measurement results may be incorrect. To compensate for this, TI ignores the first and last divisions on the time base results. Since these divisions are ignored, the function-generator frequency is selected so that the offset ramp signal is at –100 IRE (–0.714 V) at 0 seconds, crosses the 0 IRE (0 V) reference point at 0.55 seconds, and is at +100 IRE (+0.714 V) at 1.1 seconds (see Figure 3). The ramp starts at –100 IRE again and continues to rise while the data is still coming through. This compensates for the fact that for the first 0.14 seconds, the data due to possible smoothing errors is ignored. The data from 1.1 to 1.24 seconds is used instead of the data for the first 0.14 seconds. Keeping the smoothing value to 5% or below ensures that no errors are introduced into the readings.

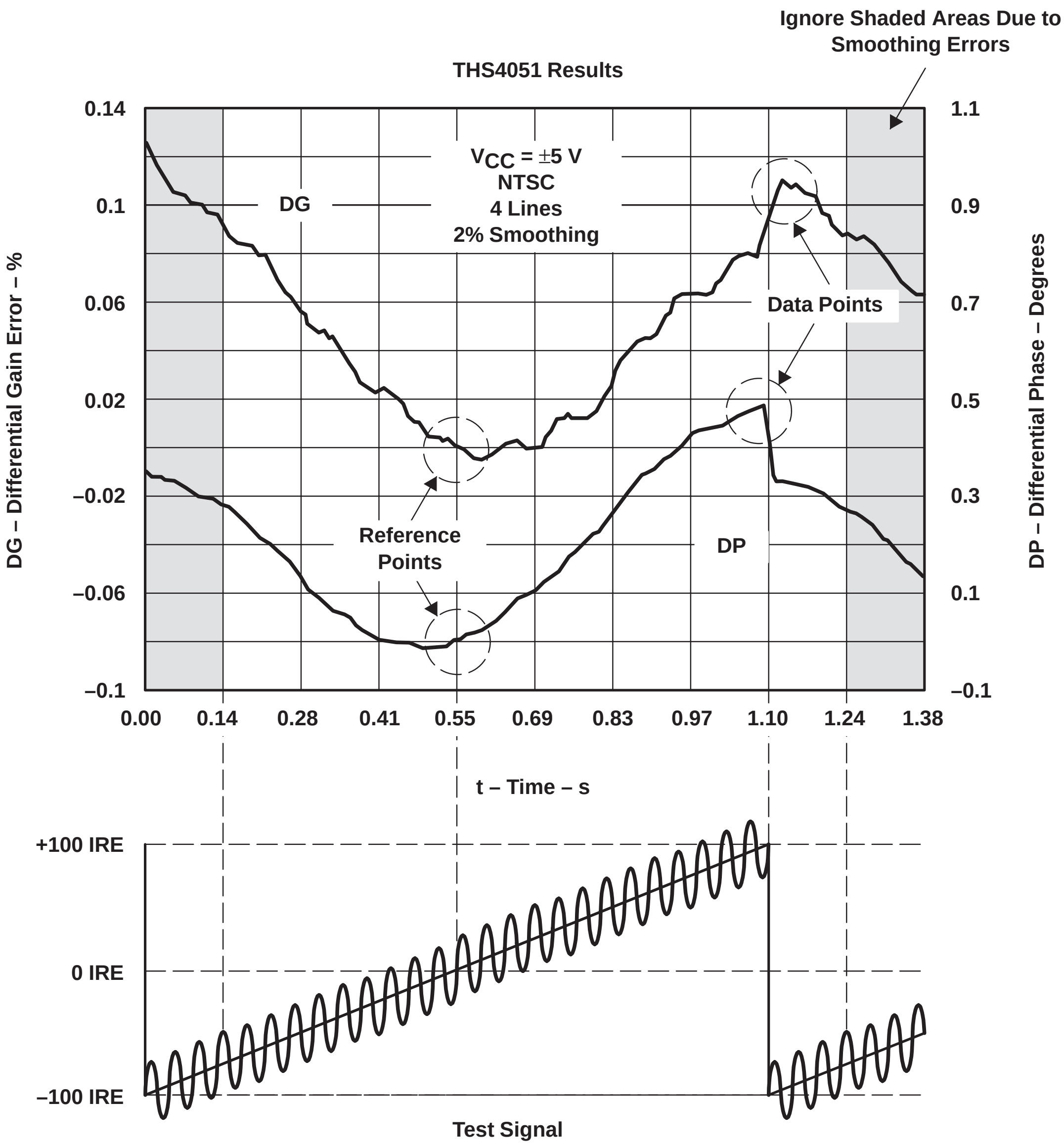

Figure 3. Actual Differential Gain and Phase Results (4 lines = 37.5 Ω)

Figure 3 shows the data points used to find the actual differential gain and phase numbers. These include the 0-IRE reference point and the maximum deviation from each reference point, as indicated in the graph. This graph shows that the worst-case differential gain occurred at –100 IRE (0.11%), and the worst-case differential phase occurred at +100 IRE (0.5°). It does not matter where the maximum error point is relative to the IRE level; only the maximum error relative to the reference points is relevant. For example, the same amplifier was tested with only 2 lines (75-Ω equivalent load). The maximum differential-gain error just happened to occur at +100 IRE (0.02%), and the maximum differential phase error occurred at +100 IRE (0.075°).

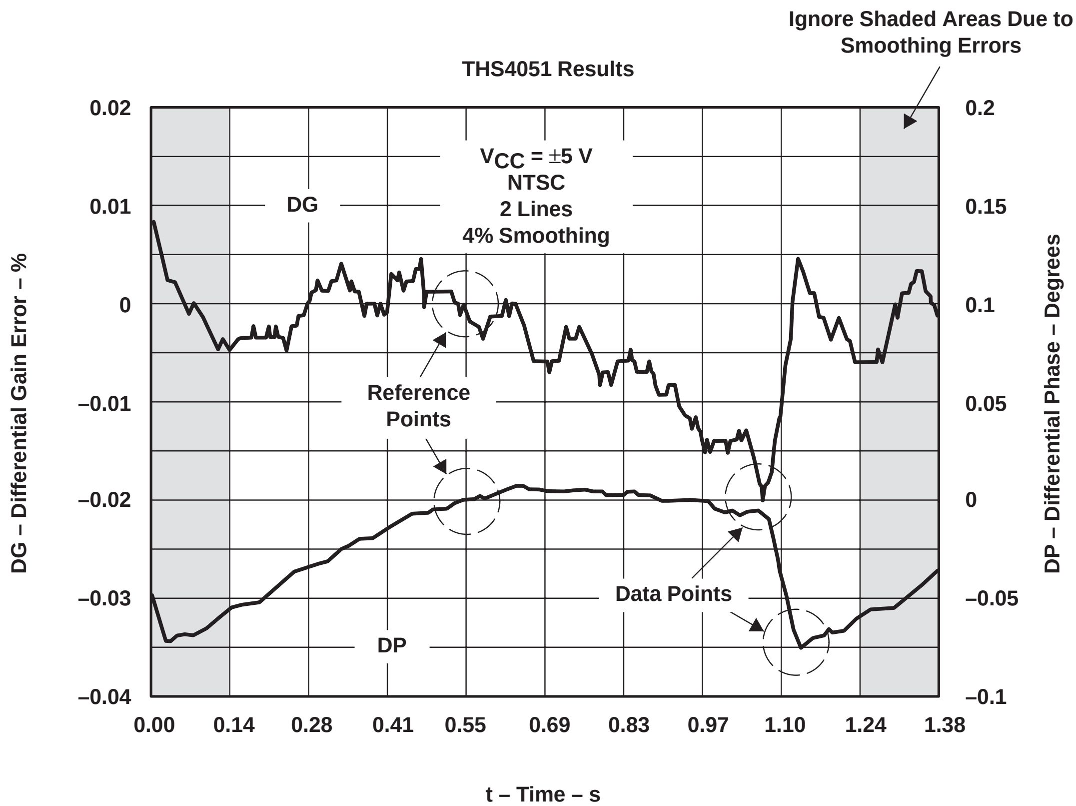

Figure 4. Actual Differential Gain and Phase Results (2 lines = 75 Ω)

There are two network-analyzer settings that may need additional explanation—the 401 data points and the 300-Hz IF bandwidth. The number of data points was selected so that enough points were taken to allow the smoothing factor to average out any random noise within the entire system. If enough data is collected, then random noise should average to a specific level. This level is subtracted from the AUT measurement by the network analyzer after calibration. This subtraction factor should become more consistent from test-to-test as the number of data points is increased. Misleading results will occur if the number of data points is too small.

The 300-Hz IF bandwidth was selected to allow the measurement of extremely small errors. Lowering the IF bandwidth from the network analyzer's default setting of 3,000 Hz will lower the noise floor of the system by about 10 dB. Additionally, lowering the IF bandwidth is better than averaging because it filters out spurious responses, odd harmonics, higher-frequency spectral noise, and line-related noise. When we are looking at operational amplifiers with differential gain and phase errors better than 0.01% and 0.01°, respectively, any reduction in system noise is extremely valuable. The drawback of lowering the IF bandwidth is a dramatic increase in the amount of time required to make one sweep. But this is a small price to pay for good results.

The base-line data in Figure 5 show the actual results after calibrating the system and bypassing the AUT. This was taken with 5% smoothing and 25 averages. The maximum differential gain error shown is 0.0034%, and the maximum differential phase error is 0.0027°.

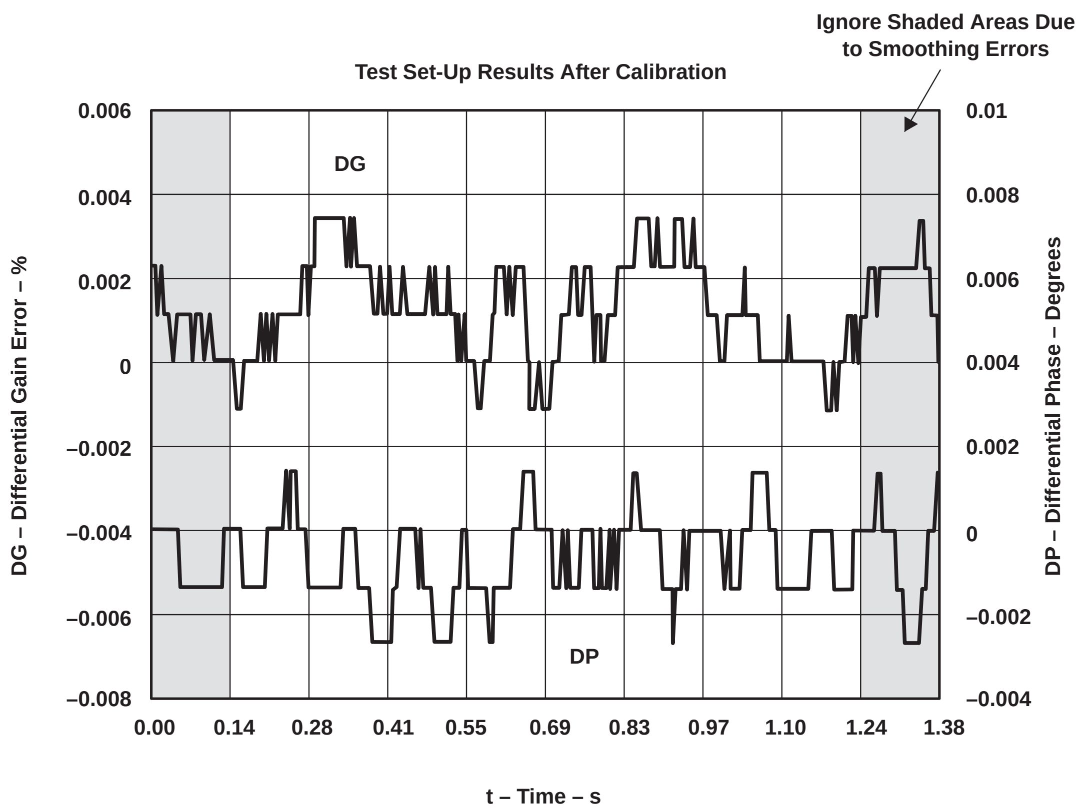

Figure 5. Base-Line Results After System Calibration

Due to the random nature of these values and the amount of uncertainty within the network analyzer, Texas Instruments will not typically publish numbers less than 0.01% and 0.01°. Even though it is quite possible for an amplifier to exhibit better results than these, it would be extremely difficult to prove time-and-time again that the errors are better than 0.01% and 0.01°. Until a new measuring device of higher precision is used, the limits previously shown shall remain a constraint.

5 Summary

The measurement techniques used at Texas Instruments are based on sound engineering methodology. The worst-case errors are used to obtain the values published in any Texas Instruments amplifier data sheet. This is the kind of data any designer should look for when selecting the proper amplifier for a system. Be sure to look at all the test conditions before making any decision on a particular video amplifier. This includes NTSC or PAL subcarrier reference at the 40-IRE amplitude, both +100-IRE and –100-IRE offset tests, proper power supply voltages for the amplifier, amplifier gain, and the load placed on the amplifier's output. Failure to do so may result in unpredictable, and possibly unacceptable, system performance.

Appendix A A Primer for the Composite Video Signal

A.1 Definitions

It is often helpful to understand the composite video signal in its entirety. The National Television System Committee (NTSC) developed this structure in 1953 with approval by the Federal Communications Commission (FCC) in 1954. The NTSC stipulated that the color composite video signal must occupy the same bandwidth as the existing monochrome video signal (for use with black-and-white television sets). Because of this constraint, the video signal (with or without color information) must use the same 0 to 4.2-MHz frequency range. The reasoning is quite simple: it allows both black-and-white and color television sets to use the exact same video signal, while the audio information uses the 4.5-MHz subcarrier.

Another video standard is the phase alternation line (PAL) system. Broadcast of the PAL system in Europe began in 1967. One of the more notable differences was the NTSC use of a 525-line, 60-fields/second, 2:1-interlaced system, while PAL uses a 625-line, 50-fields/second, 2:1-interlaced system. The other big difference that PAL attempted to overcome was the NTSC requirement for very high-quality components in-order to reproduce a video signal with high repeatability. To accomplish this, PAL uses a line-by-line phase reversal of one of the color signal components. The human eye will then hopefully average the potential differences in color lines to reproduce a better picture than NTSC. Despite these differences, both NTSC and PAL have very similar composite video signal wave-shapes and attributes. This primer will concentrate on the NTSC composite video signal for simplicity sake.

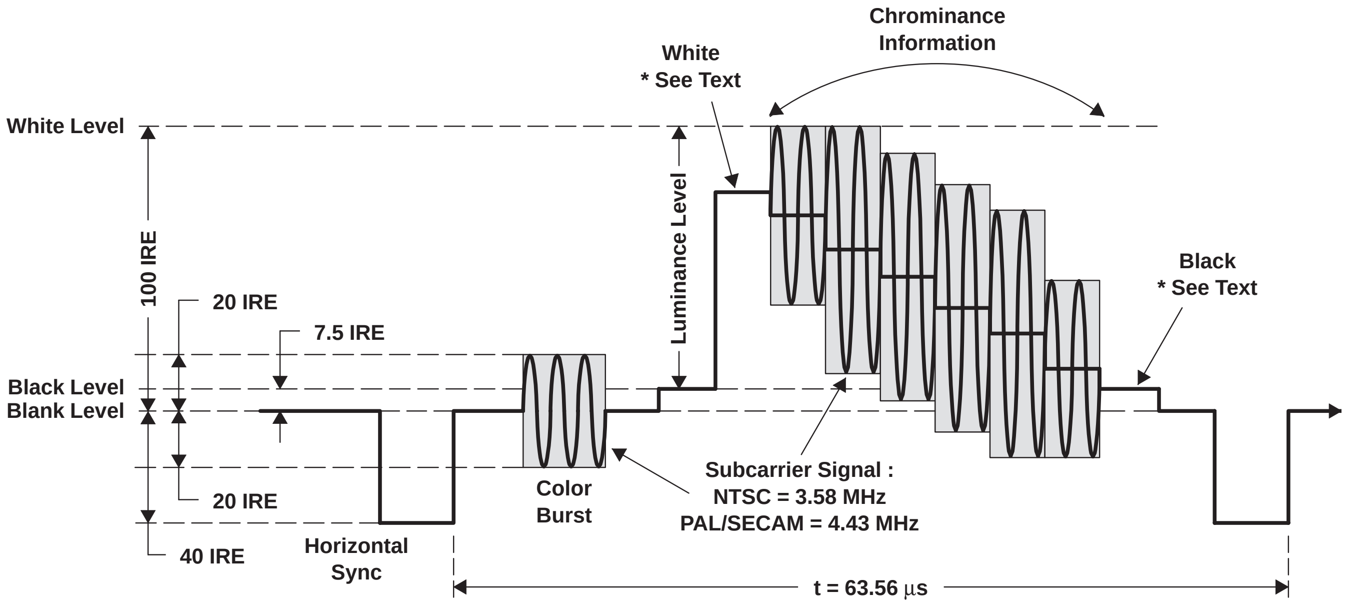

Figure A–1. A Typical 75%-Amplitude, 100%-Saturation Composite Video Signal

The composite video signal includes timing (sync), brightness (luminance), color (chrominance), and color burst information—see Figure A–1. The horizontal-sync pulse is used to keep everything within a video system at the right point in time. It tells the display device where the beginning of a new scan line is. A new scan line is generated every 63.556 µs (15.734 kHz). This concept is very similar to the use of a digital-clock in logic circuitry to keep all the signals at a common reference point in time.

The brightness information is the luminance portion of the composite video signal. Because the human eye is more sensitive to brightness variations, the entire bandwidth (0 to 4.2 MHz) is allocated to reproducing the luminance information. The brighter (or whiter) the picture is, the higher the video-signal amplitude. Video signals use the arbitrary IRE unit for amplitude measurements. A pure-white signal corresponds to a 100-IRE level, while the blanking level, which is blacker than black, corresponds to a 0-IRE level. Sometimes it is required to relate the IRE measurement to volts. The NTSC standard states that 100 IRE should correspond to 714 ±7 mV, while the PAL standard establishes it at 700 ± 7 mV. This difference is generally ignored to keep things simple. The most common conversion used is 1 V_{PP} = 140 IRE, or 1 IRE = 7.14 mV.

The color information (chrominance) is encoded within a subcarrier frequency. For most systems, this subcarrier frequency is 3.579545 MHz ±10 Hz for the NTSC standard, and 4.43361875 MHz ±5 Hz for the PAL standard. There are two parts to the color information: hue and saturation. Hue is technically the wavelength of a color. It can be thought of as the base color within the whole color spectrum. The other part to the color information, saturation, determines the intensity of a specific color. For example, the color red may look very dull with very little saturation, or it may look deep red with very high saturation. The fact that its base color is red still remains.

If the luminance portion takes up the full 0 to 4.2 MHz band, then what is the bandwidth for the color portion? To help answer this, we need to know how the color subcarrier came to be. The first thing to realize is that the NTSC color subcarrier was chosen for a reason. The audio carrier is at 4.5 MHz. The 286th harmonic of the already existing 15.75-kHz monochrome horizontal-scan frequency is about 4.5 MHz. In the color system, this 4.5-MHz frequency is divided by 286 to obtain the new horizontal-scan frequency of 15.743 kHz. This result is within the deviation limit originally set by the NTSC monochrome standards. The luminance information is then based on this 15.743-kHz separation. To minimize the visibility of the color subcarrier, its frequency is chosen to be an odd multiple of one-half the horizontal scan rate. This then performs frequency interleaving (see Figure A–2). The color information is also allowed to have a 0.6 MHz side band. This leaves a maximum upper frequency of 4.2 MHz (from the luminance bandwidth limit)—0.6 MHz = 3.6 MHz. It was found that the 455th harmonic ($5 \times 7 \times 13 = 455$) of one-half the color scan rate (15.743 kHz/2) is equal to 3.579545 MHz.

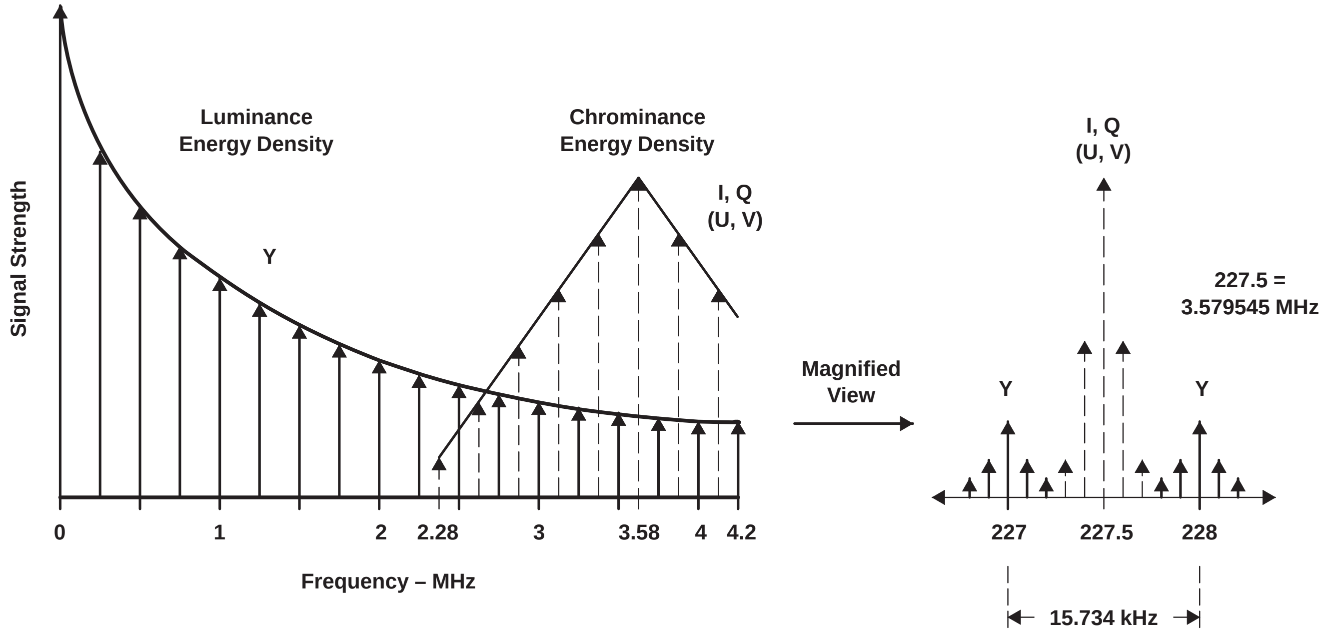

Figure A–2. Frequency Interleaving Spectrum

Let's now find the color information bandwidth. As stated earlier, the color information is allowed a 0.6-MHz side band. All three color components are allowed to use the entire bandwidth for large objects. For medium-sized objects a 1.3-MHz bandwidth is allowed, but is generally limited to two colors. The choice of these colors is discussed further in this document. A filter is placed at 4.2 MHz to insure that the 1.3-MHz upper-side band does not interfere with the sound carrier. Most consumer-grade equipment uses only the 0.6-MHz side bands, while studio quality equipment generally use the fully-allotted spectrum (see Figure A–3).

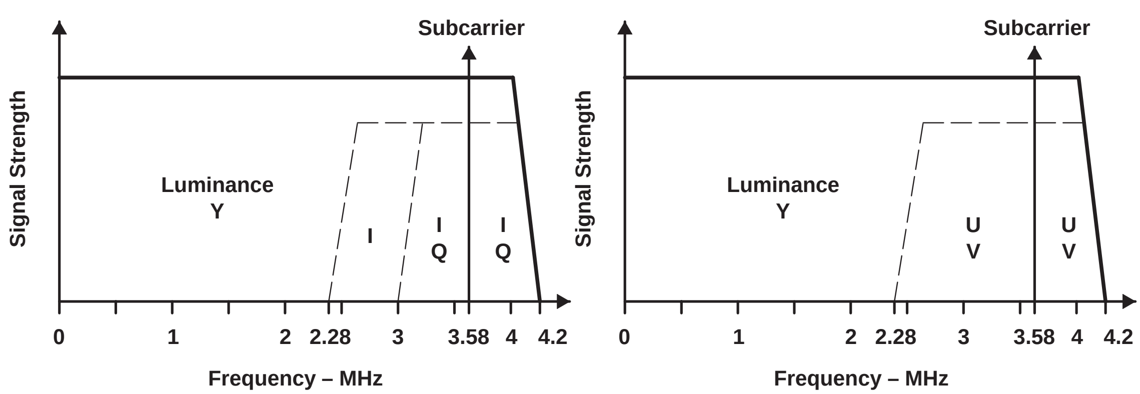

Figure A–3. NTSC Composite Video Frequency Spectrum

The last part to a composite-video signal is the color burst. The color burst tells the video circuit color decoder how to decode the color information contained within the following line of video. It ensures that the colors displayed match those of the original source material. This is accomplished by sending 8 to 11 subcarrier cycles. These cycles are sent in phase with the subcarrier used to record the original picture. The receiving circuitry can then lock this frequency and phase for its own subcarrier oscillator. Because the color burst is only transmitted with color material, the decoding circuitry can tell if the composite video signal is black-and-white or color and adjust its compensation accordingly.

A.2 Creating the Composite Video Signal

Now that we know what the composite-video signal components are, let us see how the picture information is encoded. If the signal is purely monochrome, the video signal only has the luminance component and no color subcarrier component. This is quite straightforward and should be easy to see how a monochrome signal is encoded. The tricky part is getting the hue and saturation components into the composite video signal.

The first and most important part of the composite video signal is the luminance information. The luminance (or brightness) is derived from the gamma-corrected red (R’), green (G’), and blue (B’) signals. Gamma correction takes into account the fact that the intensity of the display device (typically a cathode-ray tube—CRT) is proportional to the signal voltage raised to some power (typically 2.2 to 2.6). As a result of this relationship, high-intensity images are even brighter, and low-intensity ones become even darker. To correct for this, the red, green, and blue (RGB) transmitted signal levels are reduced by a factor of 1/2.2 = 0.45. An advantage of this correction is an increase in the signal-to-noise ratio during transmission. For clarification purposes, the component signals are shown below.

$$R_{DISPLAYED} = \left(R_{SIGNAL}\right)^{2.2} \text{ and } R_{TRANSMIT} = \left(R_{SIGNAL}\right)^{0.45} = R' \quad (2)$$

$$G_{DISPLAYED} = \left(G_{SIGNAL}\right)^{2.2} \text{ and } G_{TRANSMIT} = \left(G_{SIGNAL}\right)^{0.45} = G' \quad (3)$$

$$B_{DISPLAYED} = \left(B_{SIGNAL}\right)^{2.2} \text{ and } B_{TRANSMIT} = \left(B_{SIGNAL}\right)^{0.45} = B' \quad (4)$$

When talking about transmitted color components, the gamma-corrected color components (R’G’B’) are used for all of the signal levels in the composite-video signal.

The luminance information (or brightness) is formed by the basic R’G’B’ color combination. When these signals are set to zero, the luminance information portrays a black color. When the three signals are set to 100% level, the luminance information portrays a white color. It then becomes clear that every other color falling between black and white can be encoded. This allows a black-and-white television set to reproduce an image created from a color source.

The formula for luminance is:

$$Y\ (luminance) = 0.299\ R' + 0.587\ G' + 0.114\ B' \quad (5)$$

We can also see that color weighting is used for the transmission of each color. This is because the human eye is more sensitive to green than to red, and more sensitive to red than to blue. It makes sense to have the dominant portion of the signal be the color that our eyes are most sensitive to. This naturally brings us to how the rest of the color information is encoded.

The hue and saturation information is encoded by using a color-difference formula. This is done by running the luminance (Y) information through an inverter to obtain –Y. This signal is then added to the R’ and B’ color signals to get:

$$R' - Y = 0.701\ R' + 0.587\ G' + 0.114\ B' \quad (6)$$

$$B' - Y = 0.299\ R' + 0.587\ G' + 0.886\ B' \quad (7)$$

But what about the third color-difference signal, G' – Y? Because the luminance signal incorporates the G' component, sending the G' – Y portion is not required. At the receiver side, to extract the original R'G'B' components, we simply do the following:

$$(R' - Y) + Y = R' \tag{8}$$

$$(B' - Y) + Y = B' \tag{9}$$

$$(Y - R' - B') \times 1.704 = G' \tag{10}$$

or

$$Y - 0.510\ (R' - Y) - 0.194\ (B' - Y) = G' \tag{11}$$

Equations (5) through (11) show that it is only necessary to send two parts of the R'G'B' signal instead of all three. To simplify equations (6) and (7) even further, the color-difference components U and V were created

$$U = 0.492\ (B' - Y) \tag{12}$$

$$V = 0.877\ (R' - Y) \tag{13}$$

To get back to the R'G'B' discrete signals using the U and V methodology, use the following:

$$R' = Y + 1.140\ V \tag{14}$$

$$B' = Y + 2.032\ U \tag{15}$$

$$G' = Y - 0.394\ U - 0.581\ V \tag{16}$$

It is also common to see the I and Q components used in place of U and V. This is done because when medium-size objects are displayed on a screen, the eye is more sensitive to certain colors. These colors are the bluish-greens and the reddish-oranges. Medium-size objects use the full 1.3-MHz lower-side band (LSB) of the subcarrier energy distribution (see Figure A–3). However, the Q signal is filtered to show only the 0.6-MHz LSB portion. So, for medium-size objects, the Q-signal amplitude is reduced to zero. On the other hand, the I signal is allowed to use the full 1.3-MHz LSB spectrum. When only the I signal is used, just the sensitive color range from the bluish-greens to the reddish-oranges is shown (see Figure A–4). Hence, the I and Q scheme was created to exploit the sensitivity of the human eye beyond what the U and V formulas provide.

$$I = 0.596\ R' - 0.275\ G' - 0.321\ B' \tag{17}$$

$$= V\ cos\ 33^\circ - U\ sin\ 33^\circ \tag{18}$$

$$= 0.736\ (R' - Y) - 0.268\ (B' - Y) \tag{19}$$

and

$$Q = 0.212\ R' - 0.523\ G' - 0.311\ B' \tag{20}$$

$$= V\ sin\ 33^\circ - U\ cos\ 33^\circ \tag{21}$$

$$= 0.478\ (R' - Y) + 0.413\ (B' - Y) \tag{22}$$

To get back to the R'G'B' discrete signals using the I and Q methodology, we perform the following:

$$R' = Y + 0.956\ I + 0.620\ Q \tag{23}$$

$$B' = Y - 1.108\ I + 1.705\ Q \tag{24}$$

$$G' = Y - 0.272\ I - 0.647\ Q \tag{25}$$

The I and Q (or U and V) signals are then used to modulate the subcarrier frequency. Using a phase-quadrature mixing scheme produces the chrominance signal. Modulating one signal at a sine phase and adding the other signal, which is modulated with a cosine phase, results in the chrominance-signal formulas:

$$Chrominance\ signal = Q\ sin\ (\omega t + 33°) + I\ cos\ (\omega t + 33°) \tag{26}$$

or

$$Chrominance\ signal = U\ sin\ \omega t + V\ cos\ \omega t \tag{27}$$

Where $\omega = 2\,\pi\,F_{SUBCARRIER}$

The hue information is introduced into the signal by a phase shift relative to the subcarrier reference signal, as shown in Equations 26 and 27. For example, the color red is at a phase-shift of 103° relative to the sub-carrier, green is at 241° relative to the subcarrier, and blue is 347° relative to the subcarrier. By setting the correct phase relative to the subcarrier reference signal, any color can be realized on a vector diagram (see Figure A–4). The only thing left to produce the proper color is to find the proper color saturation.

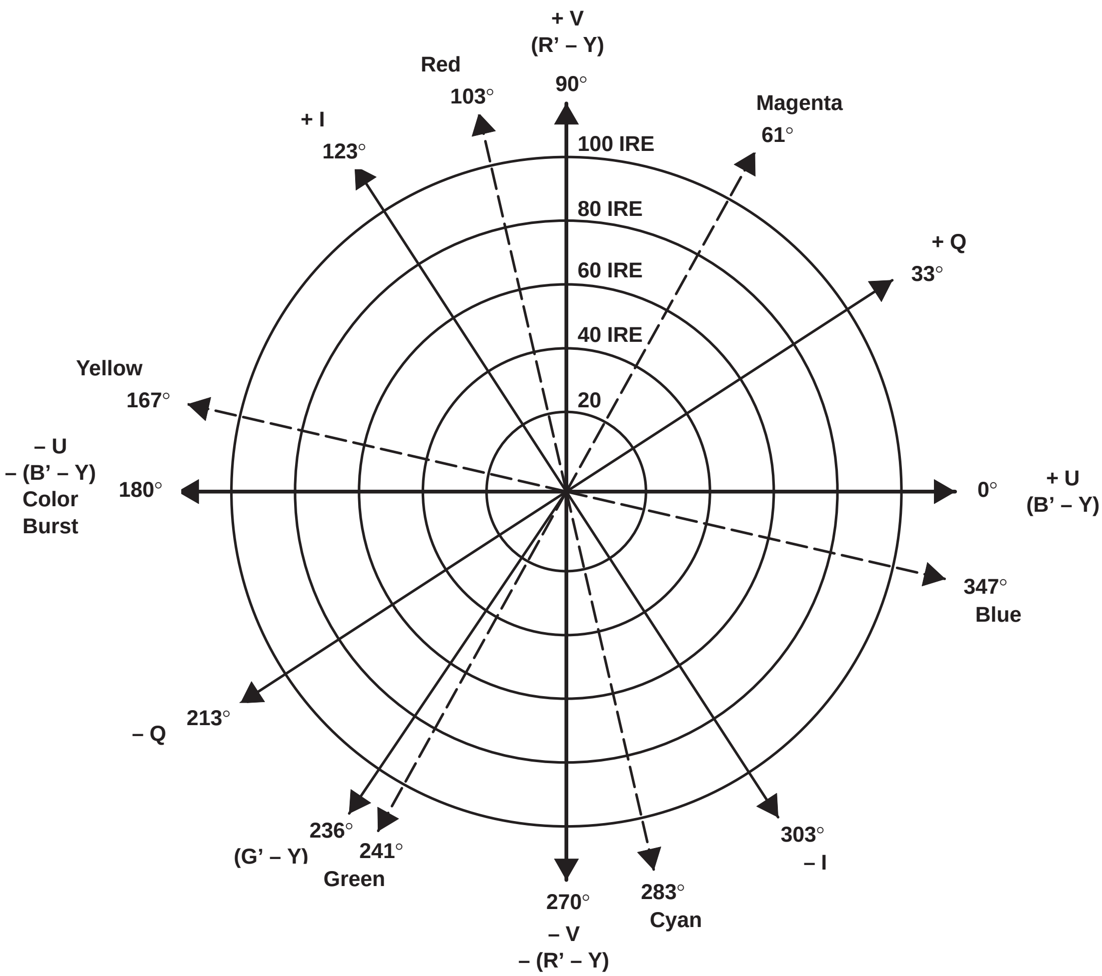

Figure A–4. Vector Diagram

The saturation information is introduced by the ratio of the chrominance amplitude (both hue and saturation) and the corresponding luminance level (see Figure A–5). The chrominance amplitude is easily obtained from the previous formulas by simply using the square root of the sum-of-the-squares method:

$$Chrominance\ amplitude = \sqrt{\left(I^2 + Q^2\right)} = \sqrt{\left(U^2 + V^2\right)} \quad (28)$$

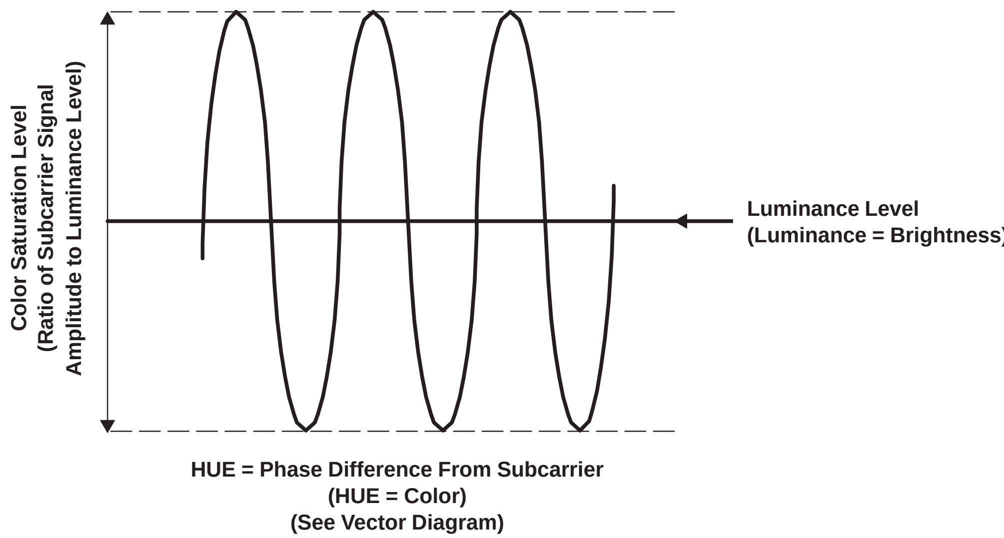

Figure A–5. Chrominance Signal Components

This chrominance signal is then added to the luminance (Y) signal to create the composite-video signal shown in Figure A–1. The additional portions to the composite-video signal are added by different means generally not related to the encoding algorithm. These portions include, but are not limited to, the horizontal sync, color burst, and blanking signals. The exact formula for the composite-video signal is:

$$Chrominance\ NTSC = Y + Q\ sin\ (\omega t + 33°) + I\ cos\ (\omega t + 33°) \tag{29}$$

or

$$Chrominance\ NTSC = Y + U\ sin\ \omega t + V\ cos\ \omega t \tag{30}$$

It may be helpful to note that if no color is encoded, there will be no subcarrier modulation within the composite-video signal. This can be seen in the example composite-video line shown in Figure A–1. The first section within the video information portion shows a flat-luminance line, coupled with no subcarrier amplitude and a high IRE level. This will produce the color white on a screen. Additionally, the last portion shown within the video information with a 7.5 IRE level will show the color black on the screen.

To see all of these concepts and equations in action, let us look at the standard test pattern shown in A of Figure A–6. This represents a single horizontal line being scanned from left to right across the screen. It consists of every combination of the three basic colors; red, green, and blue. B, C, and D of Figure A–6 show the relative saturation levels of the three basic colors using 100% as the maximum saturation level. The luminance formula of Equation 5 using the saturation levels shown results in E of Figure A–6. The next step is to create the I and Q signals using Equations 17 and 20, as shown in F and G of Figure A–6. Using Equations 26 and 27 results in the 3.58-MHz modulated-subcarrier waveform shown in H and I of Figure A–6. Next, combine the chrominance signal (H and I) with the luminance signal (E). Before the final composite waveform is produced, we should take into account that the black level is at 7.5 IRE and not at 0 IRE as mandated by the NTSC. A small correction factor has to be added to account for the 92.5 IRE full scale instead of a 100 IRE full scale. The resulting corrected composite waveform is shown in J of Figure A–6.

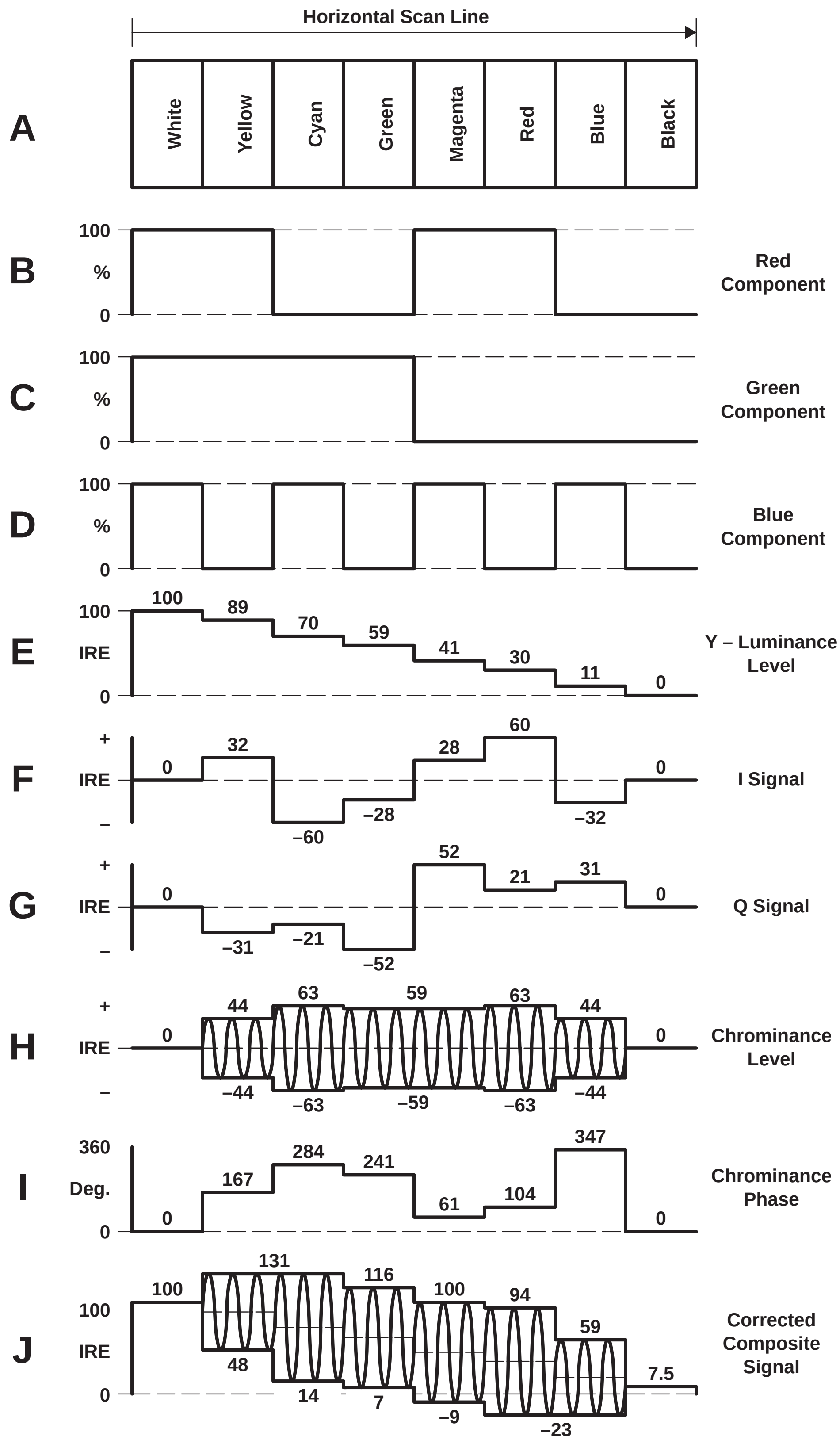

Figure A–6. The Construction of an NTSC Composite Video Signal

There are two standard test-bar patterns used for the NTSC signal. The two types are the 75% amplitude-100% saturation, which is most commonly used, and the 100% amplitude-100% saturation shown in Figure A–6. The *amplitude* number refers to the IRE level required for the color white. The luminance and chrominance levels for the two basic types of color bar formats are listed in Tables A–1 and A–2.

Table A–1. 75% Amplitude, 100% Saturation NTSC Color Bars

COLOR	LUMINANCE (IRE)	CHROMINANCE LEVEL (IRE)	CHROMINANCE PHASE (degrees)
White	76.9	0	0
Yellow	69.0	62.1	167.1
Cyan	56.1	87.7	283.5
Green	48.2	81.9	240.7
Magenta	36.2	81.9	60.7
Red	28.2	87.7	103.5
Blue	15.4	62.1	347.1
Black	7.5	0	0

Table A–2. 100% Amplitude, 100% Saturation NTSC Color Bars

COLOR	LUMINANCE (IRE)	CHROMINANCE LEVEL (IRE)	CHROMINANCE PHASE (degrees)
White	100.0	0	0
Yellow	89.5	82.8	167.1
Cyan	72.3	117.0	283.5
Green	61.8	109.2	240.7
Magenta	45.7	109.2	60.7
Red	35.2	117.0	103.5
Blue	18.0	82.8	347.1
Black	7.5	0	0

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