Designing Video Circuits

Part One of Three

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With the advent of low-cost VLSI components, designers are placing greater emphasis on incorporating images into computer-based systems. Many of their designs will use existing broadcast video standards to capture, process, and display images. Already, PC-based frame grabbers and image processors are adopting these standards.

Designers must grasp the underlying theory behind the video signal itself, the number of standards in existence, and the different methods of pre- and postprocessing. This article, the first in a three-part series, describes the video signal as well as several methods of preprocessing. Part 2, scheduled to appear next month, will discuss the several types of sampling standards, as well as video synchronization and genlock. The final installment will describe video output from digital systems, along with time-base correction and frame synchronization.

What Is Video?

Figure 1a shows the signal for one scan of a horizontal line of video with negative sync polarity, the 1V pk-pk composite color standard of video equipment. In this standard, black level is more negative than the active video. The horizontal period, which is approximately 63.6 msec, can be divided into two sections: active video and blanking.

During the active video time, the video level controls the CRT's electron guns, causing them to produce variations in color and brightness. During the blanking interval, the level of the video signal turns off the electron guns, enabling the beam to retrace without drawing to the screen (**Figure 1b**). H-sync and color burst are processed by the receiver to provide synchronization information.

Video images are created by scanning an electron beam back and forth across the face of the CRT. The left-to-right positioning of the CRT beam is synchronized to the signal produced by the studio camera. Such timing information is provided by adding an H-sync pulse to the video signal. Receivers lock onto this signal and synchronize the horizontal sweep timing to the H-sync pulse. A similar scheme is used to synchronize the vertical sweep to a 60-Hz V-sync pulse.

Color burst is added to the video signal to provide a reference for the decoding of color information. From line to line, the burst maintains the correct phase and frequency to a high degree of precision. Active video consists of a luminance signal, which controls picture brightness, and a chrominance signal, which sets the hue and saturation. Color information from the camera is first modulated by the subcarrier signal at approximately 3.58 MHz. This modulated signal is added to the luminance signal to form the composite NTSC signal. At any instant, hue is deter-

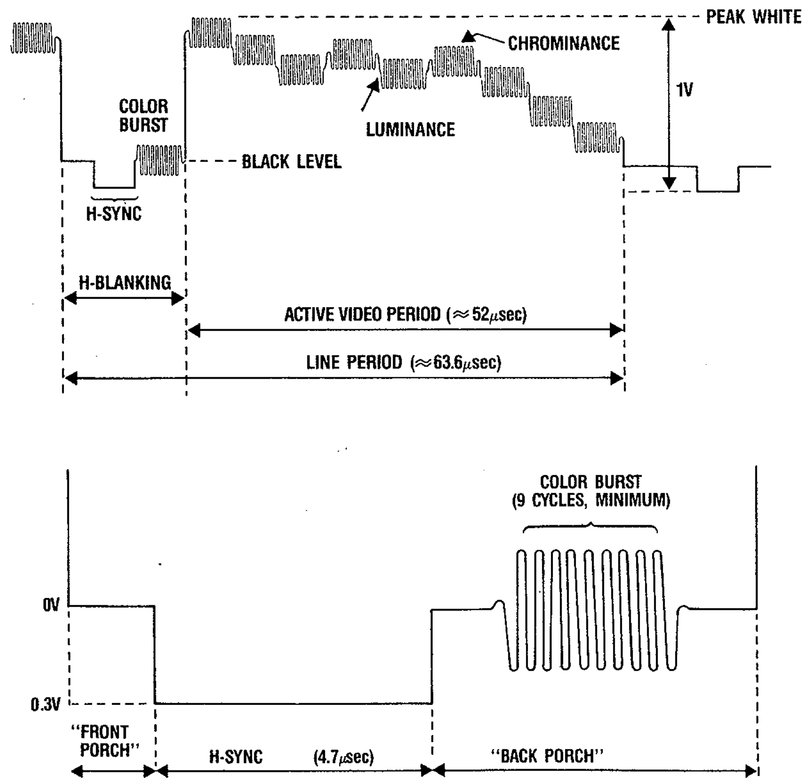


Figure 1: (a) Composite color standards of video equipment are made up of video signals like this one. During the active video period, the voltage level is used to provide input to a CRT's electron guns, causing them to produce variations in color and brightness. (b) To enable the CRT beam to retrace without drawing to the screen, a horizontal blanking period is defined. In this period, the color reference information, or color burst, is transmitted.

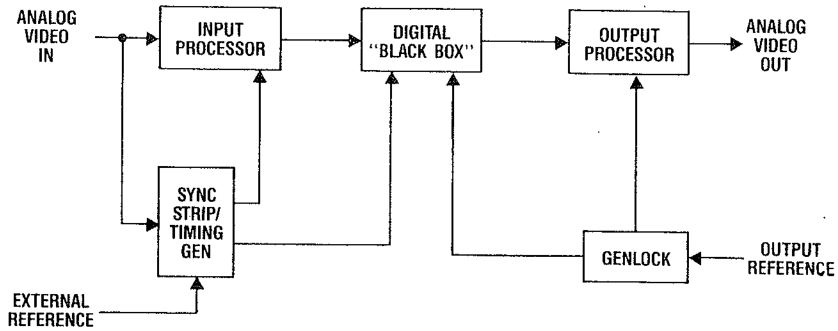


Figure 2: In digital video systems, an input processor operates on analog video to prepare it for A/D conversion. Sync stripper/timing generators remove the sync and color-burst information from the signal for further digital processing. After processing, timing information is reinserted either from the incoming video source or external reference.

mined by the phase difference between the chrominance signal and the color burst. Thus, a range of colors can be encoded in 360° of phase.

Since a hue shift of a few degrees is noticeable, a high degree of precision must be maintained in the color burst. Static errors of a few nanoseconds for the color burst or its decoding circuitry will cause noticeable misregistration of hues.

Standard Digital Video

Figure 2 shows a block diagram of a standard digital video system. The input processor operates on the analog input video signal and prepares it for A/D conversion. Gain adjustment, antialiasing, and phase equalization are usually provided. If it is necessary to operate on video components instead of the composite signal, a decoding function is implemented.

The sync stripper/timing generator takes as its input either the input video waveform or an external reference, and locks two separate voltage-controlled oscillators (VCOs) to the horizontal sync and color burst. Outputs of this module are used by both analog and digital processing sections. The A/D converter needs a clock pulse to tell it when to sample the incoming video. The sampling pulse may occur at a harmonic of the subcarrier frequency or the H-pulse.

Processing of video takes place in a digital processor. Data is read into the processor from the analog section and is clocked by the timing generator. All processing takes place in the digital domain without the degradations associated with analog processing.

The output processor returns the video data to the analog domain for interfacing to analog video systems. This digital "black box" will contain a D/A conversion, reconstruction filtering, and timing insertion (e.g., blanking and burst).

Genlock provides basically the same function as does timing generation. Input and output video are clocked independently, and the genlock module synchronizes the timing of the output video to an external output reference. If the same clocking is used for both inputs and outputs, the genlock module may be omitted.

For digital processing, an analog signal is digitized by an ADC. However, prior to conversion, the signal must be low-pass filtered to avoid the phenomena of aliasing.

Shannon's sampling theorem states that a signal must be sampled at a rate that is at least twice as high as its highest frequency component. This minimum sampling frequency is commonly called the Nyquist rate. If signals are sampled at lower than the Nyquist rate, alias components will result, and it will be impossible to recover the original signal exactly. For a baseband input signal with high-frequency noise, aliasing will result unless some form of low-pass filtering is performed prior to A/D conversion.

Choosing Filters

Choice of antialiasing filters depends on the rate of sampling of the input signal. If the signal can be oversampled (at a rate much higher than the Nyquist rate), a low-pass filter with a slow roll-off can be used. But, for video signals with a nominal bandwidth of 4.2 MHz, this method is in-

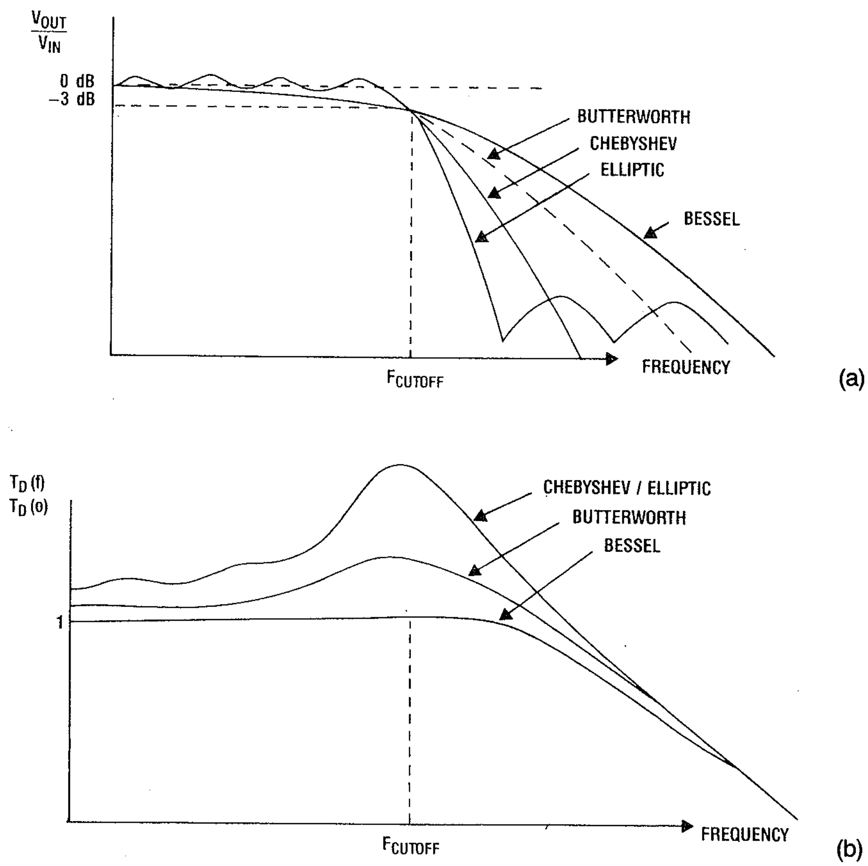


Figure 3: Prior to A/D conversion, the video signal must be filtered to remove aliasing components. The input signal's rate of sampling dictates filter choice. (a) Popular filter designs include Butterworth, Bessel, Chebyshev, and elliptic approximations. (b) Magnitude and phase responses will have varying degrees of phase nonlinearity and high variance in group-delay response.

efficient because of the high digital sample rate. Alternatively, a more complicated filter with a faster roll-off can be used.

It is impossible to generate an infinitely fast filter roll-off with a finite number of components. Complex designs approach, but never reach, the ideal. There are several popular filter designs that are generated from polynomial approximations to the ideal low-pass filter characteristic. These include Butterworth, Bessel, Chebyshev, and elliptic approximations (**Figure 3a**).

Butterworth filters exhibit the flattest possible response at dc, and moderately fast roll-off. Their mathematical function was derived on the assumption that the response near zero frequency is important. Advantages of Butterworth filters are mathematical simplicity and the fact that they can be easily synthesized with off-the-shelf components.

Bessel's approximation ensures maximally flat phase response. This is important in pulse-transmission systems where the maintenance of pulse shape without overshoot or undershoot is critical. The response in the passband is not as flat as for the Butterworth approximation, and the roll-off in the stop band is fairly slow. The filter starts rolling off from its dc value at well below the cutoff frequency.

Chebyshev filters have a predetermined passband ripple and, compared to the Butterworth design, a more rapid attenuation above the cutoff frequency. There are an infinite number of Chebyshev designs for a specified filter order; a design with a 1-dB ripple in the passband will have a faster roll-off than one with a maximum ripple of 0.1 dB.

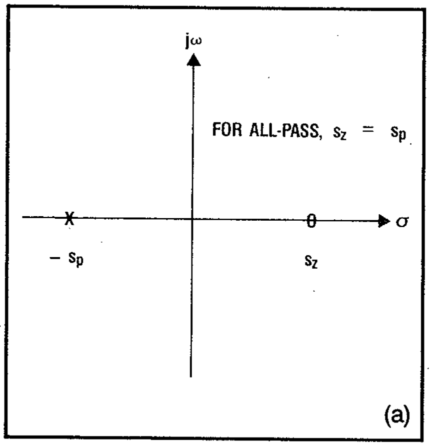
Elliptic filters (alternatively called Cauer-Chebyshev or brick-wall filters) are a subset of Chebyshev designs. They exhibit a response with ripple in both passband and stopband, and a very sharp roll-off rate.

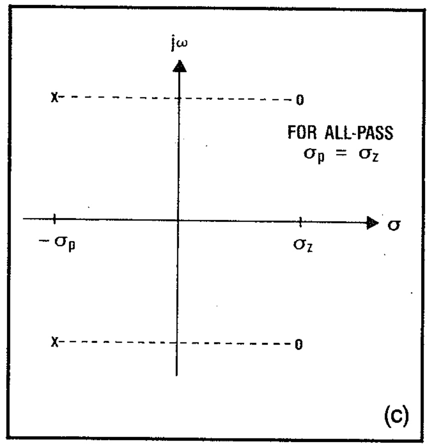
Group Delay

Group delay is a measure of how much a group of frequencies is delayed when passing through a system. Mathematically, it is the negative derivative of the phase response, with respect to frequency, and can be described as:

$$D(\omega) = \frac{-dP(\omega)}{d\omega} \qquad \text{Equation 1}$$

where $P(\omega)$ is the phase response in relation to frequency. For a modulated waveform, the group-delay function





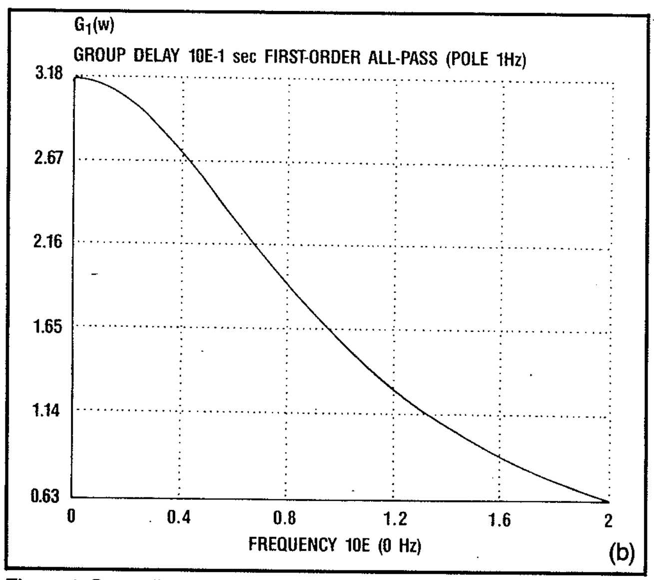
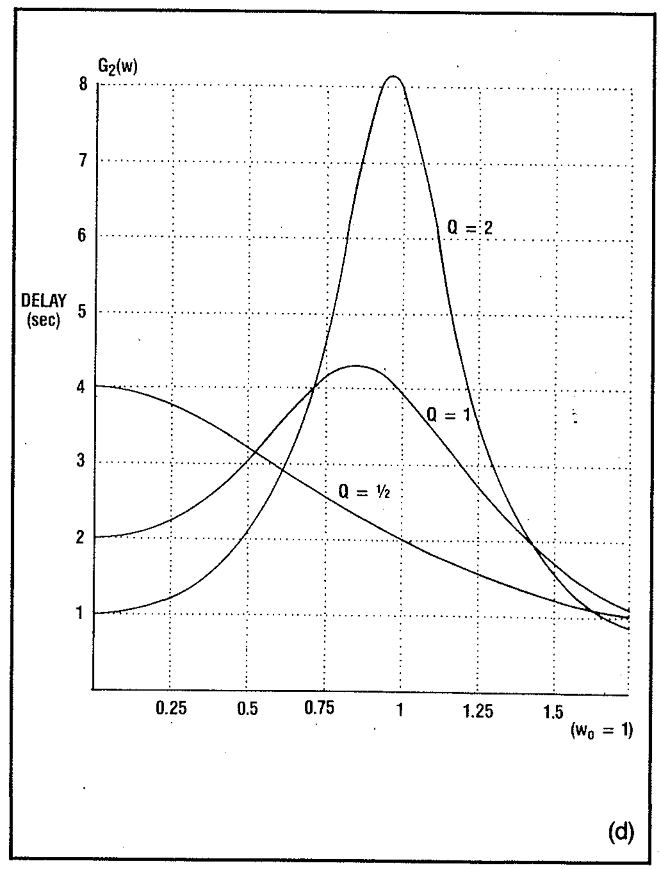


Figure 4: Cascading low-pass filters with all-pass functions compensates for phase nonlinearities. (a) First-order all-pass pole-zero plots like this one correspond to the transfer function shown in Equation 2. For sp = sz, the magnitude is constant for all frequencies. (b) By changing the pole and zero locations, the group delay over frequency can be varied. Here, sp = 1 radian/sec is plotted. (c) For some choices of filter, the filter must be tuned to meet the group-delay specification. (d) Here, a second-order all-pass function allows a better fit to the ideal group-delay response for a cascaded filter system.



shows how the transient response of the envelope behaves. Thus, group delay is also called envelope delay.

All filters add a phase shift when a sinusoid passes through them. A linear phase shift with frequency implies constant group delay. Near-constant group delay is a requirement for pulse and information transmission systems.

Consider a square wave input to the antialias filter. The square wave contains frequency components at multiples of the fundamental frequency. If each sinewave component is not delayed by the same amount when passing through the filter, the square wave will be distorted. For any processing system, there is an upper limit to the amount of phase distortion that can be tolerated.

Phase distortion in an NTSC signal will cause luminance overshoot and hue shifts on vertical color transitions. Subjective tests have been performed to assess tolerances of the human visual system. To insure good image quality, great care must be taken to account for phase distortion and group delay.

In practice, magnitude and phase responses cannot be independently specified. The implications are that a very sharp cutoff filter (an elliptic, for instance) will have very nonlinear phase and high variance in group-delay response. In **Figure 3b**, group-delay responses for the previously discussed filters are compared.

Video filter design involves trade-offs in the required roll-off rate, attenuation floor, and amount of phase distortion. For an NTSC signal with 4.2-MHz bandwidth, a 10.7-MHz sampling rate with 8 bits/sample is often used. Utilizing these figures, and assuming a maximum allowable group-delay deviation, the designer may derive a specification for the antialiasing low-pass filter.

An elliptic approximation is often chosen for video filtering because, above the cutoff frequency, its behavior approaches the brick-wall response. However, the phase response for this type of filter leaves much to be desired. As an alternative, the designer could use a Butterworth filter to increase the order of the filter until it meets the magnitude response specifications. For this type of filter, the phase distortion below the cutoff frequency increases as the filter order increases.

All-Pass Transfer

Another solution is to cascade the low-pass filter with a system that will compensate for phase nonlinearities. To accomplish this, a set of *all-pass* transfer functions can be used. These exhibit constant magnitude and a phase response that varies with frequency.

All-pass functions have pole-zero plots in the complex plane with poles in the left-half plane and mirror zeros in the right-half plane. Consider the pole-zero diagram of **Figure 4a.** This corresponds to a transfer function of:

$$H(s) = K\frac{s - s_z}{s + s_p} \qquad \text{Equation 2}$$

From this function, a magnitude function $M(\omega)$ can be derived:

$$M(\omega) = K\sqrt{\frac{\omega^2 + (-s_z)^2}{\omega^2 + (s_p)^2}} \qquad \text{Equation 3}$$

For $s_p = s_z$, the magnitude is constant for all frequencies. If the value of K is 1, cascading this system with an elliptic filter will not affect the magnitude response. Phase response $P(\omega)$ of the all-pass network is given by:

$$P(\omega) = -2\tan^{-1}\left(\frac{\omega}{s_p}\right); s_p = s_z \qquad \text{Equation 4}$$

The group delay is the negative derivative of the phase response with respect to frequency.

$$G_1(\omega) = \frac{\frac{2}{s_p}}{1 + \left(\frac{\omega}{s_p}\right)^2} \qquad \text{Equation 5}$$

By changing the pole and zero locations, the group delay over frequency can be varied. The group-delay function for sp = 1 radian/sec is plotted in **Figure 4b.** By cascading a properly scaled version of this system with the elliptic filter, the total delay through the filter will increase, and the group-delay deviation from the value at zero frequency will be reduced.

For some antialiasing filters, the first-order all-pass function may not allow the filter to be tuned to meet the group-delay spec. The second-order all-pass function (**Figure 4c**) has two poles and two zeros. Its transfer function is given by:

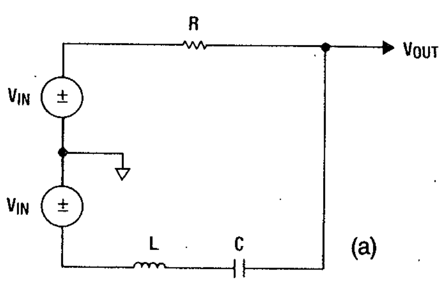
$$H_2(\omega) = \frac{s^2 - \left(\frac{\omega_o}{Q}\right) s + \omega_o^2}{s^2 + \left(\frac{\omega_o}{Q}\right) s + \omega_o^2} \qquad \text{Equation 6}$$

The group-delay response, $G_2(\omega)$, is plotted in **Figure 4d** for various values of Q.

The two degrees of freedom in this design (ω_o and Q) allow a better fit to the ideal group-delay response for the cascaded system. In many video designs, a second-order all-pass delay equalizer is sufficient.

In **Figure 5a,** the transfer function V_o/V_i takes the form of $H_2(\omega)$. In terms of circuit parameters, $\omega_o^2 = 1/LC$ and $\omega_o/Q = R/L$. By varying R, L, and C, the natural frequency and filter Q may be independently specified.

Figure 5b is a transistor circuit implementation of Figure 5a. Transistor Q_1



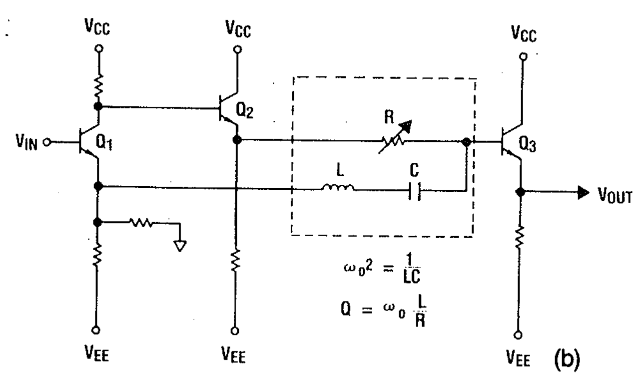


Figure 5: (a) All-pass filter implementation. Here, the transfer function V_o/V_i takes the form $H_2(\omega)$. (b) Transistor implementation of Figure 5a, with Q_1 acting as phase splitter, Q_2 acting to provide a low-impedance source to the R-L-C filter network, and Q_3 acting as output buffer.

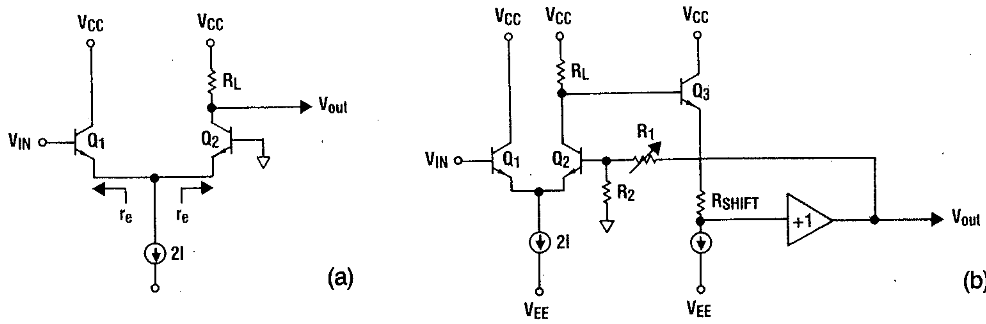


Figure 6: (a) Differential amps provide gain down to dc. The input voltage signal range in which the I/O relationship stays linear is only a few mV. To solve this, feedback amps can be used. (b) Negative feedback ensures that the amp tries to drive bases of Q_1 and Q_2 to the same potential.

is a phase splitter that provides the input signal and its negative. Q_2 buffers the high impedance seen at the collector of Q_1 to provide a low-impedance source to the all-pass R-L-C network. Q_3 provides buffering to insure that the output does not load the all-pass circuit.

Gain Adjustment

In general, any video system needs to have a gain adjustment in its input processing section to calibrate the signal amplitude in order to use the full dynamic range of the A/D converter. One advantage in placing the gain stage at the front end is that it will increase the S/N ratio throughout the string of analog processing. Specifications for a video amplifier may include gain adjustment from 1 to 2, low-frequency cutoff of less than a few hertz, and high-frequency cutoff greater than 10 MHz or more.

An important gain topology that provides gain down to dc is the differential amplifier (**Figure 6a**). Transistor Q_2 is connected as a common base amplifier, while transistor Q_1 is connected as an emitter follower. The two transistors are biased into the linear operating region by the current source. Assume that the transistors are perfectly matched and are at the same temperature. In equilibrium, with $V_{in} = 0$, the bias current splits equally between Q_1 and Q_2. V_{out} will be a dc value, $V_{cc}—IR_L$.

The output resistance at the emitter node of transister Q_1 is given by:

$$r_e \cong \frac{1}{g_m} \qquad \text{Equation 7}$$

where g_m is the transconductance of a transistor, and is related to the bias current, I_c.

$$g_m = I_c/26\text{ mV at room temperature} \qquad \text{Equation 8}$$

Since Q_2 is biased at the same current level as Q_1, its output resistance is also the same. To determine the ac gain, consider a small wiggle of input voltage (V_{in}). Since Q_1 is an emitter follower, the wiggle will be reproduced at the emitter of Q_1. This voltage will cause an incremental current to flow into the emitter of Q2. The value of this current is given by:

$$i_{inc} = V_{in}/2r_e \qquad \text{Equation 9}$$

where V_{in} is the small signal wiggle and r_e is the resistance looking into the emitter of Q_2. Q_2 has a current gain from emitter to collector of approximately 1. This incremental current will flow through the load resistor R_L and generate an incremental voltage

$$V_{out} = \frac{R_L}{2r_e} \times V_{in} \qquad \text{Equation 10}$$

Or, equivalently,

$$\frac{V_{out}}{V_{in}} = g_m \frac{R_L}{2}$$

Unfortunately, the differential amplifier of **Figure 6a** has several practical limitations. The input voltage signal range in which the input/output relationship remains approximately linear is only a few millivolts. Furthermore, gain and dc bias levels are very dependent upon ambient temperature and transistor device parameters.

Feedback can be used to solve these problems. Operational amplifiers rely on the differential gain stage as their inputs. Assuming that the open-loop gain of the amplifier is large at all frequencies of interest, the application of feedback ensures that the closed-loop performance will essentially be independent of amplifier parameters. Gain can be set by varying a single resistor.

A basic IV gain-stage feedback amplifier is shown in **Figure 6b.** Q_1 and Q_2 are the differential gain stage. Q_3 and resistor R_{shift} provide a dc-level shift. The +1 gain stage buffers the high-impedance level shifter from the outside world. The application of negative feedback ensures that the amplifier tries to drive the bases of Q_1 and Q_2 to the same potential. By providing a resistive voltage divider between the output and the base of Q_2, gain can be adjusted by varying the ratio of R_1 to R_2.

Video A/D Conversion

When digitizing video, too much is always better than not enough. Thus, video signals must be sampled considerably faster than the Nyquist rate, and have a larger number of bits than the S/N ratio of video would indicate.

Much work has been directed at establishing the encoding parameters for digital television. Generally, 8 bits are needed for digital composite video, providing an S/N ratio of approximately 60 dB. In many applications, however, 7 or fewer bits will suffice.

Sampling of a video signal for digital processing results in a sampling grid in three dimensions: horizontal, vertical, and temporal. For efficient processing, the sampling rate should be as low as possible without introducing aliasing. For sampling in the horizontal direction, the NTSC nominal bandwidth of 4.2 MHz is the constraining factor.

In practice, to determine the minimum sampling frequency, the Nyquist limit must be multiplied by a factor of at least 1.1 to 1.2 to allow for a finite antialiasing filter transition width. For NTSC, this translates to a minimum sampling frequency of approximately 10 MHz.

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