Chapter

17.1

Television Reception Principles

K. Blair Benson

L. H. Hoke, Jr., L. E. Donovan, J. D. Knox, D. E. Manners, W. G. Miller, R. J. Peffer, J. G. Zahnen

17.1.1 Introduction

Television receivers provide black-and-white or color reproduction of pictures and the accompanying monaural or stereophonic sound from signals broadcast through the air or via cable distribution systems. The broadcast channels in the U.S. are 6 MHz wide for transmission on conventional 525-line standards.

17.1.2 Basic Operating Principles

The minimum signal level at which television receivers provide usable pictures and sound, called the *sensitivity level*, generally is on the order of 10 to 20 μV. The maximum level encountered in locations near transmitters may be as high as several hundred millivolts. The FCC has set up two standard signal level classifications, Grades A and B, for the purpose of licensing television stations and allocating coverage areas. Grade A is to be used in urban areas relatively near the transmitting tower, and Grade B use ranges from suburban to rural and fringe areas a number of miles from the transmitting antenna. The FCC values are expressed in microvolts per meter (μV/m), where meter is the signal wavelength [1].

The standard transmitter field-strength values for the outer edges of these services for Channels 2 through 69 are listed in Table 17.1.1. Included for reference in the table are the signal levels for what may be considered "city grade" in order to give an indication of the wide range in signal level that a receiver may be required to handle. The actual antenna terminal voltage into a matched receiver load, listed in the second column, is calculated from the following equation:

$$e = E\frac{96.68}{\sqrt{f_1 f_2}} \qquad (17.1.1)$$

Table 17.1.1 Television Service Operating Parameters

Band and Channels	Frequency (MHz)	City grade		Grade A		Grade B	
		μV/m	μV	μV/m	μV	μV/m	μV
VHF 2–6	54–88 MHz	5,010	7030	2510	3520	224	314
VHF 7–13	174–216 MHz	7,080	3550	3550	1770	631	315
UHF 14–69	470–806 MHz	10,000	1570	5010	787	1580	248
UHF 70–83[1]	806–890 MHz	10,000	1570	5010	571	1580	180

1. Receiver coverage of Channels 70 to 83 has been on a voluntary basis since July 1982. This frequency band was reallocated by the FCC to land mobile use in 1975 with the provision that existing transmitters could continue indefinitely.

Where
e = terminal voltage, μV, 300 Ω
E = field, μV/m
f_1 and f_2 = band-edge frequencies, MHz

Many sizes and form factors of receivers are manufactured. Portable personal types include pocket-sized or hand-held models with picture sizes of 2 to 4 in (5 to 10 cm) diagonal for monochrome and 5 to 6 in (13 to 15 cm) for color powered by either batteries or ac. Conventional cathode ray tubes (CRTS) for picture displays in portable sets have essentially been supplanted by flat CRTs and liquid crystal displays.

Larger screen sizes are available in monochrome where low cost and light weight are prime requirements. However, except where extreme portability is important, the vast majority of television program viewing is in color. The 19-in (48-cm) and 27-in (69-cm) sizes now dominate the market, although the smaller 13-in (33-cm) size is popular as a second or semiportable set.

The television receiver functions can be broken down into several interconnected blocks. With the rapidly increasing use of large-scale integrated circuits, the isolation of functions has become more evident in the design and service of receivers, while at the same time the actual parts count has dropped dramatically. The typical functional configuration of a receiver using a tri-gun picture tube, shown in Figure 17.1.1, will serve as a guide for the following description of receiver design and operation. The discussions of each major block, in turn, are accompanied with more detailed subblock diagrams.

17.1.2a Tuner Principles

The purpose of the tuner, and the following *intermediate amplifier* (IF), is to select the desired radio frequency (RF) signals in a 6 MHz channel, to the exclusion of all other signals, available from the antenna or cable system and to amplify the signals to a level adequate for demodulation. Channel selection is accomplished with either mechanically switched and manually tuned circuits, or in varactor tuners with electrically switched and controlled circuit components. A mechanical tuner consists of two units, one for the VHF band from 54 to 88 and 174 to 216 MHz, and the other for the UHF band from 470 to 806 MHz. Two separate antenna connections are provided for the VHF and UHF sections of the tuner.

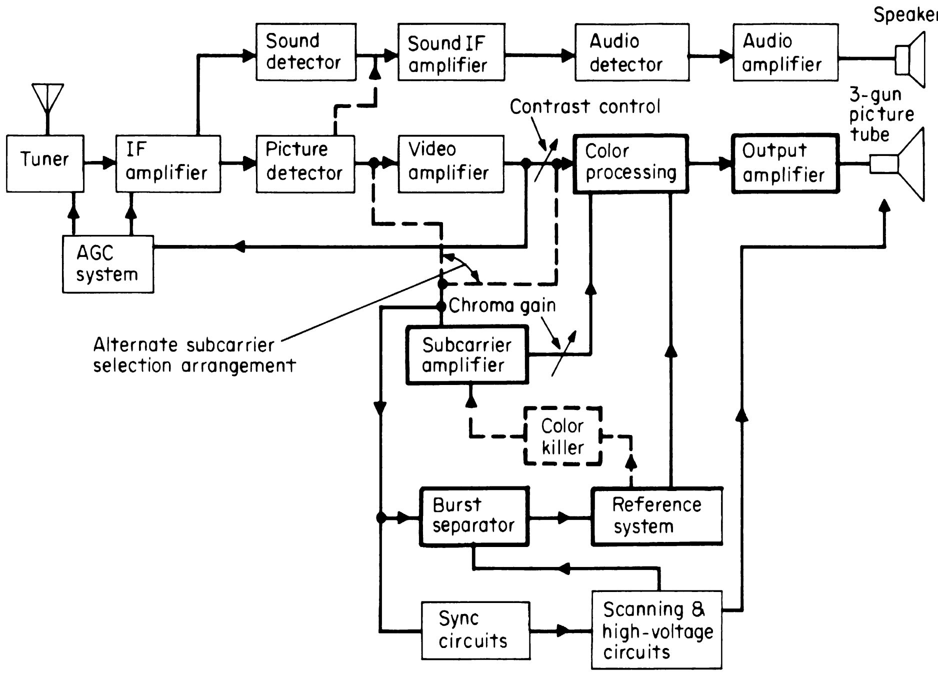


Figure 17.1.1 Fundamental block diagram of a color receiver with a tri-gun picture tube display.

Varactor tuners, on the other hand, have no moving parts or mechanisms and consequently are less than a third the volume of their mechanical equivalent. Part of this smaller size is the result of combining the VHF and UHF circuits on a single printed circuit board in the same shielded box with a common antenna connection, thus eliminating the need for an outrigger coupling unit.

Selectivity

The tuner bandpass generally is 10 MHz in order to ensure that the picture and sound signals of the full 6 MHz television channel are amplified with no significant imbalance in levels or phase distortion by the skirts of the bandpass filters. This bandpass characteristic usually is provided by three tuned circuits:

- A single-tuned preselector between the antenna input and the RF amplifier stage
- A double-tuned interstage network between the RF and mixer stages
- A single-tuned coupling circuit at the mixer output.

The first two circuits are frequency-selective to the desired channel by varying either or both the inductance and capacitance. The mixer output is tuned to approximately 44 MHz, the center frequency of the IF channel.

Table 17.1.2 Potential VHF Interference Problems

Desired Channel	Interfering Signals	Mechanism
5	Channel. 11 picture	2 × ch. 5 osc. – ch. 11 pix = IF
6	Channel 13 picture	2 × ch. 6 osc. – ch. 13 pix = IF
7 and 8	Channel 5, FM (98–108 MHz)	Ch. 5 pix + FM = ch. 7 and 8
2–6	Channel 5, FM (97–99 MHz)	2 × (FM – ch. 5) = IF
7–13	FM (88–108 MHz)	2 × FM = ch. 7–13
6	FM (89–92 MHz)	Ch. 6 pix + FM – ch. 6 osc. = IF
2	6 m amateur (52–54 MHz)	2 × ch. 2 pix – 6 m = ch. 2
2	CB (27 MHz)	2 × CB = ch. 2
5 and 6	CB (27 MHz)	3 × CB = ch. 5 and 6

The purpose of the RF selectivity function is to reduce all signals that are outside of the selected television channel. For example, the input section of VHF tuners usually contains a high-pass filter and trap section to reject signals lower than Channel 2 (54 MHz), such as standard broadcast, amateur, and citizen's band (CB) emissions. In addition, a trap is provided to reduce FM broadcast signal in the 88 to 108 MHz band. A list of the major interference problems is tabulated in Table 17.1.2 for VHF channels. In Table 17.1.3 for UHF channels, the formula for calculation of the interfering channels is given in the second column, and the calculation for a receiver tuned to Channel 30 is given in the third column.

VHF Tuner

A block diagram of a typical mechanical tuner is shown in Figure 17.1.2. The antenna is coupled to a tunable RF stage through a bandpass filter to reduce spurious interference signals in the IF band, or from FM broadcast stations and CB transmitters. Another bandpass filter is provided in the UHF section for the same purpose. The typical responses of these filters are shown in Figures 17.1.3.*a* and *b*.

The RF stage provides a gain of 15 to 20 dB (approximately a 10:1 voltage gain) with a bandpass selectivity of about 10 MHz between the –3 dB points on the response curve. The response between these points is relatively flat with a dip of only a decibel or so at the midpoint. Therefore, the response over the narrower 6 MHz television channel bandwidth is essentially flat.

VHF tuners have a rotary shaft that switches a different set of three or four coils or coil taps into the circuit at each VHF channel position (2 to 13). The circuits with these switched coils are the following:

- RF input preselection
- RF input coupling (single-tuned for monochrome, double-tuned for color)
- RF-to-mixer interstage

In the first switch position (Channel 1), the RF stage is disabled and the mixer stage becomes an IF amplifier stage, centered on 44 MHz for the UHF tuner.

Table 17.1.3 Potential UHF Interference Problems

Interference Type	Interfering Channels	Channel 30 Example
IF beat	$N \pm 7, \pm 8$	22, 23, 37, 38
Intermodulation	$N \pm 2, \pm 3, \pm 4, \pm 5$	25–28, 32–35
Adjacent channel	$N + 1, - 1$	29, 31
Local oscillator	$N \pm 7 \times$	23, 37
Sound image	$N + 1/6\ (2 \times 41.25)$	44
Picture image	$N + 1/6\ (2 \times 45.75)$	45

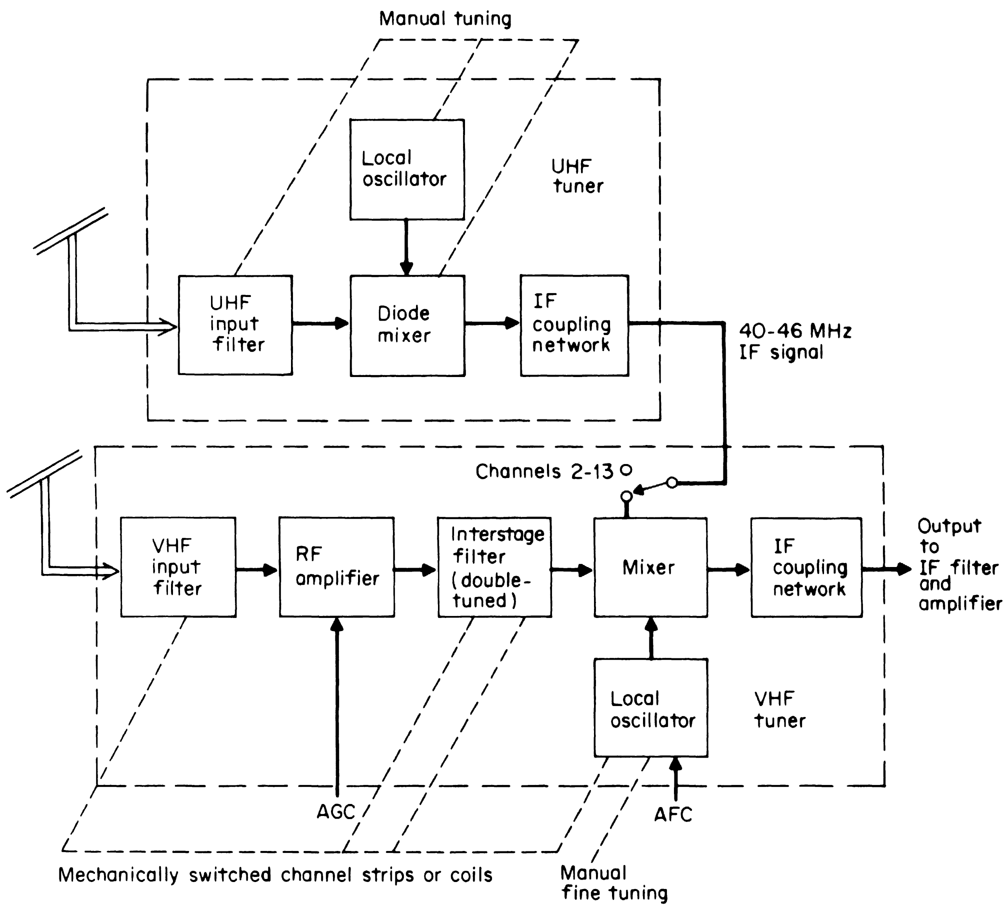


Figure 17.1.2 Typical mechanical-tuner configuration.

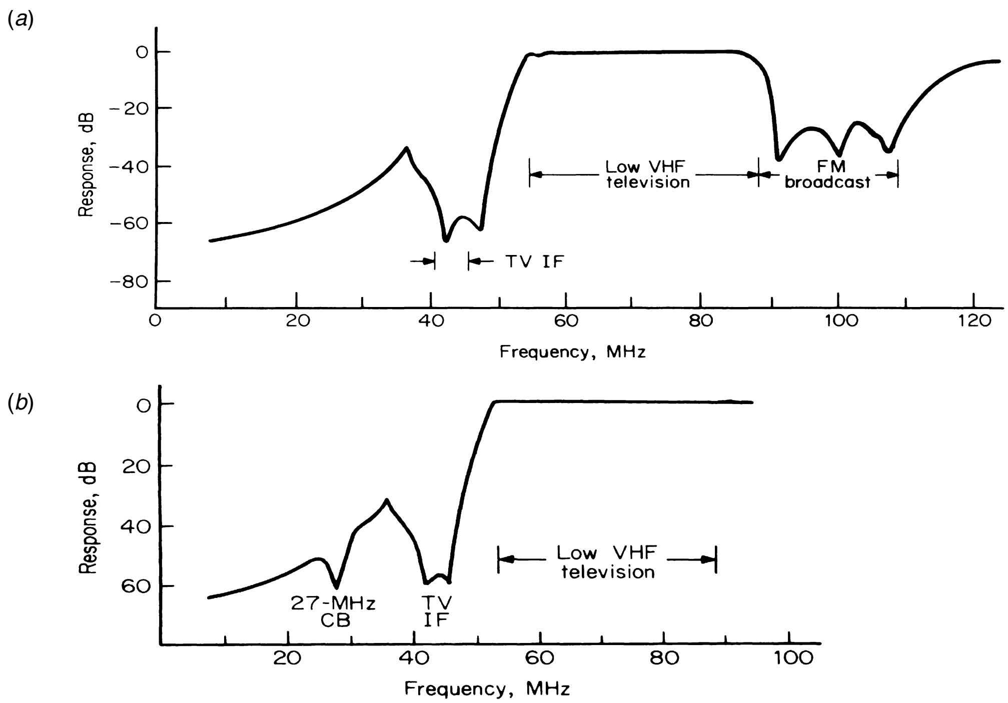


Figure 17.1.3 Filter response characteristics: (*a*) response of a tuner input FM bandstop filter, (*b*) response of a tuner input with CB and IF traps.

The mixer stage combines the RF signal with the output of a tunable local oscillator to produce an IF of 43.75 MHz for the picture carrier signal and 42.25 MHz for the sound carrier signal. The local oscillator signal thus is always 45.75 MHz above that of the selected incoming picture signal. For example, the frequencies for Channel 2 are listed in Table 17.1.4.

These frequencies were chosen to minimize interference from one television receiver into another by always having the local-oscillator signal above the VHF channels. Note that the oscillator frequencies for the low VHF channels (2 to 6) are between Channels 6 (low VHF) and 7 (high VHF), and the oscillator frequencies for the high VHF fall above these channels.

The picture and sound signals of the full 6 MHz television channel are amplified with no significant imbalance in levels or phase distortion by the skirts of the bandpass filters. This bandpass characteristic usually is provided by three tuned circuits:

- A single-tuned preselector between the antenna input and the RF amplifier stage
- Double-tuned interstage network between the RF and mixer stages
- Single-tuned coupling circuit at the mixer output

The first two are frequency-selective to the desired channel by varying either or both the inductance and capacitance. The mixer output is tuned to approximately 44 MHz, the center frequency of the IF channel.

Table 17.1.4 Local Oscillator and IF Frequencies

Parameter	Channel 2, MHz	Channel 6, MHz	Channel 7, MHz
Channel width	54–60	82–88	174–180
Local oscillator	101.00	129.00	221.00
Less picture signal	55.25	83.25	175.25
IF picture signal	45.75	45.75	45.75

UHF Tuner

The UHF tuner contains a tunable input circuit to select the desired channel, followed by a diode mixer. As in a VHF tuner, the local oscillator is operated at 45.75 MHz above the selected input channel signal. The output of the UHF mixer is fed to the mixer of the accompanying VHF tuner, which functions as an IF amplifier. Selection between UHF and VHF is made by applying power to the appropriate tuner RF stage.

Mechanical UHF tuners have a shaft that when rotated moves one set of plates of variable air-dielectric capacitors in three resonant circuits. The first two are a double-tuned preselector in the amplifier-mixer coupling circuit, and the third is the tank circuit of the local oscillator. In order to meet the discrete selection requirement of the FCC, a mechanical detent on the rotation of the shaft and a channel-selector indicator are provided, as illustrated in Figure 17.1.4.

The inductor for each tuned circuit is a rigid metal strip, grounded at one end to the tuner shield and connected at the other end to the fixed plate of a three-section variable capacitor with the rotary plates grounded. The three tuned circuits are separated by two internal shields that divide the tuner box into three compartments.

Tuner-IF Link Coupling

With mechanically switched tuners, it has usually been necessary to place the tuner behind the viewer control panel and connect it to the IF section, located on the chassis, with a foot or so length of shielded 50 or 75 Ω coaxial cable. Because the output of the tuner and the input of the IF amplifier are high-impedance-tuned circuits, for maximum signal transfer, it is necessary to couple these to the cable with impedance-matching networks.

Two common resonant circuit arrangements are shown in Figure 17.1.5. The low-side capacitive system has a low-pass characteristic that attenuates the local oscillator and mixer harmonic currents ahead of the IF amplifier. This can be an advantage in controlling local-oscillator radiation and in reducing the generation of spurious signals in the IF section. On the other hand, the low-side inductance gives a better termination to the link cable and therefore reduces interstage cable loss. The necessary bandpass characteristics can be obtained either by undercoupled stagger tuning or by overcoupled synchronous tuning, as illustrated in Figure 17.1.5.

17.1.2b Advanced Tuner Systems

Mechanically tuned television receivers were the mainstay of consumer sets since the beginning of TV broadcasting. It was not until the 1970s and 1980s that electronically tuned systems became practical. More recent technological trends include microprocessor-based control of the

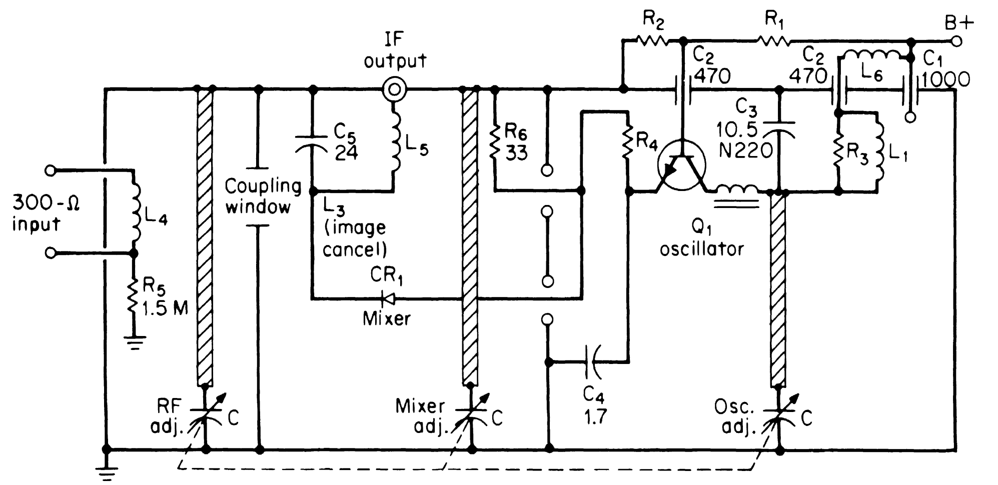


Figure 17.1.4 UHF rotary turner. (*Courtesy of Philips Consumer Electronics.*)

tuning functions. Many consumer models have, in fact, completely dispensed with the conventional tuner controls, in favor of using the remote control as the primary user interface. This being the case, considerable integration of functions can be gained, resulting in performance improvements and cost savings.

The preceding sections on mechanically tuned receivers are, however, still important today because they form the foundations for the all-electronic tuning systems that are prevalent today.

Varactor Tuner

The varactor diode forms the basis for electronic tuning, which is accomplished by a change in capacitance with the applied dc voltage to the device. One diode is used in each tuned circuit. Unlike variable air-dielectric capacitors, *varicaps* have a resistive component in addition to their capacitance that lowers the Q and results in a degraded noise figure. Therefore, varactor UHF tuners usually include an RF amplifier stage, making it functionally similar to a VHF tuner. (See Figure 17.1.6.)

The full UHF band can be covered by a single varicap in a tuned circuit because the ratio of highest and lowest frequencies in the UHF bands is less than 2:1 (1.7). However, the ratio of the highest to lowest frequencies in the two VHF bands is over twice (4.07) that of the UHF band. This is beyond the range that typically can be covered by a tuned circuit using varicaps. This problem is solved by the use of band switching between the low and high VHF channels. This is accomplished rather simply by short-circuiting a part of the tuning coil in each resonant tank circuit to reduce its inductance. The short circuit is provided by a diode that has a low resistance in the forward-biased condition and a low capacitance in the reverse-biased condition. A typical RF and oscillator circuit arrangement is shown in Figure 17.1.7. Applying a positive voltage to V, switches the tuner to high VHF by causing the diodes to conduct and lower the inductance of the tuning circuits.

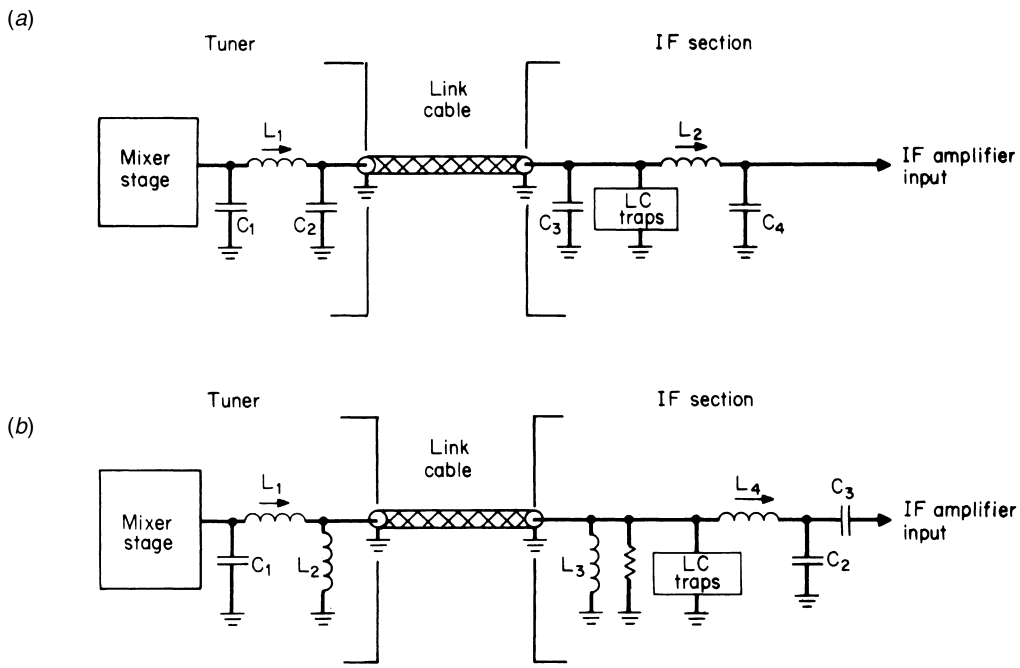


Figure 17.1.5 Tuner-to-IF section link coupling: (*a*) low-side capacitive coupling, (*b*) low-side inductive coupling.

Tuning Systems

The purpose of the tuning system is to set the tuner, VHF or UHF, to the desired channel and to fine-tune the local oscillator for the video carrier from the mixer to be set at the proper IF frequency of 46.75 MHz. In mechanical tuners, this obviously involves an adjustment of the rotary selector switch and the capacitor knob on the switch shaft. In electronically tuned systems, the dc tuning voltage can be supplied from the wiper arm of a potentiometer control connected to a fixed voltage source as shown in Figure 17.1.8*a*.

Alternatively, multiples of this circuit, as shown in Figure 17.1.8*b*, can provide preset fine-tuning for each channel. This arrangement most commonly is found in cable-channel selector boxes supplied with an external cable processor.

In digital systems, such as that shown in Figure 17.1.8*c*, the tuning voltage can be read as a digital word from the memory of a keyboard and display station (or remote control circuit). After conversion from a digital code to an analog voltage, the tuning control voltage is sent to the tuner.

Figure 17.1.8*d* shows a microprocessor system using a phase-locked loop to compare a medium-frequency square-wave signal from the channel selector keyboard, corresponding to a specific channel, with a signal divided down by 256 from the local oscillator. The error signal generated by the difference in these two frequencies is filtered and used to correct the tuning voltage being supplied to the tuner.

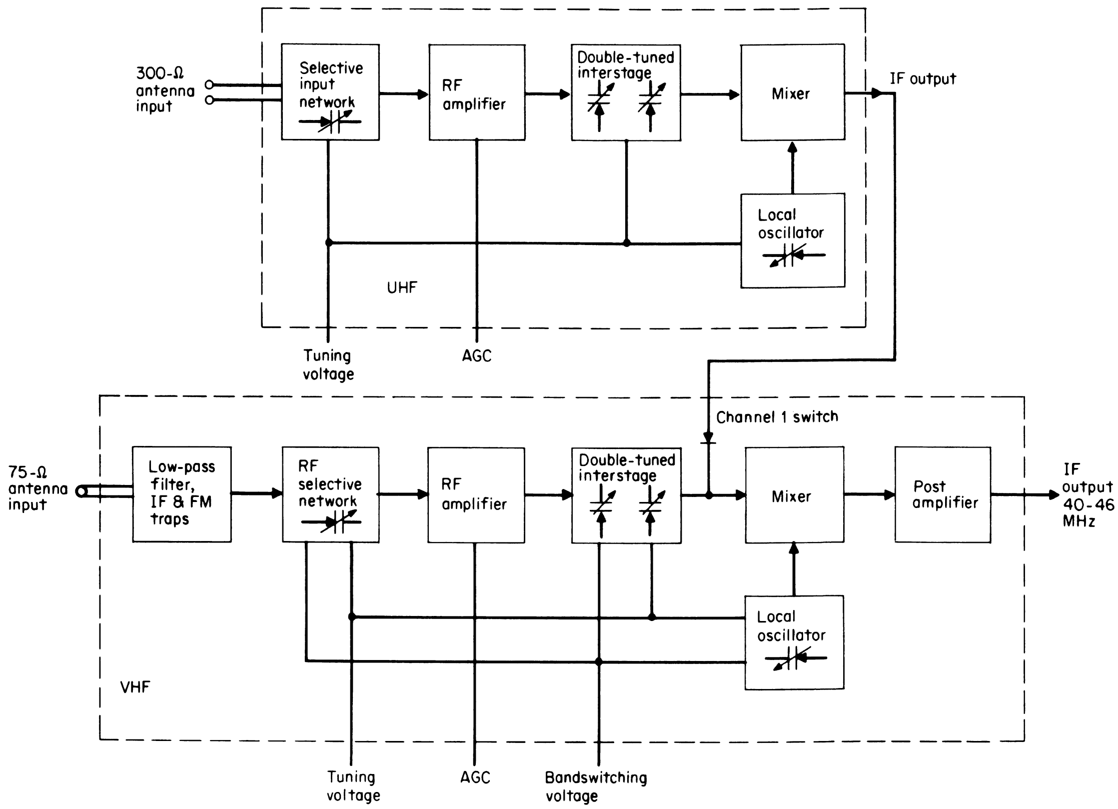


Figure 17.1.6 Typical varactor-tuned block configuration.

17.1.2c Intermediate-Frequency (IF) Amplifier Requirements

The IF picture and sound carrier frequencies standardized for conventional television receivers were chosen with prime consideration of possible degradation of the picture from interfering signals. The picture-carrier frequency is 45.75 MHz and, with the local oscillator above the received RF signal, the sound carrier frequency is 41.25 MHz.

The three factors given greatest emphasis in the choice were the following:

- Interference from other nontelevision services
- Interference from the fundamental and harmonics of local oscillators in other television receivers
- Spurious responses from the image signal in the mixer conversion and from harmonics of the IF signals

Analysis of the relationships indicates the soundness of the choice of 45.75 MHz for the IF picture carrier. The important advantages include the following:

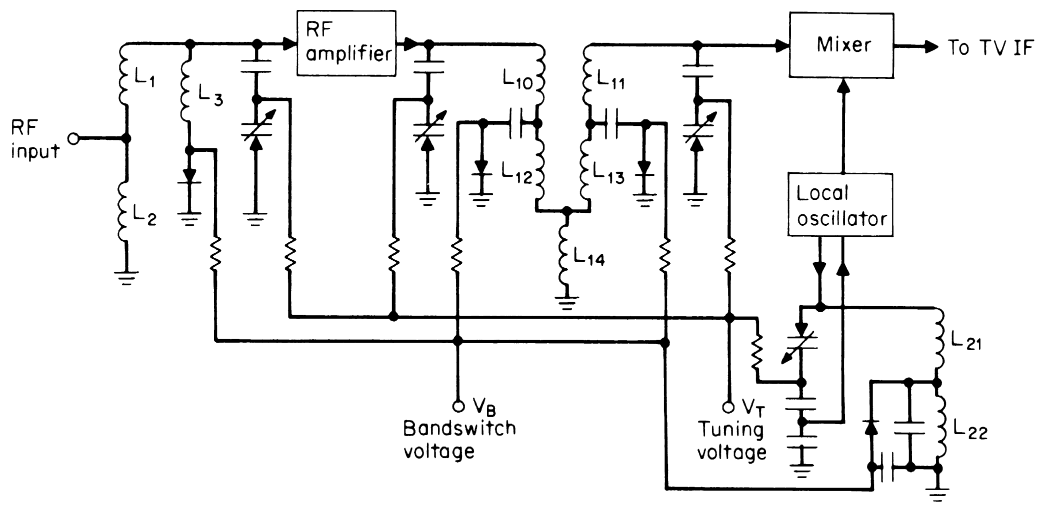


Figure 17.1.7 VHF tuner band-switching and tuning circuits. $+V_B$ = high VHF (active tuning inductors = L_3 in parallel with L_1 and L_2, L_{10}, L_{11}, L_{21}). $-V_B$ = low VHF (active tuning inductors = $L_1 + L_2$, L_{12}, $L_{11} + L_{13}$, L_{14}, $L_{21} + L_{22}$).

- No images from the mixer conversion process fall within the VHF band selected by the tuner except for a negligible interference on the edge of the Channel 7 passband from the image from another receiver tuned to Channel 6.
- All channels are clear of picture harmonics except that the fourth harmonic of the IF picture carrier falls near the Channel 8 picture carrier. This can cause a noticeable beat pattern in another receiver if the offending receiver has not been designed with adequate shielding.
- Local oscillator radiation does not interfere with another receiver on any channel or on any channel image.
- No UHF signal falls on the image frequency of another station.

On the other hand, it should be pointed out that channels for certain public safety communications are allocated in the standard IF band. Because these transmitters radiate high power levels, receivers require thorough shielding of the IF amplifier and IF signal rejection traps in the tuner ahead of the mixer. In locations where severe cases of interference are encountered, the addition of a rejection filter in the antenna input may be necessary.

Gain Characteristics

The output level of the picture and sound carriers from the mixer in the tuner is about 200 µV. The IF section provides amplification to boost this level to about 2 V, which is required for linear operation of the detector or demodulator stage. This is an overall gain of 80 dB. The gain distribution in a typical IF amplifier using discrete gain stages is shown in Figure 17.1.9*a*.

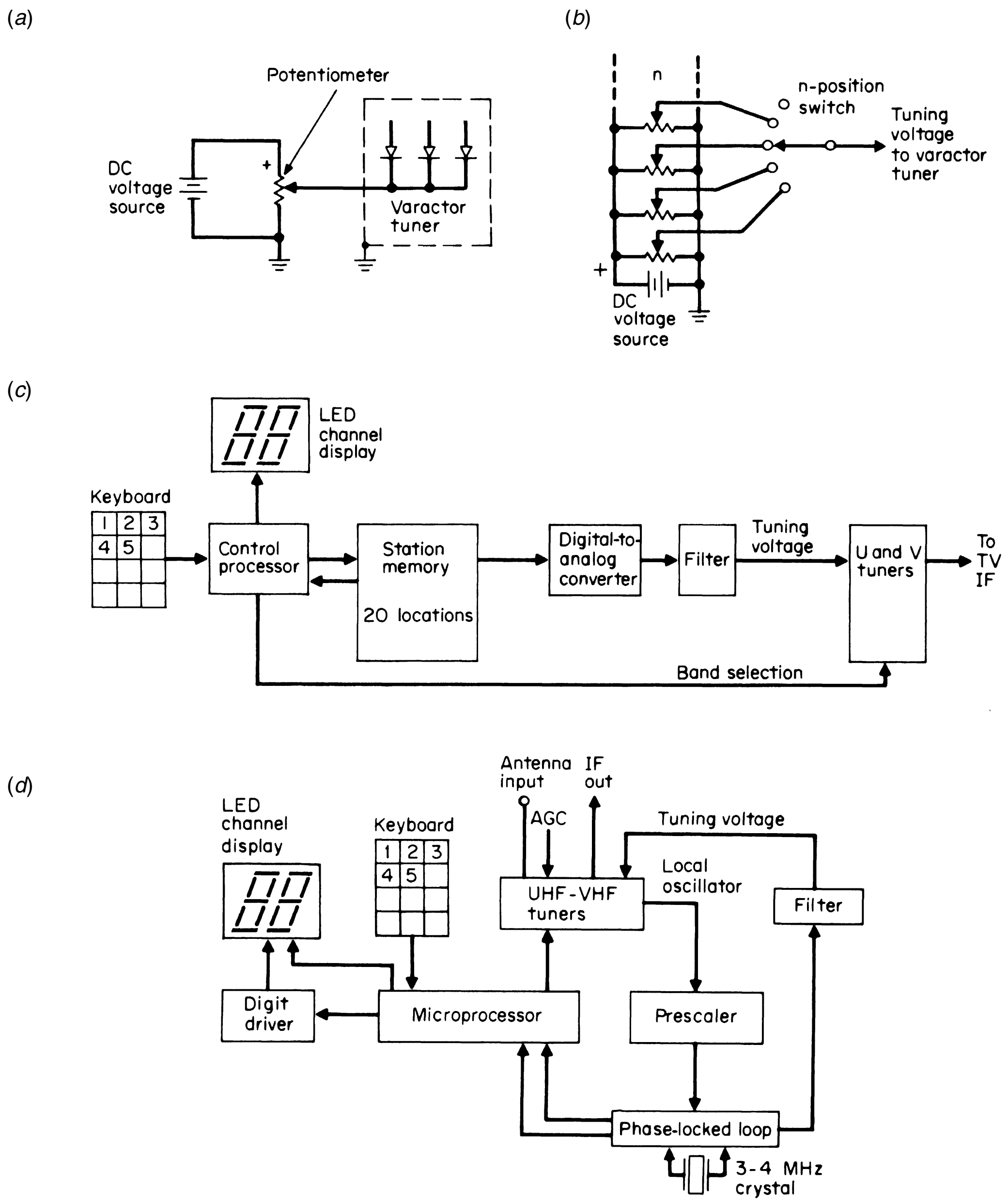


Figure 17.1.8 Varactor-tuned systems: (*a*) simple potentiometer controlled varactor-tuned system, (*b*) multiple potentiometers providing *n*-channel selection, (*c*) simplified memory tuning, (*d*) microprocessor-based PLL tuning system.

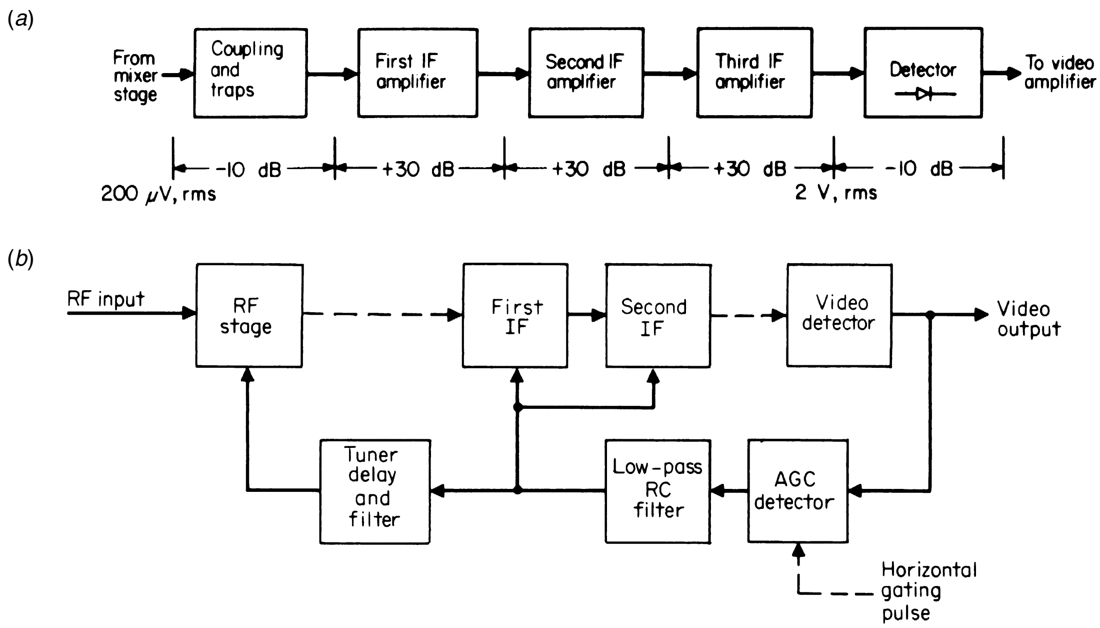


Figure 17.1.9 IF amplifier system: (*a*) typical IF amplifier strip block diagram and gain distribution, (*b*) receiver AGC system.

Automatic gain control (AGC) in a closed feedback loop is used to prevent overload in the IF, and the mixer stage as well, from strong signals (see Figure 17.1.9*b*). Input-signal levels may range from a few microvolts to several hundred microvolts, thus emphasizing the need for AGC. The AGC voltage is applied only to the IF for moderate signal levels so that the low-noise RF amplifier stage in the tuner will operate at maximum gain for relatively weak tuner input signals. A "delay" bias is applied to the tuner gain control to block application of the AGC voltage except at very high antenna signal levels. As the antenna signal level increases, the AGC voltage is applied first to the first and second IF stages. When the input signal reaches about 1 mV, the AGC voltage is applied to the tuner, as well.

Response Characteristics

The bandpass of the IF amplifier must be wide enough to cover the picture and sound carriers of the channel selected in the tuner while providing sharp rejection of adjacent channel signals. Specifically the upper adjacent-picture carrier and lower adjacent-sound channel must be attenuated 40 and 50 dB, respectively, to eliminate visible interference patterns in the picture. The sound carrier at 4.5 MHz below the picture carrier must be of adequate level to feed either a separate sound IF channel or a 4.5 MHz intercarrier sound channel. Furthermore, because in the vestigial sideband system of transmission the video carrier lower sideband is missing, the response characteristic is modified from flat response to attenuate the picture carrier by 50 percent (6 dB).

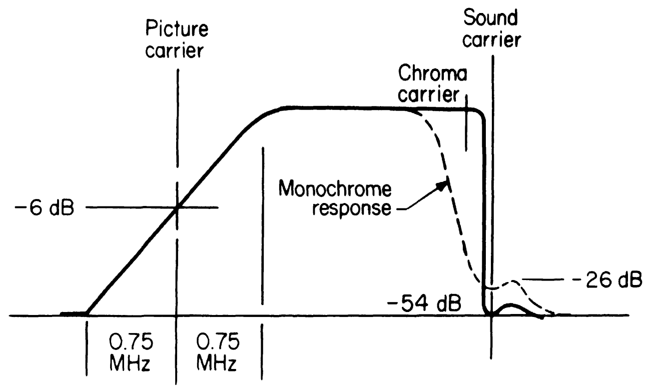


Figure 17.1.10 Ideal IF amplitude response for color and monochrome reception.

In addition, in color receivers the color subcarrier at 3.58 MHz below the picture carrier must be amplified without time delay relative to the picture carrier or distortion. Ideally, this requires the response shown in Figure 17.1.10. Notice that the color IF is wider and has greater attenuation of the channel sound carrier in order to reduce the visibility of the 920 kHz beat between the color subcarrier and the sound carrier.

These and other more stringent requirements for color reception are illustrated in Figure 17.1.11. Specifically:

- IF bandwidth must be extended on the high-frequency video side to accommodate the color subcarrier modulation sidebands that extend to 41.25 MHz (as shown in Figure 17.1.11). The response must be stable and, except in sets with *automatic chroma control* (ACC), the response must not change with the input signal level (AGC), in order to maintain a constant level of color saturation.
- More accurate tuning of the received signal must be accomplished in order to avoid shifting the carriers on the tuner IF passband response. Deviation from their prescribed positions will alter the ratio of luminance to chrominance (saturation). While this is corrected in receivers with *automatic fine tuning* (AFT) and ACC, it can change the time relationship between color and luminance that is apparent in the color picture as chroma being misplaced horizontally.
- Color subcarrier presence as a second signal dictates greater freedom from overload of amplifier and detector circuits, which can result in spurious intermodulation signals visible as beat patterns. These cannot be removed by subsequent filtering.
- Envelope delay (time delay) of the narrow-band chroma and wide-band luminance signals must be equalized so that the horizontal position of the two signals match in the color picture.

Surface Acoustic Wave (SAW) Filter

A SAW filter can provide the entire passband shape and adjacent-channel attenuation required for a television receiver. A typical amplitude response and group delay characteristic are shown

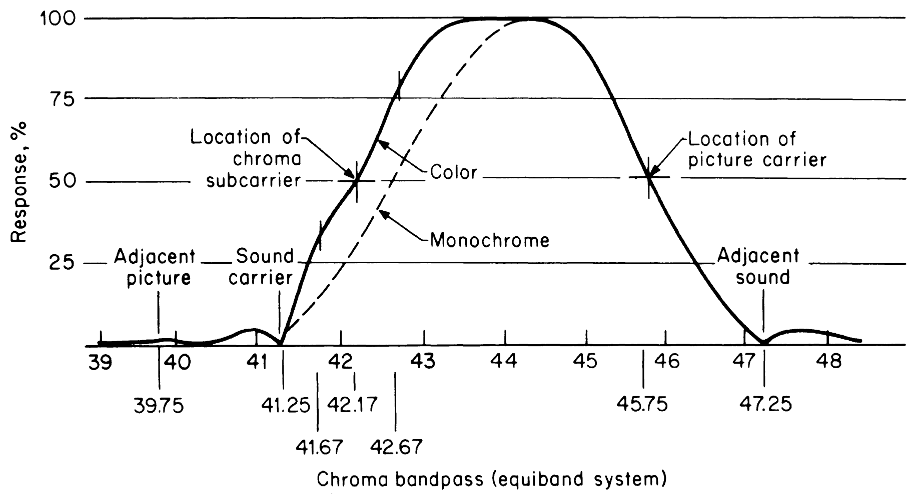


Figure 17.1.11 Overall IF bandwidth for color reception.

in Figure 17.1.12. The sound carrier (41.25 MHz) attenuation of the SAW filter has been designed to operate with a synchronous detector, hence the lesser attenuation than in a conventional LC bandpass filter. The response of an LC discrete stage configuration shows a 60 dB attenuation of the sound carrier, necessary for suppression of the 920 kHz sound-chroma beat when used with a diode detector. In addition, with SAW technology it is possible to make wider adjacent-channel traps, which improve their performance and—in part—makes allowance for the temperature coefficient of the substrate materials used in SAW filters. This drift may be as great as 59 kHz per 10°C.

The schematic diagram of a SAW filter IF circuit is shown in Figure 17.1.13. The filter typically has an insertion loss of 15 to 20 dB and therefore requires a preamplifier to maintain a satisfactory overall receiver SNR.

The SAW filter consists of a piezoelectric substrate measuring 4 to 8 mm by 0.4 mm thick, upon which has been deposited a pattern of two sets of interleaved aluminum fingers. The width of the fingers may be on the order of 50 to 500 mm thick and 10 to 20 μm wide. (See Figure 17.1.14.

Although quartz and other materials have been researched for use as SAW filter substrates, for television applications lithium niobate and lithium tantalate are typical. When one set is driven by an electric signal voltage, an acoustic wave moves across the surface to the other set of fingers that are connected to the load. The transfer amplitude-frequency response appears as a (sin x)/x (Figure 17.1.15).

A modification to the design in Figure 17.1.14 that gives more optimum television bandpass and trap response consists of varying the length of the fingers to form a diamond configuration, illustrated in Figure 17.1.16. This is equivalent to connecting several transducers with slightly different resonant frequencies and bandwidths in parallel. Other modification consists of varying the aperture spacings, distance between the transducers, and the passive coupler strip patterns in the space between the transducers.

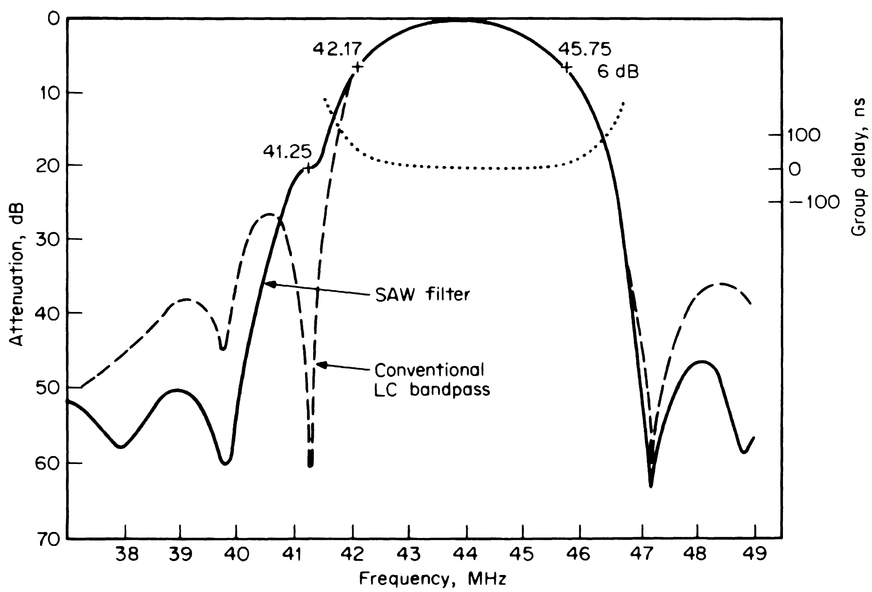


Figure 17.1.12 Surface acoustic wave (SAW) filter response.

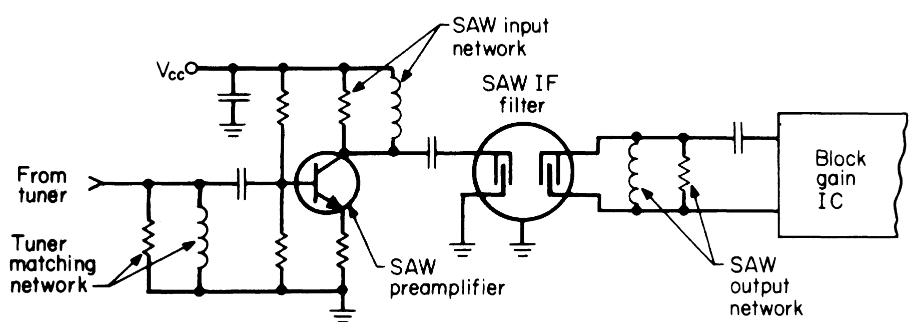


Figure 17.1.13 Schematic of a block filter/block gain configuration.

Gain Block Requirements

Integrated-circuit gain blocks have the same basic requirements as discrete stage IF amplifiers (see Figure 17.1.17). These are: high gain, low distortion, and a large linear gain-control range under all operating conditions. One typical differential amplifier configuration used in IC gain blocks that meets these objectives has a gain of nearly 20 dB and a gain-control range of 24 dB. A direct-coupled cascade of three stages yields an overall gain of 57 dB and a gain-control range of 64 dB. The gain-control system internal to this IC begins to gain-reduce the third stage at an

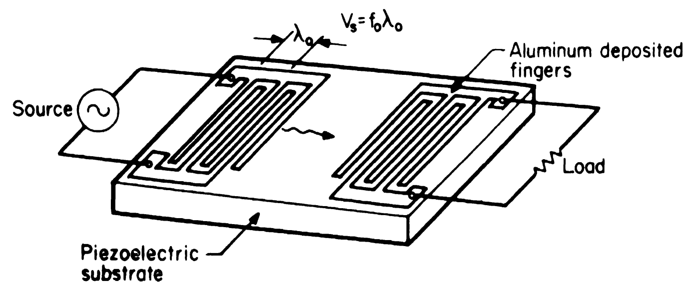


Figure 17.1.14 Physical implementation of a SAW filter.

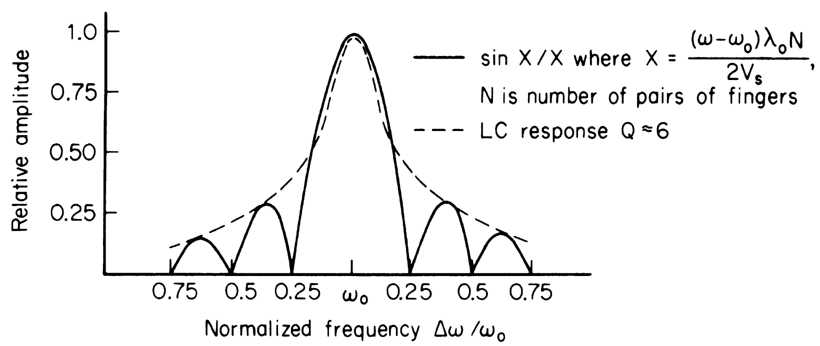


Figure 17.1.15 Response of a uniform interdigitated SAW filter. (After [2].)

IC input level of 100 μV of IF carrier. With increasing input signal level, the third stage gain reduces to 0 dB and then is followed by the second stage to a similar level, followed by the first in the same manner. By this means, a noise figure of 7 dB is held constant over an IF input signal range of 40 dB. The need for a preamplifier ahead of the SAW IF becomes less important when the IF amplifier noise figure is maintained constant by this cascaded control system.

The high gain and small size of an integrated-circuit IF amplifier places greater importance on PC layout techniques and ground paths if stability is to be achieved under a wide range of operating conditions. These considerations also carry over to the external circuits and components.

17.1.2d Video Signal Demodulation

The function of the video demodulator is to extract the picture signal information that has been placed on the RF carrier as amplitude modulation. The demodulator receives the modulated carrier signal from the IF amplifier that has boosted the peak-to-peak level to 1 or 2 V. The modula-

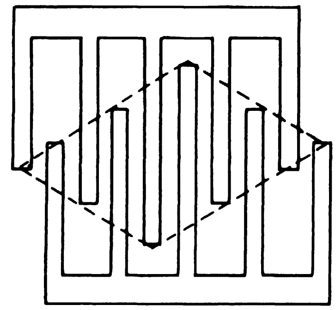


Figure 17.1.16 SAW apodized IDT pattern.

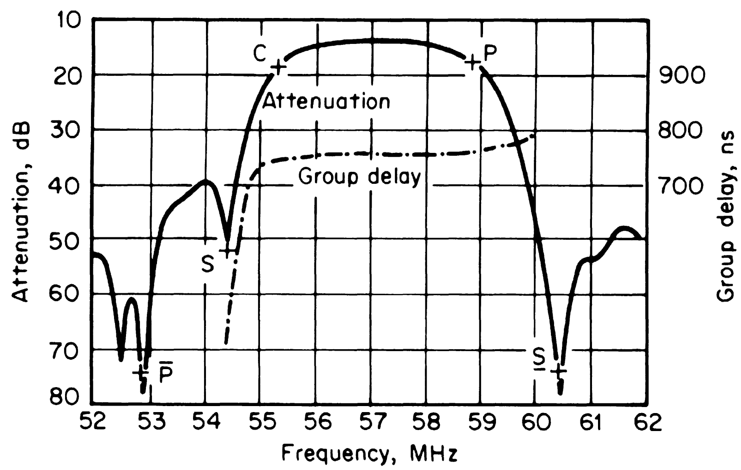


Figure 17.1.17 Frequency response of SAW filter picture-carrier output. (After [3].)

tion components extend from dc to 4.5 MHz. The output of the demodulator is fed directly to the video amplifier.

There are four types of demodulators commonly used in television receivers:

- Envelope detector
- Transistor detector
- Synchronous detector
- Feedback balanced diode

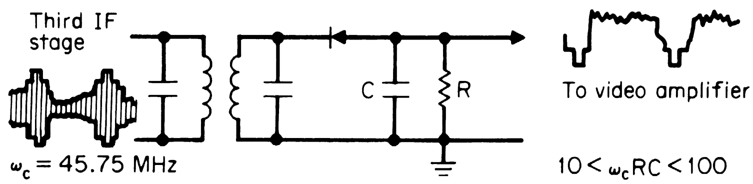


Figure 17.1.18 Television receiver envelope detector.

Envelope Detector

Of the several types of demodulators, the envelope detector is the simplest. It consists of a diode rectifier feeding a parallel load of a resistor and a capacitor (Figure 17.1.18). In other words, it is a half-wave rectifier that charges the capacitor to the peak value of the modulation envelope.

Because of the large loss in the diode, a high level of IF voltage is required to recover 1 or 2 volts of demodulated video. In addition, unless the circuit is operated at a high signal level, the curvature of the diode impedance curve near cutoff results in undesirable compression of peak-white signals. The requirement for large signal levels and the nonlinearity of detection result in design problems and certain performance deficiencies, including the following:

- Beat signal products will occur between the color subcarrier (42.17 MHz), the sound carrier (41.25 MHz), and high-amplitude components in the video signal. The most serious is a 920 Hz (color to sound) beat and 60 Hz buzz in sound from vertical sync and peak-white video modulation.
- Distortion of luminance toward black of as much as 10 percent and asymmetric transient response. Referred to as *quadrature distortion*, this characteristic of the vestigial sideband is aggravated by nonlinearity of the diode. (See Figure 17.1.19.)
- Radiation of the fourth harmonic of the video IF produced by the detection action directly from the chassis, which can interfere with reception of VHF Channel 8 (180 to 186 MHz).

Even with these deficiencies, the diode envelope detector was used in the majority of the monochrome and color television receivers dating back to vacuum tube designs up to the era of discrete transistors.

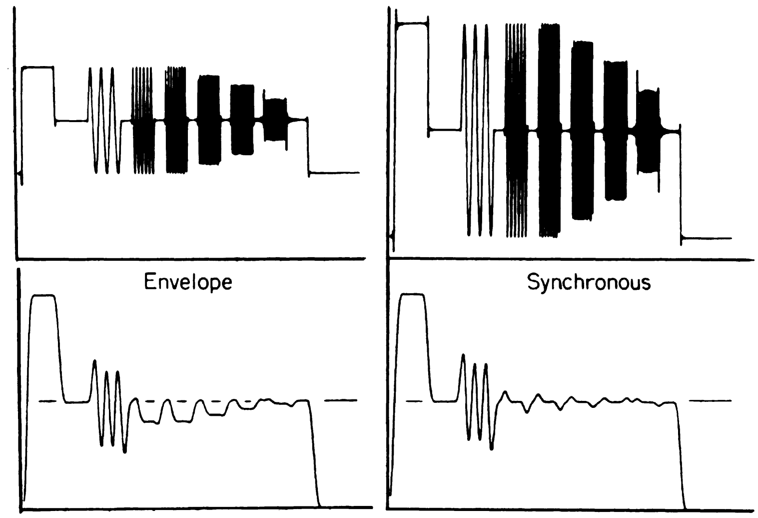
Transistor Detector

A transistor biased near collector cutoff and driven with a modulated carrier at an amplitude greater than the bias level (see Figure 17.1.20) provides a demodulator that can have a gain of 15 or 20 dB over that of a diode. Consequently, less gain is required in the IF amplifier; in fact, in some receiver designs, the third IF stage has been eliminated. Unfortunately, this is offset by the same deficiencies in signal detection as the diode envelope detector.

Synchronous Detector

The synchronous detector is basically a balanced rectifier in which the carrier is sampled by an unmodulated carrier at the same frequency as the modulated carrier. The unmodulated reference

(*a*)



(*b*)

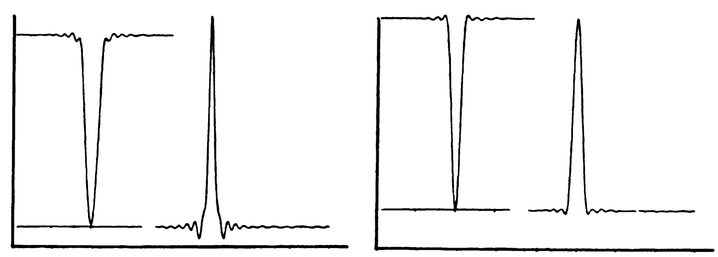


Figure 17.1.19 Comparison of quadrature distortion in envelope and synchronous detectors: (*a*) axis shift, (*b*) inverted and normal 2T-pulse response. (*After* [4].)

signal is generated in a separate high-Q limiting circuit that removes the modulation. An alternative system for generating the reference waveform is by means of a local oscillator phase-locked to the IF signal carrier.

The advantages of synchronous demodulation are:

- Higher gain than a diode detector
- Low level input, which considerably reduces beat-signal generation
- Low level detection operation, which reduces IF harmonics by more than 20 dB
- Little or no quadrature distortion (see Figure 17.1.19), depending upon the lack of residual phase modulation (purity) of the reference carrier used for detection

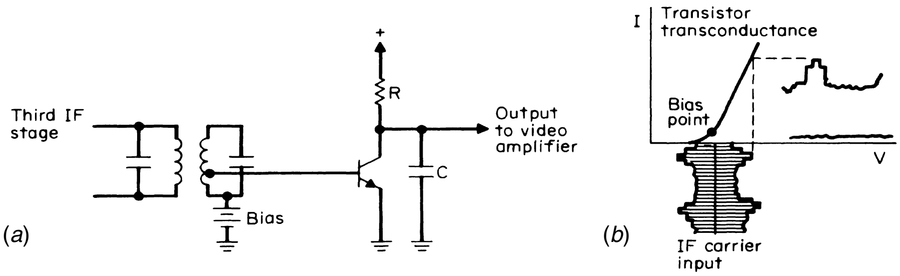


Figure 17.1.20 Transistor detector: (*a*) schematic diagram, (*b*) detection characteristic.

- Circuit easily formatted in an IC

17.1.2e Automatic Gain Control (AGC)

Each amplifier, mixer, and detector stage in a television receiver has a number of operating conditions that must be met in order to achieve optimum performance. Specifically, these conditions include that:

- The input level is greater than the internally generated noise by a factor in excess of the minimum acceptable SNR.
- Input level does not overload the amplifier stages, thus causing a bias shift.
- Bias operates each functional component of the system at its optimum linearity point, that is, the lowest third-order product for amplifiers, and highest practical second-order for mixers and detectors.
- Spurious responses in the output are 50 dB below the desired signal.

The function of the automatic gain control system is to maintain signal levels in these stages at the optimum value over a large range of receiver input levels. The control voltage to operate the system usually is derived from the video detector or the first video amplifier stage. Common implementation techniques include the following:

- Average AGC, which operates on the principle of keeping the carrier level constant in the IF. Changes in modulation of the carrier will affect the gain control, and therefore it is used only in low-cost receivers.
- Peak or sync-clamp AGC, which compares the video sync-tip level with a fixed dc level. If the sync-tip amplitude exceeds the reference level, a control voltage is applied to the RF and IF stages to reduce their gain and thus restore the sync-tip level to the reference level.
- Keyed or gated AGC, which is similar to sync-clamp AGC. The stage where the comparison of sync-tip and reference signals takes place is activated only during the sync-pulse interval by a horizontal flyback pulse. Because the AGC circuit is insensitive to spurious noise signals between sync pulses, the noise immunity is considerably improved over the other two systems.

AGC Delay

For best receiver SNR, the tuner RF stage is operated at maximum gain for RF signals up to a threshold level of 1 mV. In discrete amplifier chains, the AGC system begins to reduce the gain of the second IF stage proportionately as the RF signal level increases from just above the sensitivity level to the second-stage limit of gain reduction (20 to 25 dB). For increasing signals, the first IF stage gain is reduced. Finally, above the control delay point of 1mV, the tuner gain is reduced. A plot of the relationships between receiver input RF level and the gain characteristics of the tuner and IF are shown in Figure 17.1.21*a* and the noise figure is shown in Figure 17.1.21*b*.

System Analysis

The interconnection of the amplifier stages (RF, mixer, and IF), the detector, and the lowpass-filtered control voltage is—in effect—a feedback system. The loop gain of the feedback system is nonlinear, increasing with increasing signal level. There are two principal constraints that designers must cope with:

- First, the loop gain should be large to maintain good regulation of the detector output over a wide range of input signal levels. As the loop gain increases, the stability of the system will decrease and oscillation can occur.
- Second, the ability of a receiver to reject impulse noise is inversely proportional to the loop gain. Excessive impulse noise can saturate the detector and reduce the RF-IF gain, thereby causing either a loss in picture contrast or a complete loss of picture. This problem can be alleviated by bandwidth-limiting the video signal fed to the AGC detector, or the use of keyed or gated AGC to block false input signals (except during the sync pulse time).

A good compromise between regulation of video level and noise immunity is realized with a loop-gain factor of 20 to 30 dB.

The filter network and filter time constants play an important part in the effectiveness of AGC operation. The filter removes the 15.750 kHz horizontal sync pulses and equalizing pulses, the latter in blocks in the 60 Hz vertical sync interval. The filter time constants must be chosen to eliminate or minimize the following problems:

- *Airplane flutter*, a fluctuation in signal level caused by alternate cancellation and reinforcement of the received signal by reflections from an airplane flying overhead in the path between the transmitting and receiving antennas. The amplitude may vary as much as 2 to 1 at rates from 50 to 150 Hz. If the time constants are too long, especially that of the control voltage to the RF stage, the gain will not change rapidly enough to track the fluctuating level of the signal. The result will be a flutter in contrast and brightness of the picture.
- *Vertical sync pulse sag*, resulting from the AGC system speed of response being so fast that it will follow changes in sync pulse energy during the vertical sync interval. Gain increases during the initial half-width equalizing pulses, then decreases during the slotted vertical sync pulse, increases again during the final equalizing pulses, and then returns to normal during the end of vertical blanking and the next field of picture. Excessive sag can cause loss of proper interlace and vertical jitter or bounce. Sag can be reduced by limiting the response of the AGC control loop, or through the use of keyed AGC.

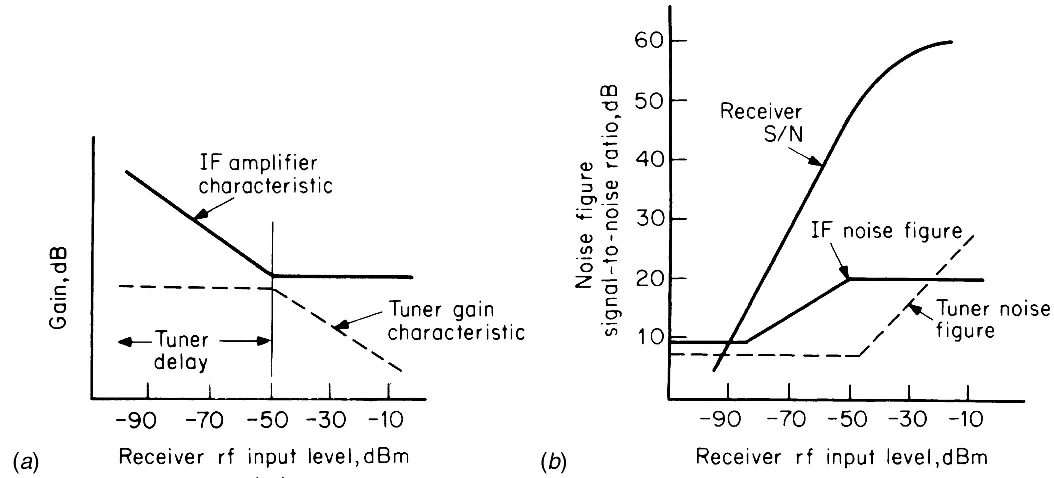


Figure 17.1.21 Automatic gain control principles: (*a*) gain control as a function of input level, (*b*) noise figure of RF and IF stages with gain control and the resulting receiver SNR.

- Lock-out of the received signal during channel switching, caused by excessive speed of the AGC system. This can result in as much as a 2:1 decrease in pull-in range for the AGC system. In keyed AGC systems, if the timing of the horizontal gating pulses is incorrect, excessive or insufficient sync can result at the sync separator, which—in turn—will upset the operation of the AGC loop.

17.1.2f Automatic Frequency Control (AFC)

Also called *automatic fine-tuning* (AFT), the AFC circuit senses the frequency of the picture carrier in the IF section and sends a correction voltage to the local oscillator in the tuner section if the picture carrier is not on the standard frequency of 45.75 MHz.

Typical AFC systems consist of a frequency discriminator prior to the video detector, a low-pass filter, and a varactor diode controlling the local oscillator. The frequency discriminator in discrete transistor IF systems has typically been the familiar balanced-diode type used for FM radio receivers with the components adjusted for wide-band operation centering on 45.75 MHz. A small amount of unbalance is designed into the circuit to compensate for the unbalanced sideband components of vestigial sideband signal characteristics. The characteristics of AFC closed loops are shown in Table 17.1.5.

In early solid-state designs, the AFC block was a single IC with a few external components. Current designs have included the AFC circuit in the form of a synchronous demodulator on the same IC die as the other functions of the IF section.

Table 17.1.5 Typical AFC Closed-Loop Characteristics

Parameter	Value
Pull-in range	±750 kHz
Hold-in range	±1.5 MHz
Frequency error for ±500 kHz offset	< 50 kHz

17.1.2g Sound Carrier Separation Systems

Television sound is transmitted as a frequency-modulated signal with a maximum deviation of ±25 kHz (100 percent modulation) capable of providing an audio bandwidth of 50 to 15,000 Hz. The frequency of the sound carrier is 4.5 MHz above the RF picture carrier. The basic system is monaural with dual-channel stereophonic transmission at the option of the broadcaster.

The *intercarrier* sound system passes the IF picture and sound carriers (45.75 and 41.25 MHz, respectively) through a detector (nonlinear stage) to create the intermodulation beat of 4.5 MHz. The intercarrier sound signal is then amplified, limited, and FM-demodulated to recover the audio. Block diagrams of intercarrier sound systems are shown in Figures 17.1.22 and 17.1.23. In a discrete component implementation, the intercarrier detector is typically a simple diode detector feeding a 4.5 MHz resonant network. If an IC IF system is used, the sound and picture IF signals are carried all the way to the video detector, where one output port of the balanced synchronous demodulator supplies both the 4.5 MHz sound carrier and the composite baseband video.

The coupling network between the intercarrier detector and the sound IF amplifier usually has the form of a half-section high-pass filter that is resonant at 4.5 MHz. This form gives greater attenuation to the video and sync pulses in the frequency range from 4.5 MHz to dc (carrier), thereby reducing buzz in the recovered audio, especially under the conditions of low picture carrier. An alternative implementation uses a piezoelectric ceramic filter that is designed to have a bandpass characteristic at 4.5 MHz and needs no in-circuit adjustment.

Audio buzz results when video-related phase-modulated components of the visual carrier are transferred to the sound channel. The generation of *incidental carrier phase modulation* (ICPM) can be transmitter-related or receiver-related; however, the transfer to the sound channel takes place in the receiver. This can occur in the mixer or the detection circuit. Here, a synchronous detector represents little improvement over an envelope diode detector unless a narrow-band filter is used in the reference channel [5].

Split-carrier sound processes the IF picture and sound carriers as shown in Figure 17.1.24.

Quasi-parallel sound utilizes a special filter such as the SAW filter of Figure 17.1.14 to eliminate the Nyquist slope in the sound detection channel, thereby eliminating a major source of ICPM generation in the receiver. The block diagram of this system is shown in Figure 17.1.25.

Nearly all sound channels in present-day television receivers are designed as a one- or two-IC configuration. The single IC contains the functions of sound IF amplifier-limiter, FM detector, volume control, and audio output. Two-chip systems usually incorporate stereo functionality or audio power amplification.

Four types of detector circuits typically are used in ICs for demodulation of the FM sound carrier:

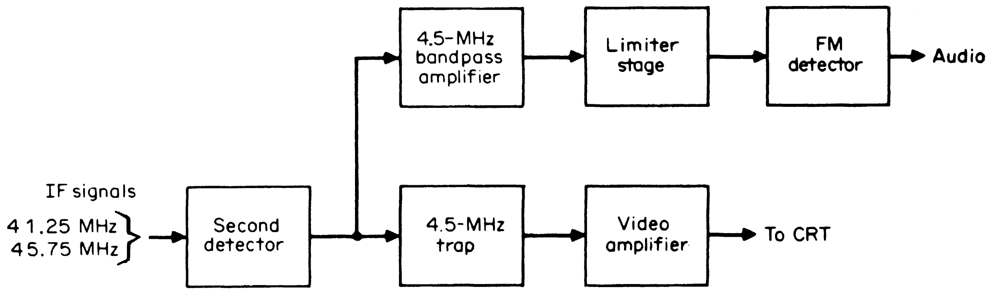


Figure 17.1.22 A typical monochrome intercarrier sound system.

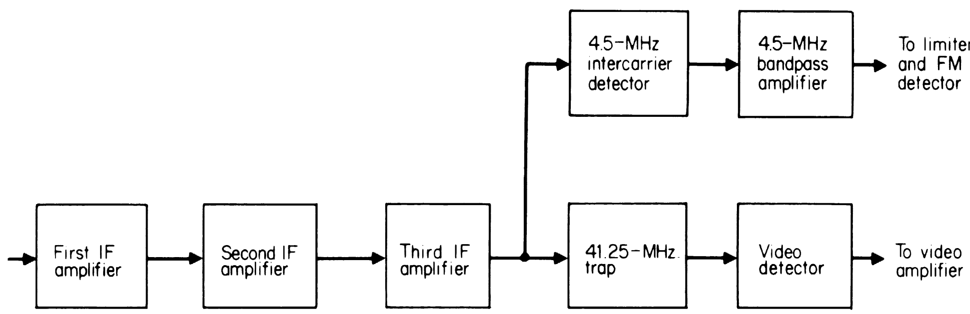


Figure 17.1.23 An intercarrier sound takeoff ahead of the video detector.

- The *quadrature detector*, also known as the *gated coincidence detector* and *analog multiplier*, measures the instantaneous phase shift across a reactive circuit as the carrier frequency shifts. At center frequency (zero deviation) the LC phase network gives a 90'° phase shift to V_2 compared with V_1. As the carrier deviates, the phase shift changes proportionately to the amount of carrier deviation and direction.
- The *balanced peak detector*, which utilizes two peak or envelope detectors, a differential amplifier, and a frequency-selective circuit or piezoceramic discriminator.
- The *differential peak detector*, which operates at a low voltage level and does not require square-wave switching pulses. Therefore, it creates less harmonic radiation than the quadrature detector. In some designs, a low-pass filter is placed between the limiter and peak detector to further reduce harmonic radiation and increase AM rejection.
- The *phase-locked-loop detector*, which requires no frequency-selective LC network to accomplish demodulation. In this system, the voltage-controlled oscillator (VCO) is phase-

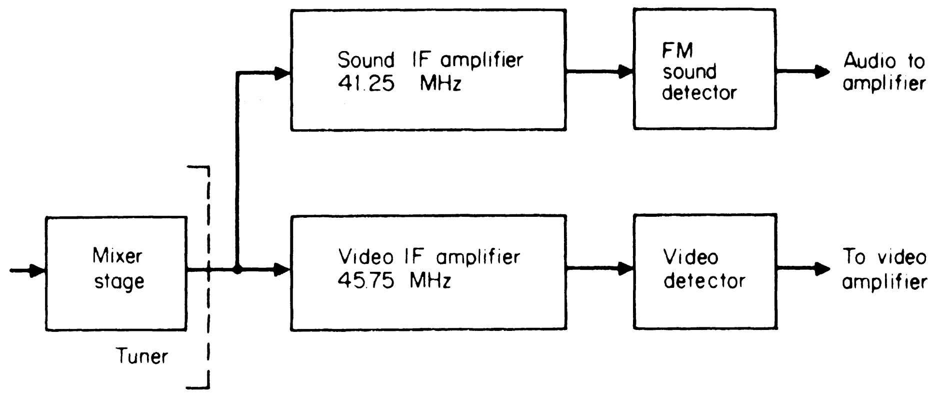


Figure 17.1.24 Split-carrier sound system.

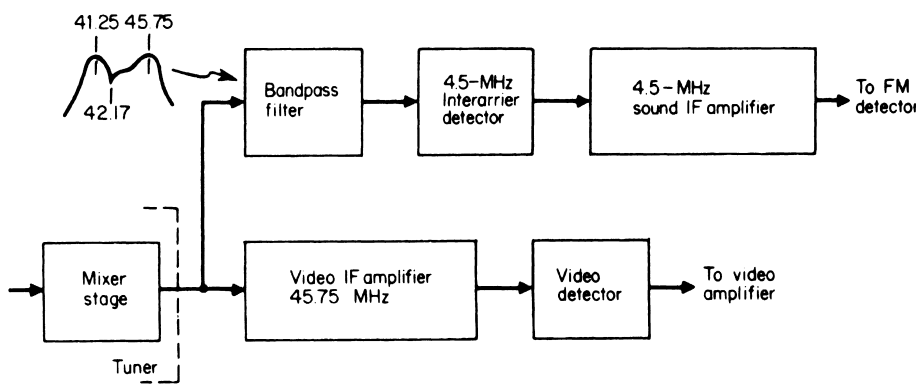


Figure 17.1.25 Quasi-parallel sound system.

locked by the feedback loop into following the deviation of the incoming FM signal. The low-frequency error voltage that forces the VCO to track is—in fact—the demodulated output.

17.1.2h Video Amplifiers

A range of video signals of 1 to 3 V at the second detector has become standard for many practical reasons, including optimum gain distribution between RF, IF, and video sections and distribution of signal levels so that video detection and sync separation may be effectively performed. The video amplifier gain and output level are designed to drive the picture tube with this input level. Sufficient reserve is provided to allow for low percentage modulation and signal strengths below the AGC threshold.

Picture Controls

A video gain or contrast control and a brightness or background control are provided to allow the viewer to select the contrast ratio and overall brightness level that produce the most pleasing picture for a variety of scene material, transmission characteristics, and ambient lighting conditions. The contrast control usually provides a 4:1 gain change. This is accomplished either by attenuator action between the output of the video stage and the CRT or by changing the ac gain of the video stage by means of an ac-coupled variable resistor in the emitter circuit. The brightness control shifts the dc bias level on the CRT to raise or lower the video signal with respect to the CRT beam cutoff voltage level. (See Figure 17.1.26.)

AC and dc Coupling

For perfect picture transmission and reproduction, it is necessary that all shades of gray are demodulated and reproduced accurately by the display device. This implies that the dc level developed by the video demodulator, in response to the various levels of video carrier, must be carried to the picture tube. Direct coupling or dc restoration is often used, especially in color receivers where color saturation is directly dependent upon luminance level. (See Figure 17.1.27.)

Many low cost monochrome designs utilize only ac coupling with no regard for the dc information. This eases the high-voltage power supply design as well as simplifying the video circuitry. These sets will produce a picture in which the average value of luminance remains nearly constant. For example, a night scene having a level of 15 to 20 IRE units and no peak-white excursions will tend to brighten toward the luminance level of the typical daytime scene (50 IRE units). Likewise a full-raster white scene with few black excursions will tend to darken to the average luminance level condition by use of partial dc coupling in which a high-resistance path exists between the second detector and the CRT. This path usually has a gain of one-half to one-fourth that of the ac signal path.

The transient response of the video amplifier is controlled by its amplitude and phase characteristics. The low-frequency transient response, including the effects of dc restoration, if used, is measured in terms of distortion to the vertical blanking pulse. Faithful reproduction requires that the change in amplitude over the pulse duration, usually a decrease from initial value called *sag* or *tilt*, be less than 5 percent. In general, there is no direct and simple relationship between the sag and the lower 3 dB cutoff frequency. However, lowering the 3 dB cutoff frequency will reduce the tilt, as illustrated in Figure 17.1.28.

Low-Frequency Response Requirements

The effect of inadequate low-frequency response appears in the picture as vertical shading. If the response is so poor as to cause a substantial droop of the top of the vertical blanking pulse, then incomplete blanking of retrace lines can occur.

It is not necessary or desirable to extend the low-frequency response to achieve essentially perfect LF square-wave reproduction. First, the effect of tilt produced by imperfect LF response is modified if dc restoration is employed. Direct-current restorers, particularly the fast-acting variety, substantially reduce tilt, and their effect must be considered in specifying the overall response. Second, extended LF response makes the system more susceptible to instability and low-frequency interference. Current coupling through a common power supply impedance can produce the low-frequency oscillation known as “motorboating.” Motorboating is not usually a

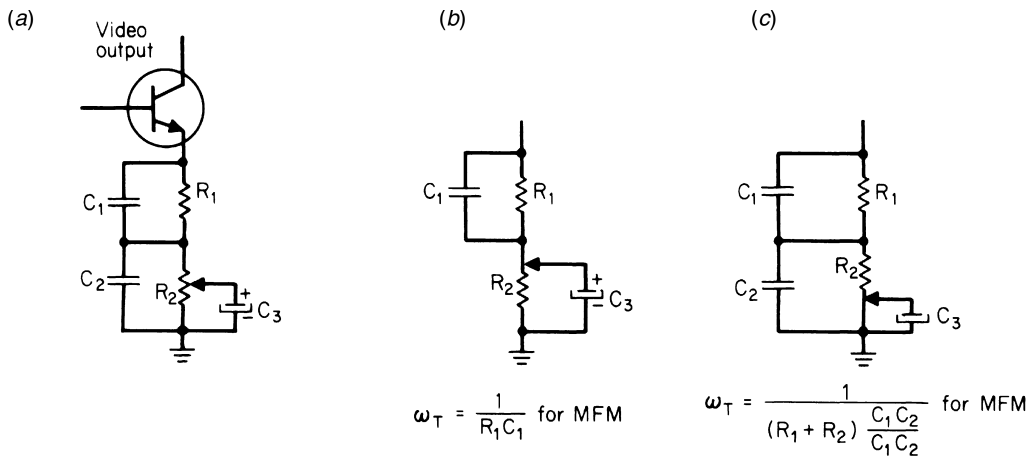


Figure 17.1.26 Contrast control circuits: (*a*) contrast control network in the emitter circuit, (*b*) equivalent circuit at maximum contrast (maximum gain), (*c*) minimum contrast.

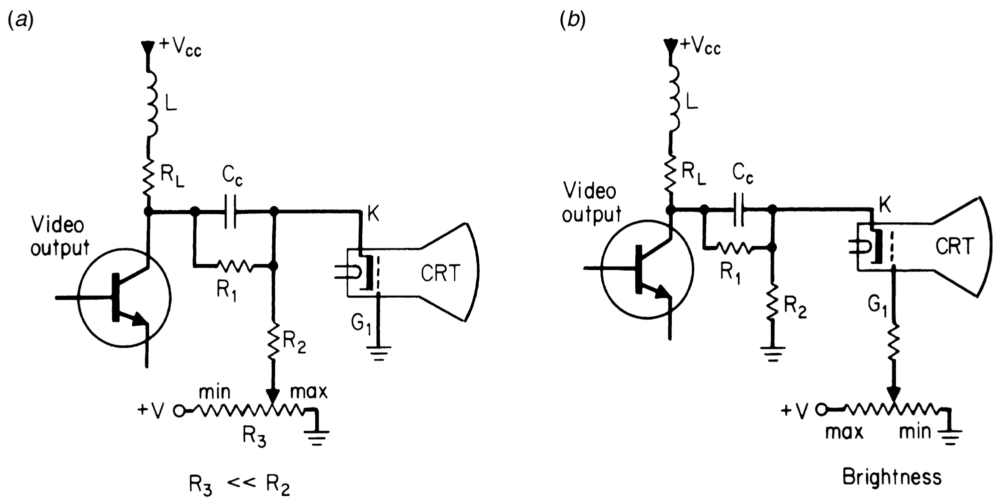


Figure 17.1.27 CRT luminance drive circuit: (*a*) brightness control in CRT cathode circuit, (*b*) brightness control in CRT grid circuit.

problem in television receiver video amplifiers because they seldom employ the number of stages required to produce regenerative feedback, but in multistage amplifiers the tendency toward motorboating is reduced as the LF response is reduced.

A more commonly encountered problem is the effect of airplane flutter and *line bounce*. Although a fast-acting AGC can substantially reduce the effects of low-frequency amplitude

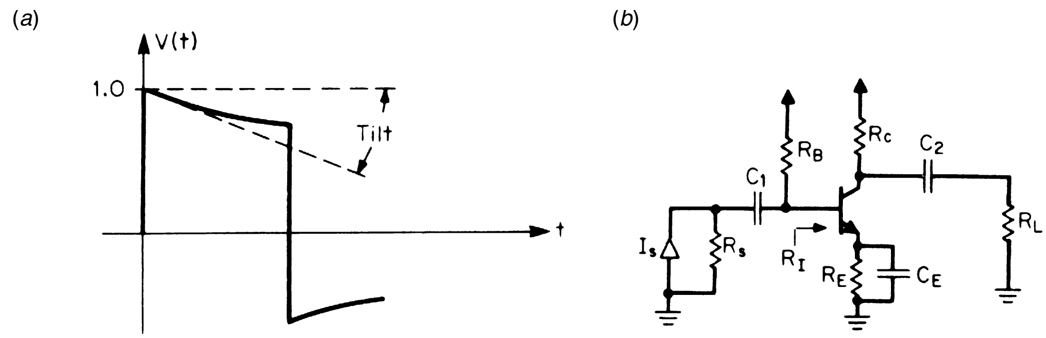


Figure 17.1.28 Video stage low-frequency response: (*a*) square-wave output showing tilt, (*b*) RC time constant circuits in the common-emitter stage that affect low-frequency response.

variations produced by airplane reflections, the effect is so annoying visually as to warrant a sacrifice in LF response to bring about further reduction. A transient in-line voltage amplitude, commonly called a line bounce, also can produce an annoying brightness transient that can similarly be reduced through a sacrifice of LF response. Special circuit precautions against line bounce include the longest possible power supply time constant, bypassing the picture tube electrodes to the supply instead of ground, and the use of coupling networks to attenuate the response sharply below the LF cutoff frequency. The overall receiver response is usually an empirically determined compromise.

The high-frequency transient characteristic is usually expressed as the amplifier response to an ideal input voltage or current step. This response is shown in Figure 17.1.29 and described in the following terms:

- Rise time τ_R is the time required for the output pulse to rise from 10 to 90 percent of its final (steady-state) value.
- Overshoot is the amplitude by which the transient rise exceeds its final value, expressed as a percentage of the final value.
- Preshoot is the amplitude by which the transient oscillatory output waveform exceeds its initial value.
- Smear is an abnormally slow rise as the output wave approaches its final value.
- Ringing is an oscillatory approach to the final value.

In practice, rise times of 0.1 to 0.2 μs are typical. Overshoot, smear, and ringing amplitude are usually held to 5 percent of the final value, and ringing is restricted to one complete cycle.

17.1.2i Color Receiver Luminance Channel

Suppression of the chroma subcarrier is necessary to reduce objectionable dot crawl in and around colored parts of the picture, as well as reduce the distortion of luminance levels resulting from the nonlinear transfer characteristic of CRT electron guns. Traditionally, a simple high-*Q*

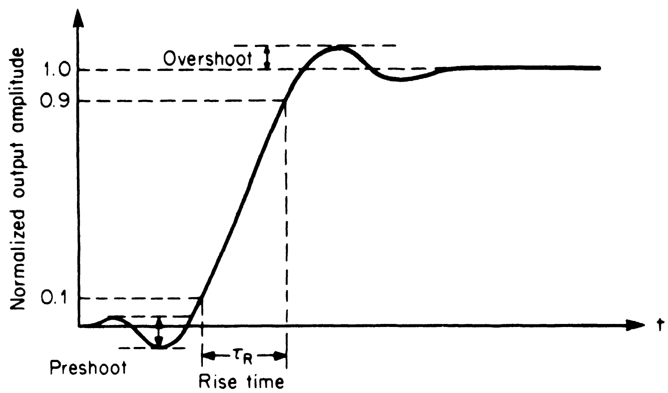


Figure 17.1.29 Video amplifier response to a step input.

LC trap, centered around the color subcarrier, has been used for rejection, but this necessitates a trade off between luminance channel bandwidth and the stop band for the chroma sidebands. Luminance channel comb filtering largely avoids this compromise and is one reason why it is commonly used.

The luminance channel also provides the time delay required to correct the time delay registration with the color difference signals, which normally incur delays in the range of from 300 to 1000 ns in their relatively narrow-bandwidth filters.

While delay circuits having substantially flat amplitude and group delay out to the highest baseband frequency of interest can and have been used, this is not necessarily required nor desirable for cost-effective overall design. Because the other links in the chain (i.e., tuner, IF, traps at 4.5 and 3.58 MHz, and CRT driver stage) may all contribute significant linear distortion individually, it is frequently advantageous to allow these distortions to occur and use the delay block as an overall group delay and/or amplitude equalizer.

Although it is well known that, for “distortionless” transmission, a linear system must possess both uniform amplitude and group delay responses over the frequency band of interest, the limitations of a finite bandwidth lead to noticeably slower rise and fall times, rendering edges less sharp or distinct. By intentionally distorting the receiver amplitude response and boosting the relative response to the mid and upper baseband frequencies to varying degrees, both faster rise and fall times can be developed along with enhanced fine detail. If carried too far, however, objectionable outlining can occur, especially to those transients in the white direction. Furthermore, the visibility of background noise is increased.

For several reasons—including possible variations in transient response of the transmitted signal, distortion due to multipath, antennas, receiver tolerances, SNR, and viewer preference—it is difficult to define a fixed response at the receiver that is optimum under all conditions. Therefore, it is useful to make the amplitude response variable so it can be controlled to best suit the individual situation. Over the range of adjustment, it is assumed that the overall group delay shall remain reasonably flat across the video band. The exact shape of the amplitude response is directly related to the desired time domain response (height and width of preshoot, overshoot, and ringing) and chroma subcarrier sideband suppression.

Because the peaked signal later operates on the nonlinear CRT gun characteristics, large white preshoots and overshoots can contribute to excessive beam currents, which can cause CRT spot defocusing. To alleviate this, circuits have been developed that compress large excursions of the peaking component in the white direction. For best operation, it is desirable that the signals being processed have equal preshoot and overshoot.

Low level, high frequency noise in the luminance channel can be removed by a technique called *coring*. One coring technique involves nonlinearly operating on the peaking or edge-enhancement signal, discussed earlier in this section. The peaking signal is passed through an amplifier having low or essentially no gain in the mid-amplitude range. When this modified peaking signal is added to the direct video, the large transitions will be enhanced, but the small ones (noise) will not be, giving the illusion that the picture sharpness has been increased while the noise has been decreased.

17.1.2j Chroma Subcarrier Processing

In the *equiband chroma* system, typical of practically all consumer receivers, the chroma amplifier must be preceded by a bandpass filter network that complements the chroma sideband response produced by the tuner and IF. Frequencies below 3 MHz also must be attenuated to reduce not only possible video cross-color disturbances but also crosstalk caused by the lower-frequency *I* channel chroma information.

Another requirement is that the filter have a gentle transition from passband to stop band in order to impart a minimum amount of group delay in the chroma signal, which then must be compensated by additional group delay circuitry in the luminance channel. The fourth-order high-pass filter is a practical realization of these requirements.

As described previously, this stage also serves as the chroma gain control circuit. The usual implementation in an IC consists of a differential amplifier having the chroma signal applied to the current source. The gain-control dc voltage is applied to one side of the differential pair to divert the signal current away from the output side.

Burst Separation

Complete separation of the color synchronizing burst from video requires *time gating*. The gate requirements are largely determined by the horizontal sync and burst specifications, illustrated in Figure 17.1.30. It is essential that all video information be excluded. It is also desirable that both the leading and trailing edges of burst be passed so that the complementary phase errors introduced at these points by quadrature distortion average to zero. Widening the gate pulse to minimize the required timing accuracy has a negligible effect on the noise performance of the reference system and may be beneficial in the presence of echoes. The ≈ 2 μs spacing between the trailing edges of burst and horizontal blanking determines the total permissible timing variation. Noise modulation of the gate timing should not permit noise excursions to encroach upon the burst because the resulting cross modulation will have the effect of increasing the noise power delivered to the reference system.

Burst Gating Signal Generation

The gate pulse generator must provide both steady-state phase accuracy and reasonable noise immunity. The horizontal flyback pulse has been widely used for burst gating because it is

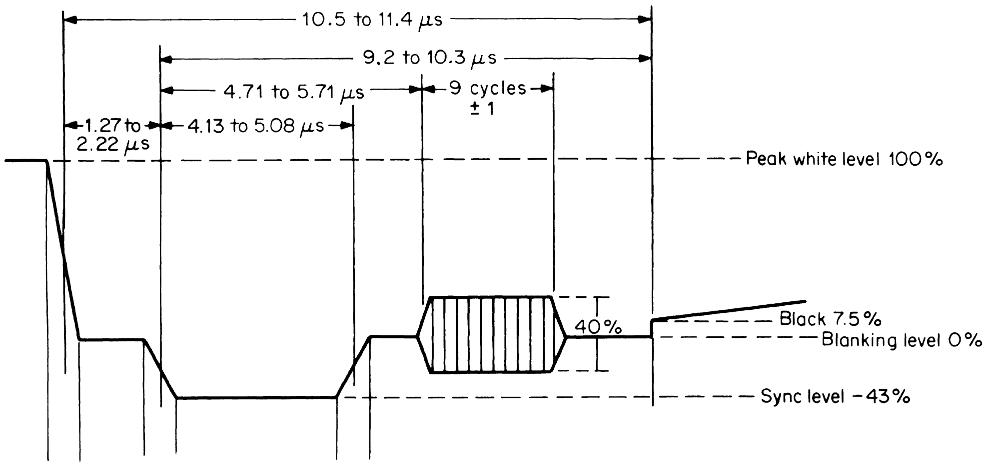


Figure 17.1.30 Horizontal blanking interval specification.

derived from the horizontal scan oscillator system, which meets the noise immunity requirements and, with appropriate design, can approximate the steady-state requirements. A further improvement in steady-state phase accuracy can be achieved by deriving the gating pulse directly from the trailing edge of the horizontal sync pulse. This technique is utilized in several chroma system ICs.

The burst gate in conventional discrete component circuits has the form of a conventional amplifier that is biased into linear conduction only during the presence of the gating pulse. In the IC implementation, the complete chroma signal is usually made available at one input of the *automatic phase control* (APC) burst-reference phase detector. The gating pulse then enables the phase detector to function only during the presence of burst.

Color Subcarrier Reference Separation

The color subcarrier reference system converts the synchronizing bursts to a continuous carrier of identical frequency and close phase tolerance. Theoretically, the long-term and repetitive phase inaccuracies should be restricted to the same value, approximately ±5°. Practically, if transmission variations considerably in excess of this value are encountered, and if operator control of phase ("hue control") is provided, the long-term accuracy need not be so great. Somewhat greater instantaneous inaccuracies can be tolerated in the presence of thermal noise so that an rms phase error specification of 5 to 10° at an S/N of unity may be regarded as typical.

Reference Systems

Three types of reference synchronization systems have been used:

- Automatic phase control of a VCO
- Injection lock of a crystal

- Ringing of a crystal filter

Best performance can be achieved by the APC loop. In typical applications, the figure of merit can be made much smaller (better) for the APC loop than for the other systems by making the factor $(1/y) + m$ have a value considerably less than 1, even as small as 0.1. The parts count for each type system, at one time much higher for the APC system, is no longer a consideration because of IC implementations where the oscillator and phase detector are integrated and only the resistors and capacitors of the filter network and oscillator crystal are external.

The APC circuit is a phase-actuated feedback system consisting of three functional components:

- A phase detector
- Low-pass filter
- DC voltage-controlled oscillator

The overall system is illustrated in Figure 17.1.31 The characteristics of these three units define both the dynamic and static loop characteristics and hence the overall system performance.

The phase detector generates a dc output E whose polarity and amplitude are proportional to the direction and magnitude of the relative phase difference $d\phi$ between the oscillator and synchronizing (burst) signals.

The VCO is an IC implementation that requires only an external crystal and simple phase-shift network. The oscillator can be shifted ±45° by varying the phase-control voltage. This leads to symmetrical pull-in and hold-in ranges.

Chroma Demodulation

The chroma signal can be considered to be made up of two amplitude-modulated carriers having a quadrature phase relationship. Each of these carriers can be individually recovered through the use of a synchronous detector. The reference input to the demodulator is that phase which will demodulate only the I signal. The output contains no Q signal.

Demodulation products of 7.16 MHz in the output signal can contribute to an optical interference moiré pattern in the picture. This is related to the line geometry of the shadow mask. The 7.16 MHz output also can result in excessively high line-terminal radiation from the receiver. A first-order low-pass filter with cutoff of 1 to 2 MHz usually provides sufficient attenuation. In extreme cases, an LC trap may be required.

Demodulation Axis

Over the double sideband region (±500 kHz around the subcarrier frequency) the chrominance signal can be demodulated as pure quadrature AM signals along either the I and Q axis or R–Y and B–Y axis. The latter signal leads to a simpler matrix for obtaining the color drive signals R, G, and B.

Current practice has moved away from the classic demodulation angles for two main reasons:

- Receiver picture tube phosphors have been modified to yield greater light output and can no longer produce the original NTSC primary colors. The chromaticity coordinates of the primary colors, as well as the RGB current ratios to produce white balance, vary from one CRT manufacturer to another.

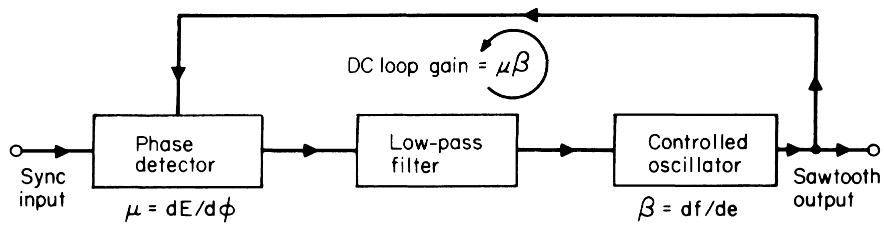


Figure 17.1.31 Automatic phase control (APC) system block diagram.

- The white-point setup has, over the years, moved from the cold 9300 K of monochrome tubes to the warmer illuminant C and D65, which produce more vivid colors, more representative of those that can be seen in natural sunlight.

17.1.2k RGB Matrixing and CRT Drive

Because RGB primary color signals driving the display are required as the end output in a television receiver, it is necessary to combine or "matrix" the demodulated color-difference signals with the luminance signal. Several circuit configurations can be used to accomplish this task.

In the color-difference drive matrixing technique, R–Y, G–Y, and B–Y signals are applied to respective control grids of the CRT while luminance is applied to all three cathodes; the CRT thereby matrixes the primary colors. This approach has the advantage that gray scale is not a function of linearity matching among the three channels because at any level of gray, the color-difference driver stages are at the same dc level. Also, because the luminance driver is common, any dc drift shows up only as a brightness shift. Luminance channel frequency response uniformity is ensured by the common driver.

RGB drive, wherein RGB signals are applied to respective cathodes and G_1 is dc biased, requires less drive and has none of the potential color fidelity errors of the color-difference system. RGB drive, however, places higher demands on the drive amplifiers for linearity, frequency response, and dc stability, plus requiring a matrixing network in the amplifier chain.

Low-level RGB matrixing and CRT drive are commonly used, especially with CRT devices that have unitized guns in which the common G_1 and G_2 elements require differential cathode bias adjustments and drive adjustments to yield gray-scale tracking. In this technique, RGB signals are matrixed at a level of a few volts and then amplified to a higher level (100 to 200 V) suitable for CRT cathode drive.

Direct current stability, frequency response, and linearity of the three stages, even if somewhat less than ideal, should be reasonably well matched to ensure overall gray-scale tracking. Bias and gain adjustments should be independent in their operation, rather than interdependent, and should minimally affect those characteristics listed previously in this section.

Figure 17.1.32 illustrates a simple example of one of the three CRT drivers. If the amplifier black-level bias voltage equals the black level from the RGB decoder, drive adjustment will not change the amplifier black-level output voltage level or affect CRT cutoff. Furthermore, if R_B >>

R_E, drive level will be independent of bias setting. Note also that frequency response-determining networks are configured to be unaffected by adjustments.

Frequently, the shunt peaking coil can be made common to all three channels, because differences between the channels are predominantly color-difference signals of relatively narrow bandwidth. Although the frequency responses could be compensated to provide the widest possible bandwidth, this is usually not necessary when the frequency response of preceding low-level luminance processing (especially the peaking stage) is factored in. One exception in which output stage bandwidth must be increased to its maximum is in an application, television receiver or video monitor, where direct RGB inputs are provided for auxiliary services, such as computers, teletext, and S-VHS wideband VCRs.

Comb Filter Processing

The frequency spectrum of a typical NTSC composite video signal is shown in Figure 17.1.33*a*. A comb filter, characterized by 100 percent transmission of desired frequencies of a given channel and substantially zero transmission for the undesired interleaved signal spectrum, can effectively separate chroma and luminance components from the composite signal. Such a filter can be easily made, in principle, by delaying the composite video signal one horizontal scan period (63.555 μs in NTSC-M) and adding or subtracting to the undelayed composite video signal (Figure 17.1.33*b* and *c*).

The output of the sum channel will have frequencies at f (horizontal), and all integral multiples thereof reinforce in phase, while those interleaved frequencies will be out of phase and will cancel. This can be used as the luminance path. The difference channel will have integral frequency multiples cancel while the interleaved ones will reinforce. This channel can serve as the chrominance channel. The filter characteristic and interleaving are shown in Figure 17.1.33*c*.

Automatic Circuits

The relative level of the chroma subcarrier in the incoming signal is highly sensitive to transmission path disorders, thereby introducing objectionable variations in saturation. These can be observed between one received signal and another or over a period of time on the same channel unless some adaptive correction is built into the system. The color burst reference, transmitted at 40 IRE units peak-to-peak, is representative of the same path distortions and is normally used as a reference for automatic gain controlling the chroma channel. A balanced peak detector or synchronous detector, having good noise rejection characteristics, detects the burst level and provides the control signal to the chroma gain-controlled stage.

Allowing the receiver chroma channel to operate during reception of a monochrome signal will result in unnecessary cross color and colored noise, made worse by the ACC increasing the chroma amplifier gain to the maximum. Most receivers, therefore, cut off the chroma channel transmission when the received burst level goes below approximately 5 to 7 percent. Hysteresis has been used to minimize the flutter or threshold problem with varying signal levels.

Burst-referenced ACC systems perform adequately when receiving correctly modulated signals with the appropriate burst-to-chroma ratio. Occasionally, however, burst level may not bear a correct relation to its accompanying chroma signal, leading to incorrectly saturated color levels. It has been determined that most viewers are more critical of excessive color levels than insufficient ones. Experience has shown that when peak chroma amplitude exceeds burst by greater than 2/1, limiting of the chroma signal is helpful. This threshold nearly corresponds to the ampli-

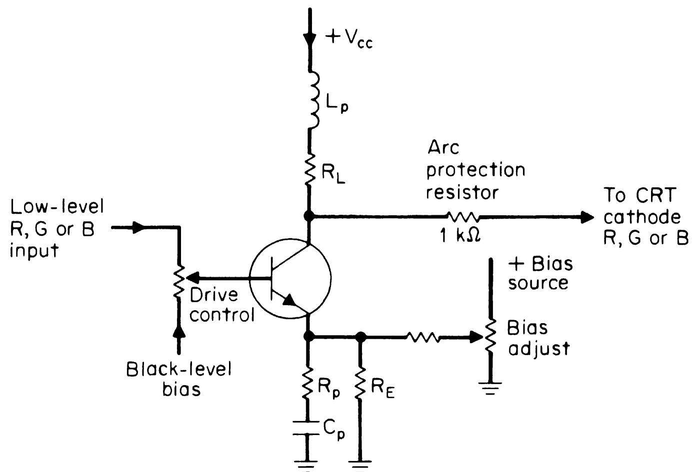


Figure 17.1.32 Simplified R, G, and B output stage.

tude of a 75 percent modulated color bar chart (2.2/1). At this level, negligible distortion is introduced into correctly modulated signals. Only those that are nonstandard are affected significantly. The output of the peak chroma detector also is sent to the chroma gain-controlled stage.

Tint

One major objective in color television receiver design is to minimize the incidence of flesh-tone reproduction with incorrect hue. Automatic hue-correcting systems can be categorized into two classes:

- Static flesh-tone correction, achieved by selecting the chroma demodulating angles and gain ratios to desensitize the resultant color-difference vector in the flesh-tone region (+*I* axis). The demodulation parameters remain fixed, but the effective *Q* axis gain is reduced. This has the disadvantage of distorting hues in all four quadrants.
- Dynamic flesh-tone corrective systems, which can adaptively confine correction to the region within several degrees of the positive *I* axis, leaving all other hues relatively unaffected. This is typically accomplished by detecting the phase of the incoming chroma signal and modulating the phase angle of the demodulator reference signal to result in an effective phase shift of 10 to 15° toward the *I* axis for a chroma vector that lies within 30° of the I axis. This approach produces no amplitude change in the chroma. In fact, for chroma saturation greater than 70 percent, the system is defeated on the theory that the color is not a flesh tone.

A simplification in circuitry can be achieved if the effective correction area is increased to the entire positive-*I* 180° sector. A conventional phase detector can be utilized and the maximum correction of approximately 20° will occur for chroma signals having phase of ±45° from the *I*

(*a*)

(*b*)

(*c*)

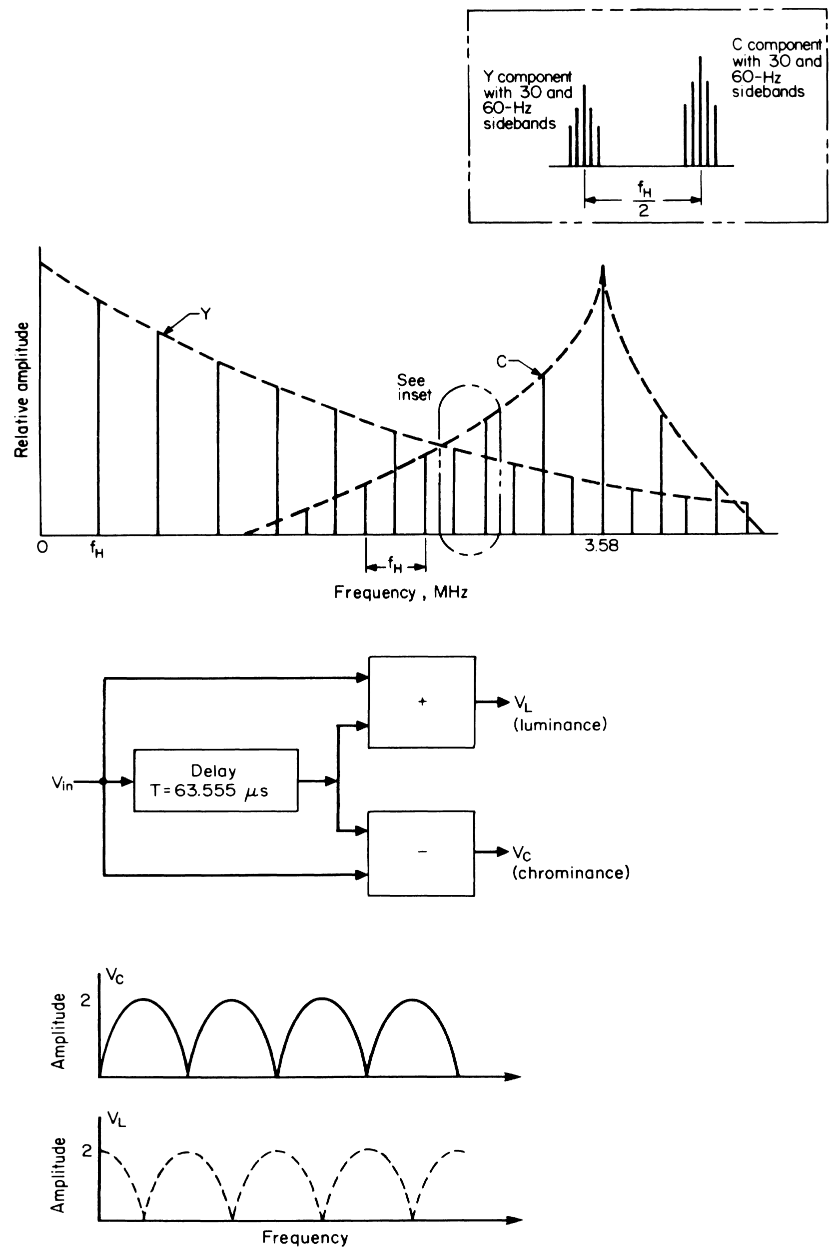


Figure 17.1.33 Color system filtering: (*a*) frequency spectrum of NTSC color system, showing interleaving of signals, (*b*) simplified 1H delay comb filter block diagram, (*c*) chrominance V_C and luminance V_L outputs of the comb filter.

axis. Signals with phase greater or less than 45° will have increasingly lower correction values, as illustrated in Figure 17.1.34.

The *vertical interval reference* (VIR) signal, as shown in Figure 17.1.34, provides references for black level, luminance, and—in addition to burst—a reference for chroma amplitude and phase. While originally developed to aid broadcasters, it has been employed in television receivers to correct for saturation and hue errors resulting from transmitter or path distortion errors.

17.1.2l Scanning Synchronization

The scan-synchronizing signal, consisting of the horizontal and vertical sync pulses, is removed from the composite video signal by means of a sync-separator circuit. The classic approach has been to ac-couple the video signal to an overdriven transistor amplifier (illustrated in Figure 17.1.35) that is biased off for signal levels smaller than V_{CO} and saturates for signal levels greater than V_{sat}, a range of approximately 0.1 V.

Vertical synchronizing information can be recovered from the output of the sync separator by the technique of integration. The classic two-section RC integrator provides a smooth, ramp waveform that corresponds to the vertical sync block as shown in Figure 17.1.36. The ramp is then sent to a *relaxation* or *blocking* oscillator, which is operating with a period slightly longer than the vertical frame period. The upper part of the ramp will trigger the oscillator into conduction to achieve vertical synchronization. The oscillator will then reset and wait for the threshold level of the next ramp. The vertical-hold potentiometer controls the free-running frequency or period of the vertical oscillator.

One modification to the integrator design is to reduce the integration (speed up the time constants) and provide for some differentiation of the waveform prior to applying it to the vertical oscillator (shown in Figure 17.1.37). Although degrading the noise performance of the system, this technique provides a more certain and repeatable trigger level than the full integrator, thereby leading to an improvement in interlace over a larger portion of the hold-in range. A second benefit is to provide a more stable vertical lock when receiving signals that have distorted vertical sync waveforms. These can be generated by video cassette recorders in nonstandard playback modes, such as fast/slow forward, still, and reverse. Prerecorded tapes having antipiracy nonstandard sync waveforms also contribute to the problem.

Vertical Countdown Synchronizing Systems

A vertical scan system using digital logic elements can be based upon the frequency relationship of 525/2 that exists between the horizontal and vertical scans. Such a system can be considered to be phase-locked for both horizontal scan and vertical scan, which will result in improved noise immunity and picture stability. Vertical sync is derived from a pulse train having a frequency of twice the horizontal rate. The sync is therefore precisely timed for both even and odd fields, resulting in excellent interlace. This system requires no hold control.

The block diagram of one design is shown in Figure 17.1.38. The 31.5 kHz clock input is converted to horizontal drive pulses by a divide-by-2 flip-flop. A two-mode counter, set for 525 counts for standard interlaced signals and 541 for noninterlaced signals, produces the vertical output pulse. Noninterlaced signals are produced by a variety of VCR cameras, games, and picture test generators. The choice of 541 allows the counter to continue until the arrival of vertical sync from the composite video waveform. The actual vertical sync then resets the counter.

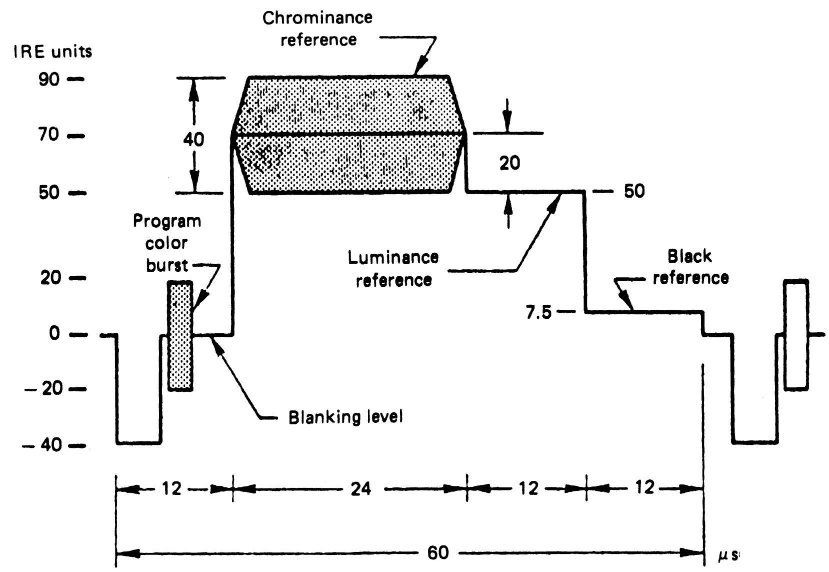


Figure 17.1.34 Vertical interval reference (VIR) signal. Note that the chrominance and the program color burst are in phase.

Other systems make use of clock frequencies of 16 and 32 times the horizontal frequency. Low-cost crystal or ceramic resonator-controlled oscillators can be built to operate at these frequencies. As in the first system, dual-mode operation is necessary in order to handle standard interlaced and nonstandard sync waveforms. An exact 525/2 countdown is used with interlaced signals. For noninterlaced signals, the systems usually operate in a free-running mode with injection of the video-derived sync pulse causing lock.

A critical characteristic, and probably the most complex portion of any countdown system, is the circuitry that properly adjusts for nonstandard sync waveforms. These can be simple 525 noninterlaced fields, distorted vertical sync blocks, blocks having no horizontal serrations, fields with excessive or insufficient lines, and a combination of the above.

Impulse Noise Suppression

The simplest type suppression that will improve the basic circuit in a noise environment of human origin consists of a parallel resistor and capacitor in series with the sync charging capacitor. Variations of this simple circuit, which include diode clamps and switches, have been developed to speed up the noise suppression performance while not permitting excessive tilt in the sync output during the vertical block.

More complex solutions to the noise problem include the noise canceler and noise gate. The canceler monitors the video signal between the second detector and the sync separator. A noise spike, which exceeds the sync or pedestal level, is inverted and added to the video after the isola-

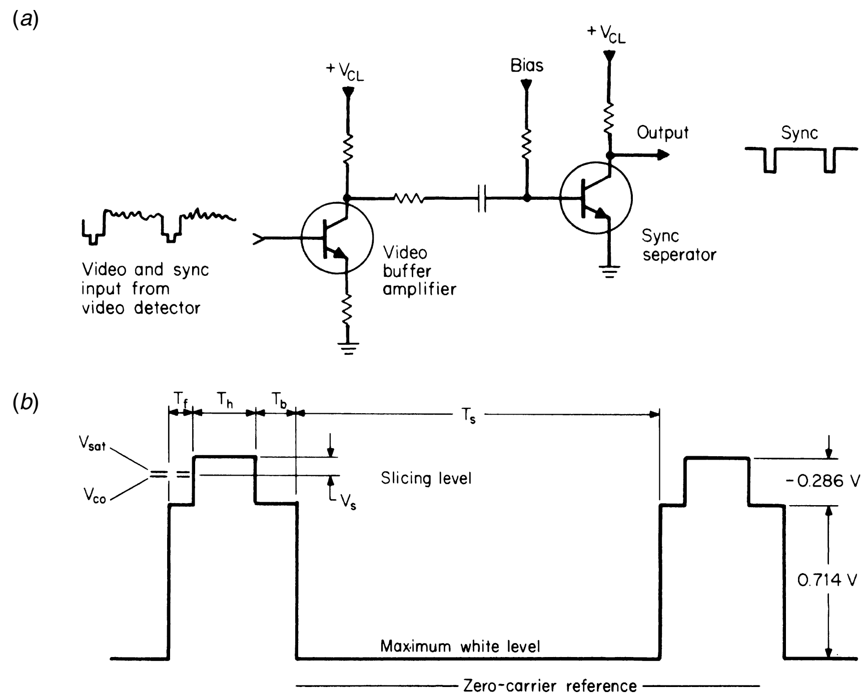


Figure 17.1.35 Sync separator: (*a*) typical circuit, (*b*) sync waveform.

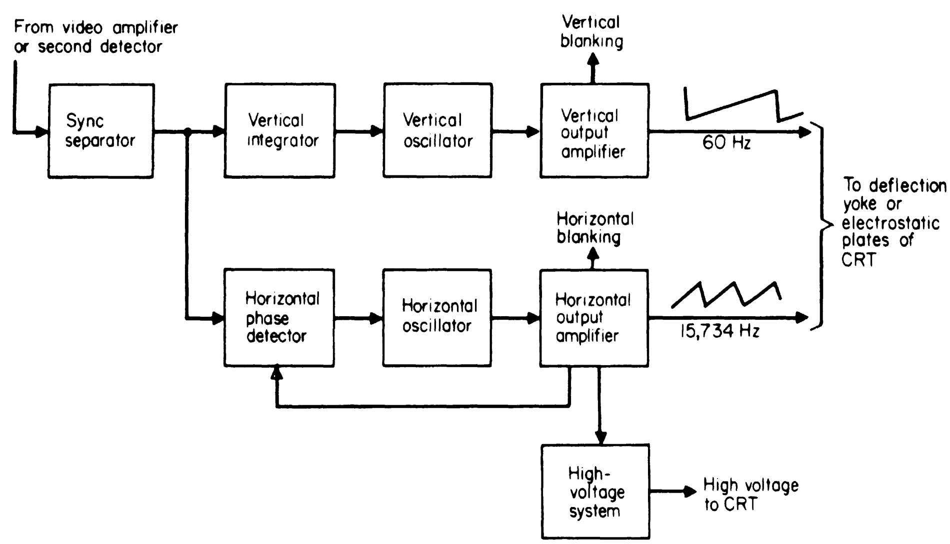


Figure 17.1.36 Picture scanning section functional block diagram.

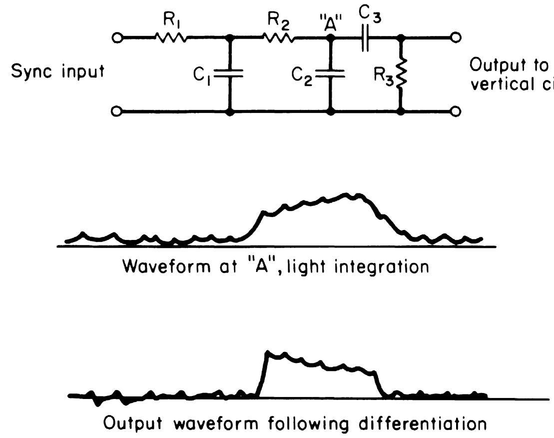


Figure 17.1.37 Modified integrator followed by a differentiator.

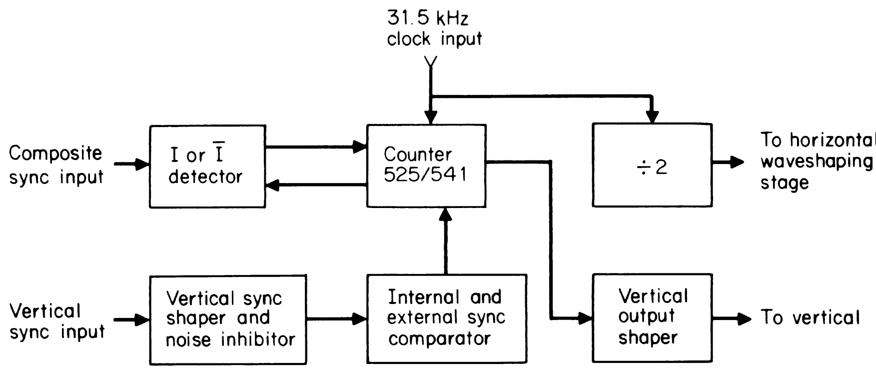


Figure 17.1.38 Vertical countdown sync processor, using a $2f_H$ clock.

tion resistor. This action cancels that part of the noise pulse which would otherwise produce an output from the sync separator.

For proper operation, the canceler circuit must track the sync-tip amplitude from a strong RF signal to the fringe level. The noise gate, in a similar manner, recognizes a noise pulse of large amplitude and prevents the sync separator from conducting, either by applying reverse bias or by opening a transistor switch in the emitter circuit.

Horizontal Automatic Phase Control

Horizontal scan synchronization is accomplished by means of an APC loop, with theory and characteristics similar to those used in the chroma reference system. Input signals to the horizontal APC loop are sync pulses from the sync separator and horizontal flyback pulses, which are integrated to form a sawtooth waveform. The phase detector compares the phase (time coinci-

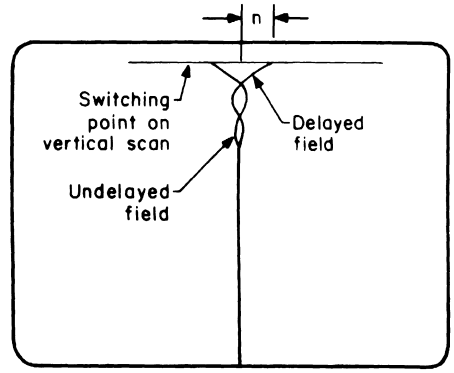


Figure 17.1.39 The pattern on a receiver screen displaying a vertical line with the even and odd fields offset by *n* ms.

dence) of these two waveforms and sends a dc-coupled low-pass filtered error signal to the voltage-controlled oscillator to cause the frequency to shift in the direction of minimal phase error between the two input signals.

The recovery of the APC loop to a step transient input involves the parameters of the natural resonant frequency, as well as the amount of overshoot permitted after the correct phase has been reached. Both of these characteristics can be evaluated by use of a jitter generator, which creates a time base error between alternate fields. The resultant sync error and system dynamics can be seen in the picture display of the receiver shown in Figure 17.1.39. This type of disturbance occurs when the two or three playback heads of a consumer helical-scan VCR switch tracks. It will also occur when the receiver is operated from a signal that does not have horizontal slices in the vertical sync block.

Phase detector circuitry in the form of classic double-diode bridge detectors, in either the balanced or unbalanced configuration, can sometimes be found in low-cost receivers. Integrated circuits commonly utilize a gated differential amplifier (Figure 17.1.40.) in which the common feed from the current source is driven by the sync pulse, and the sawtooth waveform derived from the flyback pulse is applied to one side of the differential pair.

Vertical Scanning

Class-B vertical circuits consist essentially of an audio frequency amplifier with current feedback. This approach maintains linearity without the need for an adjustable linearity control. Yoke impedance changes caused by temperature variations will not affect the yoke current, thus, a thermistor is not required. The current-sensing resistor must, of course, be temperature stable. An amplifier that uses a single NPN and a single PNP transistor in the form of a complementary output stage is given in Figure 17.1.41. Quasi-complementary, Darlington outputs and other common audio output stage configurations also can be used.

Establishing proper dc bias for the output stages is quite critical. Too little quiescent current will result in crossover distortion that will impose a faint horizontal white line in the center of the picture even though the distortion may not be detectable on an oscilloscope presentation of the yoke current waveform. Too much quiescent current results in excessive power dissipation in the output transistors.

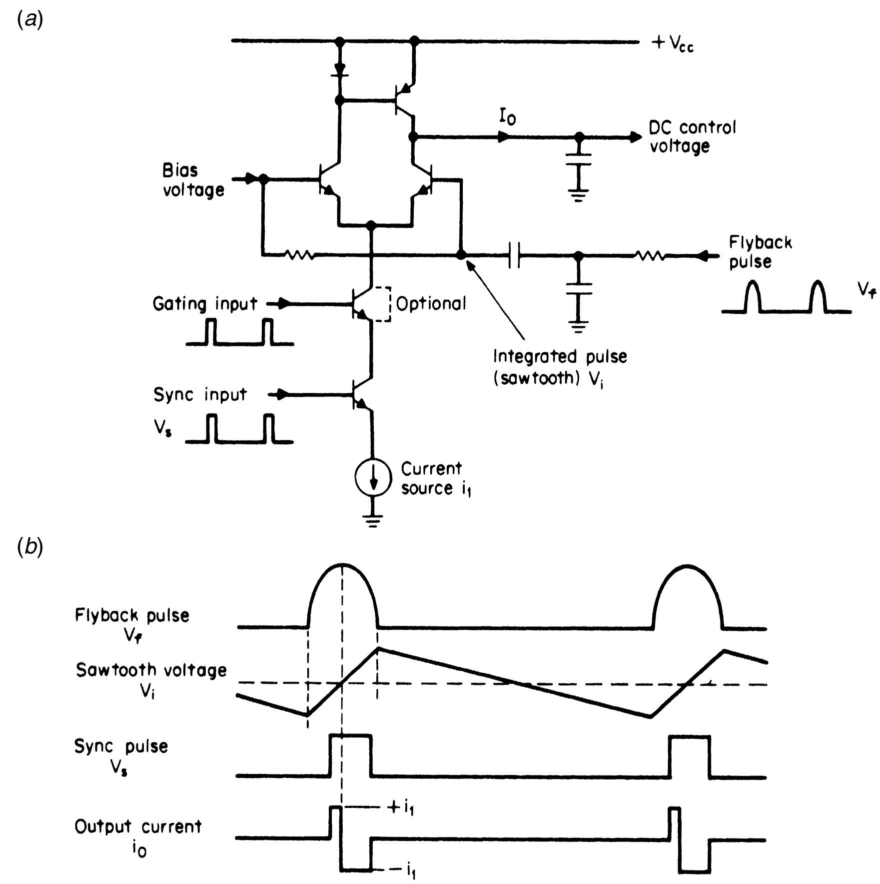


Figure 17.1.40 Gated phase detector in an IC implementation: (*a*) circuit diagram, (*b*) pulse-timing waveforms.

Retrace Power Reduction

The voltage required to accomplish retrace results in a substantial portion of the power dissipation in the output devices. The supply voltage to the amplifier and corresponding power dissipation can be reduced by using a *retrace switch* or *flyback generator* circuit to provide additional supply voltage during retrace. One version of a retrace switch is given in Figure 17.1.42. During the trace time, the capacitor is charged to a voltage near the supply voltage. As retrace begins, the voltage across the yoke goes positive, thus forcing Q_R into saturation. This places the cathode of C_R at the supply potential and the anode at a level of 1.5 to 2 times the supply, depending upon the values of R_R and C_R.

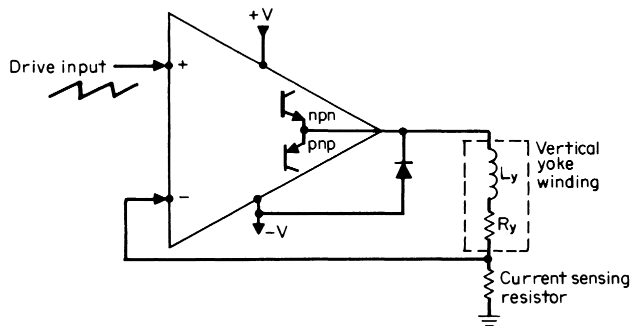


Figure 17.1.41 A class-B vertical output stage.

Horizontal Scanning

The horizontal scan system has two primary functions:

- It provides a modified sawtooth-shaped current to the horizontal yoke coils to cause the electron beam to travel horizontally across the face of the CRT.
- It provides drive to the high-voltage or flyback transformer to create the voltage needed for the CRT anode.

Frequently, low-voltage supplies also are derived from the flyback transformer. The major components of the horizontal-scan section consist of a driver stage, horizontal output device (either bipolar transistor or SCR), yoke current damper diode, retrace capacitor, yoke coil, and flyback transformer, as illustrated in Figure 17.1.43.

During the scan or retrace interval, the deflection yoke may be considered a pure inductance with a dc voltage impressed across it. This creates a sawtooth waveform of current (see Figure 17.1.44). This current flows through the damper diode during the first half scan. It then reverses direction and flows through the horizontal output transistor collector. This sawtooth-current waveform deflects the electron beam across the face of the picture tube. A similarly shaped current flows through the primary winding of the high-voltage output transformer.

At the beginning of the retrace interval, the transformer and yoke inductances transfer energy to the retrace-tuning capacitor and the accompanying stray capacitances, thereby causing a half sine wave of voltage to be generated. This high-energy pulse appears on the transistor collector and is stepped up, via the flyback transformer, to become the high voltage for the picture tube anode. Finally, at the end of the cycle, the damper diode conducts, and another horizontal scan is started.

17.1.2m High Voltage Generation

High voltage in the range 8 to 16 kV is required to supply the anode of monochrome picture tubes. Color tubes have anode requirements in the range 20 to 30 kV and focus voltage requirements of 3 to 12 kV. The horizontal flyback transformer is the common element in the generation of these high voltages. There are three common variations of this design:

- The flyback with a half-wave rectifier
- Flyback driving a voltage multiplier

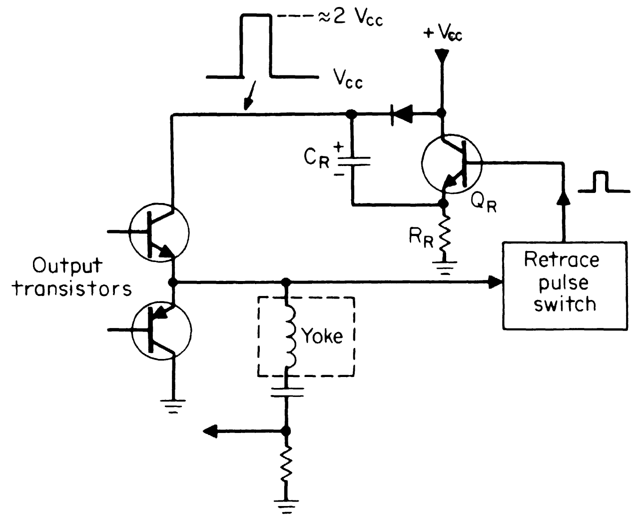


Figure 17.1.42 Simplified schematic of a retrace switch.

- Voltage-adding integrated flyback

Consider the simplified horizontal-scan circuit shown in Figure 17.1.43. The voltage at the top of the high voltage (HV) winding consists of a series of pulses delivered during retrace from the stored energy in the yoke field. The yoke voltage pulses are then multiplied by the turns ratio of the HV winding to primary winding. The peak voltage across the primary during retrace is given by:

$$V_{p(pk)} = E_{in} 0.8\left(1.79 + 1.57\frac{T_{trace}}{T_{retrace}}\right) \tag{17.1.2}$$

Where:
E_{in} = supply voltage (B+)
0.8 = accounts for the pulse shape factor with third harmonic tuning
T_{trace} = trace period, ≈ 52.0 µs
T_{retrace} = retrace period, ≈ 11.5 µs

Flyback with Half-Wave Rectifier

The most common HV supply for small-screen monochrome and color television receivers uses a direct half-wave rectifier circuit. The pulses at the top of the HV winding are rectified by the single diode, or composite of several diodes in series. The charge is then stored in the capacitance of the anode region of the picture tube. Considerable voltage step-up is required from the primary to the HV winding. This results in a large value of leakage inductance for the HV winding, which decreases its efficiency as a step-up transformer.

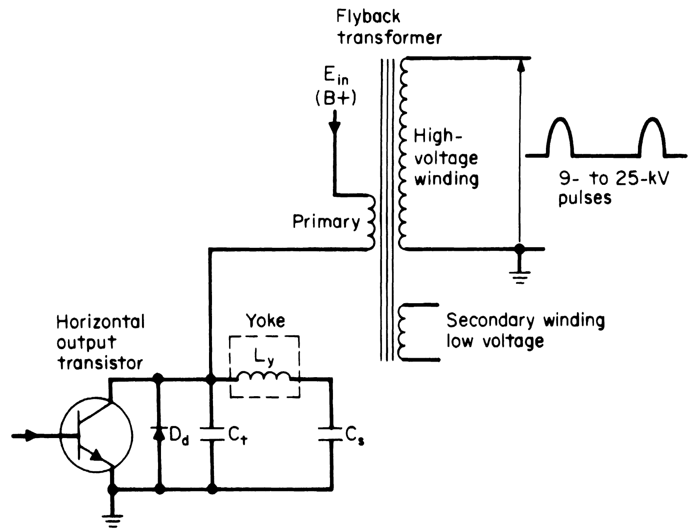


Figure 17.1.43 Simplified horizontal scan circuit.

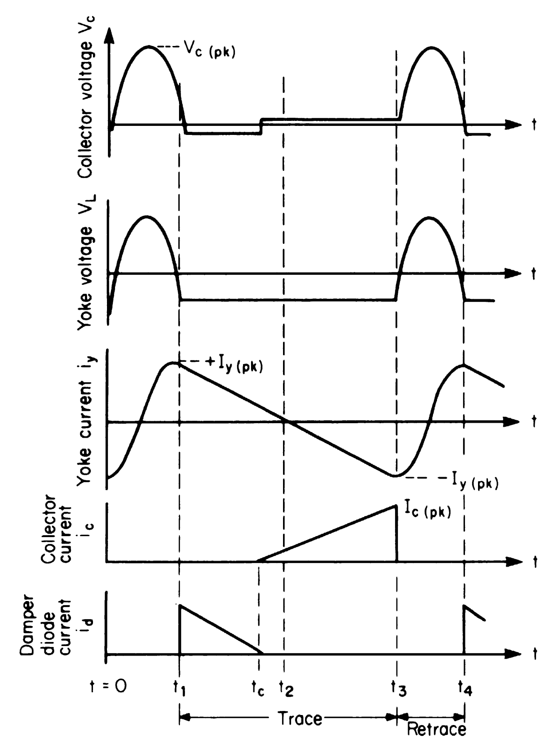


Figure 17.1.44 Horizontal scan waveform timing diagrams.

Harmonic tuning of the HV winding improves the efficiency by making the total inductance and the distributed capacitance of the winding plus the CRT anode capacitance resonate at an odd harmonic of the flyback pulse frequency. For the single-rectifier circuit, usually the third harmonic resonance is most easily implemented by proper choice of winding configuration, which results in appropriate leakage inductance and distributed capacitance values. This will result in HV pulse waveforms, which will give an improvement in the HV supply regulation (internal impedance) as well as a reduction in the amplitude of ringing in the current and voltage waveforms at the start of scan.

Voltage Multiplier Circuits

A voltage multiplier circuit consists of a combination of diodes and capacitors connected in such a way that the dc output voltage is greater than the peak amplitude of the input pulse. The tripler was a common HV system for color television receivers requiring anode voltages of 25 to 30 kV dating back to the 1970s. The considerably reduced pulse voltage required from the HV winding resulted in a flyback transformer with a more tightly coupled HV winding.

Integrated Flybacks

Most medium- and large-screen color receivers utilize an integrated flyback transformer in which the HV winding is segmented into three or four parallel-wound sections. These sections are series-connected with a diode between adjacent segments. These diodes are physically mounted as part of the HV section. The transformer is then encapsulated in high-voltage polyester or epoxy.

Two HV winding construction configurations also have been used. One, the layer or solenoid-wound type, has very tight coupling to the primary and operates well with no deliberate harmonic tuning. Each winding (layer) must be designed to have balanced voltage and capacitance with respect to the primary. The second, a bobbin or segmented-winding design, has high leakage inductance and usually requires tuning to an odd harmonic (e.g., the ninth). Regulation of this construction is not quite as good as the solenoid-wound primary winding at a horizontal-frequency rate. The +12 V, +24 V, +25 V, and –27 V supplies are scan-rectified. The +185 V, overvoltage sensing, focus voltage, and 25 kV anode voltage are derived by retrace-rectified supplies. The CRT filament is used directly in its ac mode.

Flyback-generated supplies provide a convenient means for isolation between different ground systems, as required for an iso-hot chassis.

Power Supplies

Most receivers use the flyback pulse from the horizontal transformer as the source of power for the various dc voltages required by the set. Using the pulse waveform at a duty cycle of 10 or 15 percent, by proper winding direction and grounding of either end of the winding, several different voltage sources can be created.

Scan rectified supplies are operated at a duty cycle of approximately 80 percent and are thus better able to furnish higher current loads. Also, the diodes used in such supplies must be capable of blocking voltages that are nine to ten times larger than the level they are producing. Diodes having fast-recovery characteristics are used to keep the power dissipation at a minimum during the turn-off interval because of the presence of this high reverse voltage.

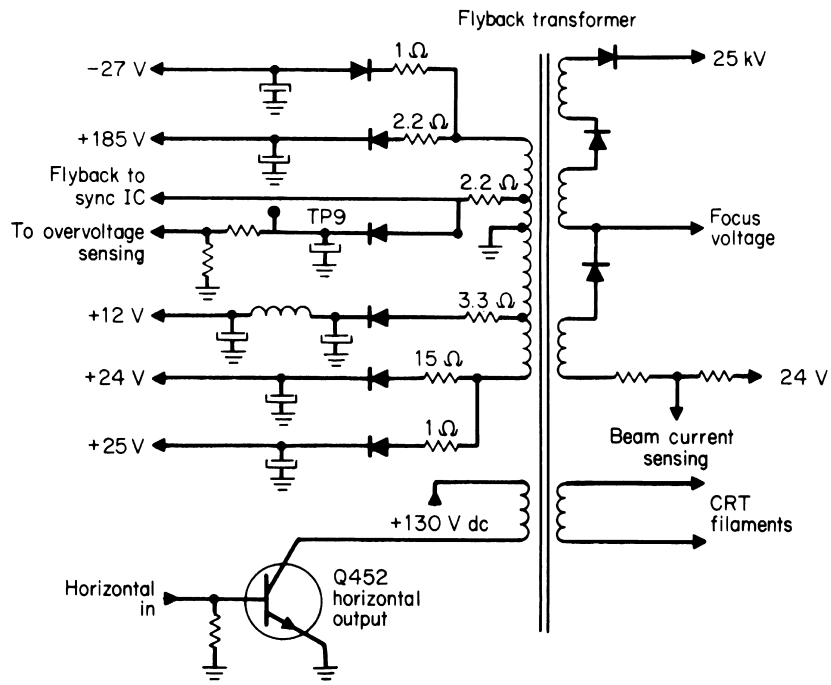


Figure 17.1.45 Auxiliary power sources derived from the horizontal output transformer. (*Courtesy of Philips.*)

A typical receiver system containing the various auxiliary power supplies derived from flyback transformer windings is shown in Figure 17.1.45. Transistor *Q* 452 switches the primary winding at a horizontal frequency rate. The +12 V, +24 V, +25 V, and –27 V supplies are scan-rectified. The +185 V, overvoltage sensing, focus voltage, and 25 kV anode voltage are derived by retrace-rectified supplies. The CRT filament is used directly in its ac mode.

As noted in the previous section, flyback-generated supplies provide a convenient means for isolation between different ground systems. Figure 17.1.46 shows the block diagram of such a television receiver power supply system [6, 7].

17.1.3 Receiver Standards and Specifications

The following sections list standards and recommendations relating to various areas of receiver design and operation. For more information on EIA/CEA documents, see the Consumer Electronics Association Web site at http://www.ce.org. For additional information on documents from the Society of Motion Picture and Television Engineers, see the SMPTE Web site at http://www.smpte.org.

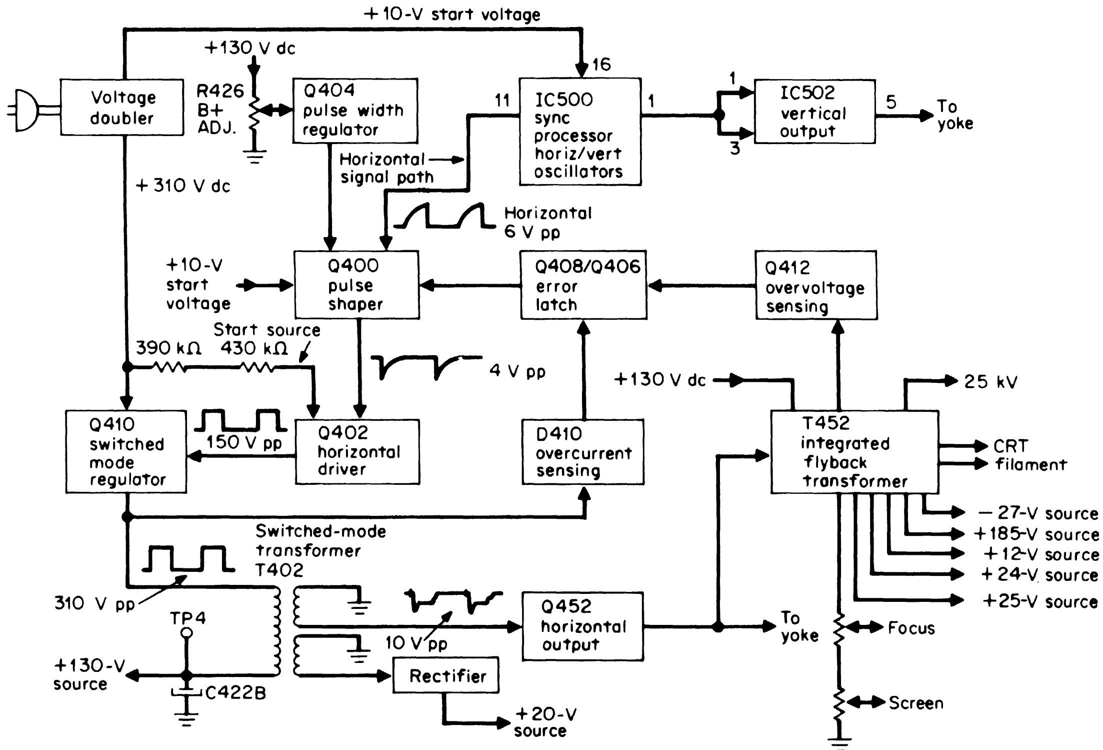


Figure 17.1.46 Color receiver power supply using a switched-mode regulator system. (*Courtesy of Philips.*)

CEB5: Recommended Practice for DTV Receiver Monitor Mode Capability

CEB5 is intended to provide recommendations to digital television (DTV) designers/manufacturers concerning a "monitor" mode capability. See EIA-762 for minimum specifications for a DTV remodulator.

CEB12: PSIP Recommended Practice

This document provides guidelines to receiver manufacturers regarding implementation of the Program and System Information Protocol used in the ATSC DTV Standard.

EIA/CEA-863: Connection Color Codes for Home Theater Systems

This standard defines the colors for marking connections commonly used for electronic devices in a home theater system. This standard adds continuity to installation information and assures consistency of information to installers.

EIA/CEA-849: Application Profiles for EIA-775A Compliant DTVs

This standard specifies profiles for various applications of the EIA-775A standard. The application areas covered here include digital streams compliant with ATSC terrestrial broadcast, direct-broadcast satellite (DBS), OpenCable™, and standard definition Digital Video (DV) camcorders.

EIA/544-A: Immunity of TV and VCR Tuners to Internally Generated Harmonic Interference from Signals in the Band 535 kHz to 30 MHz

This Standard establishes performance guidelines for rejection of interference by television receivers, video cassette recorders, and tuners. It details a measurement procedure which determines the level of interfering signal which will generate harmonics in the tuner, causing interference 40 dB below a desired signal at the intermediate frequency (IF) output of a tuner under test. This standard covers interference immunity to CB, amateur radio, and other transmissions.

EIA-761-A: DTV Remodulator Specification with Enhanced OSD Capability

This standard defines minimum specifications for a one-way data path utilizing an 8 VSB trellis or a 16 VSB remodulator in compliance with ATSC Standard A/53, Annex D. This standard also defines on-screen display (OSD) capabilities. This standard applies to any type of device used to connect to an ATSC compliant digital television receiver (DTV) receiver. Devices meeting this standard should interoperate with any ATSC compliant receiver that also supports "monitor mode." This standard addresses required RF output specifications, on-screen display (OSD) capabilities, and capability profiles for a DTV remodulator and recommendations concerning input to the remodulator. This standard does not address 8-VSB without OSD. For information concerning 8 VSB without OSD, see also EIA-762 and EIA-799.

EIA-762: DTV Remodulator Specification

This standard defines a minimum specification for a one-way data path utilizing an 8-VSB trellis remodulator in compliance with ATSC A/53, Annex D. This standard applies to any type of device used to connect to an ATSC compliant digital television receiver (DTV) receiver. Devices meeting this standard should interoperate with any ATSC compliant receiver that also supports "monitor mode" (see EIA CEB-5.) This standard addresses both required RF output specifications for a DTV remodulator and recommendations concerning input to the remodulator.

EIA-770.1-C: Analog 525 Line Component Video Interface-Three Channels

This standard defines the physical characteristics of an interface and the parameters of the signals carried across the interface, using three parallel channels for the interconnection of equipment operating with analog component video signals. This standard includes specifications for two scanning structures: 1H having 525 lines, 59.94 fields/second, 2:1 interlaced, and a horizontal scanning rate of 15.734 kHz; and 2H for doubled scanned interfaces having 525 lines, 59.94 frames/second, progressively scanned, and having a horizontal scanning rate of 31.47 kHz. Both interfaces shall be capable of either 4:3 or 16:9 aspect ratios.

EIA-770.2-A: Standard Definition TV Analog Component Video Interface

This standard defines the physical characteristics of an interface and the parameters of the signals carried across that interface, using three parallel channels for the interconnection of equipment operating with analog component video signals. The standard includes specifications for: 1) 480i video format defined by 480 active lines, 525 total lines, 2:1 interlaced at 59.94 or 60 fields/second; and 2) 480p video format defined by 480 active lines, 525 total lines, progressively scanned at 59.94 or 60 frames/second. Both video formats shall be capable of either 4:3 or 16:9 aspect ratios.

EIA-770.3-A: High Definition TV Analog Component Video Interface

This standard defines two raster-scanning systems for the representation of stationary or moving two-dimensional images sampled temporally at a constant frame rate. The first image format specified is 1920 × 1080 samples (pixels) inside a total raster of 1125 lines. The second image format specified is 1280 × 720 samples (pixels) inside a total raster of 750 lines. Both image formats shall have an aspect ratio of 16:9.

EIA-775-A: DTV 1394 Interface Specification

This standard defines a specification for a baseband digital interface to a DTV using the IEEE-1394 bus and provides a level of functionality that is similar to the analog system. It is designed to enable interoperability between a DTV compliant with this standard and various types of consumer digital audio/video sources including digital set-top boxes (STBs) and analog/digital hard disk or videocassette recorders (VCRs).

EIA-775-A: Errata Sheet

Additional information on the DTV 1394 interface is given in this errata sheet.

EIA-775.2: Service Selection Information for Digital Storage Media Interoperability

A digital storage device such as a D-VHS or hard disk digital recorder may be used by the DTV or by another source device such as a cable set-top box to record or time-shift digital television signals. This standard supports the use of such storage devices by defining *service selection information* (SSI), methods for managing discontinuities that occur during recording and playback, and rules for management of partial transport streams.

EIA-799: On-Screen Display Specification

This standard specifies syntax semantics for bitmapped graphics data typically used for on-screen display (OSD). The standard is applicable whenever it is necessary to specify a standard method for delivery of bitmapped graphics data. The pixel formats include optional alpha-blend and transparency attributes to support composition of graphics over analog or digitally decoded video within the display.

17.1.3a Receive Antennas

CEB6-C: TV Receiving Antenna Manufacturers Guide to Categorizing Antennas for Use With the CEA TV Antenna Sector Map Program, Antenna Types and Characteristics, Minimum Performance Requirements, Packaging, and Marking Specifications

This bulletin was prepared to provide manufacturers of television receiver antennas with appropriate guidance on determining antenna categories and minimum performance requirements to comply with the CEA TV Antenna Selector Map program. Essential elements of this program include color coding of various television reception environments in a market, with corresponding color-coded marking of antennas to provide reception. See EIA-774 for test and measurement procedures. The CEA copyrighted logo for marking antennas and logo use guidelines are available from CEA.

CEB7: TV Receiving Antenna Manufacturers Guide to Indoor Antennas for Use with the CEA Indoor TV Antenna Certification Program, Indoor Antenna Characteristics, Packaging and Marking Specifications, and Minimum Performance Requirements

This bulletin was prepared by the CEA R-5 Antennas Committee to provide manufacturers of indoor television receiver antennas with appropriate guidance on determining antenna characteristics and minimum performance requirements to comply with the CEA Indoor TV Antenna Certification program. Essential elements of this program include indoor TV antenna technical specifications and the use of a CEA copyrighted certification logo. Test and measurement procedures are contained in the EIA-774 standard, and the CEA copyrighted logo for marking antennas and logo use guidelines are available from CEA.

EIA-774: TV Receiving Antenna Performance Presentation and Measurement

This standard is intended to provide television receive antenna manufacturers with appropriate test and measurement procedures to examine antenna performance parameters necessary to comply with elements of the CEMA TV Antenna Selector Map program, specifically EIA CEB-6. Essential elements include procedures to determine antenna gain, front-to-back ratio, directivity, and distortion performance of active antennas with integrated amplifiers.

17.1.3b Closed Captioning

EIA Documents

EIA/CEA-CEB8: Consideration of EIA-608-B Data Within the DTV Closed Captioning (EIA-708-B) Construct

EIA/CEA CEB-8 provides guidance on the use and processing of the EIA/CEA-608-B data stream embedded within the ATSC MPEG-2 video elementary transport stream. EIA/CEA CEB-8 augments EIA-708-B (addressing DTV Closed Captioning) Sections 4.3 and 9.23.

EIA/CEA-CEB10: EIA-708-B Implementation Guidance

This document provides guidelines for receiver implementation of the DTV closed captioning specification given in EIA-708-B.

EIA/CEA-608 revision B: Line 21 Data Services

Serves as a technical guide for those providing encoding equipment and/or decoding equipment to produce material with encoded data embedded in Line 21 of the vertical blanking interval of the NTSC video signal. It is also a usage guide for those who will produce material using such equipment. Revision incorporates content advisory.

EIA-708-B: Digital Television (DTV) Closed Captioning

This document is intended as a definition of DTV Closed Captioning (DTVCC) and provides specifications and/or guidelines for caption service providers, DTVCC decoder and encoder manufacturers, DTV receiver manufacturers, and DTV signal processing equipment manufacturers. This specification includes: 1) a description of the transport method of DTVCC data in the DTV signal, 2) a description of DTVCC specific data packets and structures, 3) a specification of how DTVCC information is to be processed, 4) a list of minimum implementation recommendations for DTVCC receiver manufacturers, and 5) a set of recommended practices for DTV encoder and decoder manufacturers

SMPTE Documents

SMPTE 291M-1998: Ancillary Data Packet and Space Formatting

This standard specifies the basic formatting structure of the ancillary data space in the digital video data steam in the form of 10-bit words. Application of this standard includes 525-line, 625-line, component or composite, and high-definition digital television interfaces which provide 8- or 10-bit data ancillary data space. Space available for ancillary data packets is defined in the document specifying the connecting interface.

SMPTE 292M-1998: Bit-Serial Digital Interface for High-Definition Television Systems

This standard defines a bit-serial digital coaxial and fiber-optic interface for HDTV component signals operating at data rates in the range of 1.3 Gbits/s to 1.5 Gbits/s. Bit-parallel data derived from a specified source format are multiplexed and serialized to form the serial data stream. A common data format and channel coding are used based on modifications, if necessary, to the source format parallel data for a given high-definition television system. Coaxial cable interfaces are suitable for application where the signal loss does not exceed an amount specified by the receiver manufacturer. Typical loss amounts would be in the range of up to 20 dB at one-half the clock frequency. Fiber optic interfaces are suitable for application at up to 2 km of distance using single-mode fiber.

SMPTE 333M-1999: DTV Closed-Caption Server to Encoder Interface

This standard defines a standard for interoperation of digital television closed-caption (DTVCC) data server devices and video encoders. The caption data server devices provide partially-formatted EIA 708 data to the video encoders using the request/response protocol and interface defined in this standard. The video encoder completes the formatting and includes the EIA 708 data in the video elementary stream picture-level *user data* field. This standard describes an interface for transmission of DTVCC data from a caption server to video encoder.

SMPTE 334M-2000: Vertical Ancillary Data Mapping for Bit Serial Interface

This standard defines a method of coding which allows data services to be carried in the vertical ancillary data space of a bit-serial component television signal conforming with SMPTE 292M or ANSI/SMPTE 259M

17.1.3c Receiver Functionality

EIA/CEA-775.1: Web Enhanced DTV 1394 Interface Specification

This standard includes mechanisms to allow a source of MPEG service to utilize the MPEG decoding and display capabilities in a DTV set.

EIA/CEA-775.2: Service Selection Information for Digital Storage Media Interoperability

A digital storage device such as a D-VHS or hard disk digital recorder may be used by the DTV or by another source device such as a cable set-top box to record or time-shift digital television signals. This standard supports the use of such storage devices by defining service selection information (SSI), methods for managing discontinuities that occur during recording and playback, and rules for management of partial transport streams.

EIA/CEA-805: Data Services on the Component Video Interfaces

This standard specifies how data services are carried on component video interfaces (CVI), as described in EIA-770.1-A (for 2H 480p signals only), EIA-770.2-A (for 2H 480p signals only) and EIA-770.3-A. This standard applies to all CE devices carrying data services on the CVI vertical blanking interval (VBI). This standard does not apply to signals which originate in 1H 480I, as defined in EIA-770.1-A and EIA-770.2-A. The first data service defined is *copy generation management system* (CGMS) information, including signal format and data structure when carried by the VBI of standard definition progressive and high definition $Y P_b P_r$ type component video signals. It is also intended to be usable when the $Y P_b P_r$ signal is converted into other component video interfaces including RGB and VGA.

EIA/IS-702: Copy Generation Management System

This standard included packet description data relating to the copy generation management system (Analog) (CGMS-A). EIA/CEA-608-B incorporates this information. EIA-702 is withdrawn. See EIA/CEA-608-B.

EIA-679-B: National Renewable Security Standard (NRSS)

NRSS provides two physical designs. Part A defines a removable and renewable security element form factor that is an extension of the ISO-7816 standard. Part B defines a removable and renewable security element based on the PCMCIA ("C Card") form factor. The common attributes allow either an NRSS-A or NRSS-B device to provide security for applications involving pay and subscription cable or satellite television services, telephone, and all forms of electronic commerce.

EIA-766-A: U.S. and Canadian Region Rating Tables (RRT) and Content Advisory Descriptors for Transport of Content Advisory Information using ATSC A/65-A Program and System Information Protocol (PSIP)

This standard augments ATSC Standard A/65-A and SCTE DVS-097 Rev. 7, both titled Program and System Information Protocol for Terrestrial Broadcast and Cable (PSIP). Along with these two standards, this standard designates the RRT which provides the receiver with the definition of the rating system and the *content advisory descriptor* which provides the receiver with the specific program rating for each program. Specifically, this standard specifies the exact syntax to be used to define the U.S. and Canadian RRT in accordance with A/65-A Section 6.4 as well as the exact syntax to be used in the content advisory descriptors that convey the rating information for each program in accordance with A/65-A Section 6.7.4. Thus, DTV receivers may block unwanted programs as determined by the user.

EIA-796: NRSS Copy Protection Systems

The copy protection systems that have been included in EIA-796 are itemized for the purpose of identification. The systems outlined in EIA-796 all support the copy protection frameworks described in EIA-679-B, Parts A and B.

17.1.3d Cable Compatibility

EIA/CEA-818-B: Cable Compatibility Requirements

This standard defines the minimum requirements that shall be met by digital cable TV systems and digital TV receivers such that the receivers may be connected directly to the RF output of the cable system to provide selected baseline services.

EIA/CEA-819: Cable Compatibility Requirements for Two-Way Digital Cable TV Systems

This standard defines the minimum requirements that shall be met by two-way digital cable TV systems and two-way digital TV receivers such that the receivers may be connected directly to the RF output and input of the two-way cable system to provide specified services.

EIA/CEA-849: Application Profiles for EIA-775A Compliant DTVs

This standard specifies profiles for various applications of the EIA-775A standard. The application areas covered here include digital streams compliant with ATSC terrestrial broadcast, direct-broadcast satellite (DBS), OpenCable, and standard definition Digital Video (DV) camcorders.

EIA/IS-105.1: Decoder Interface Standard

Specifies an interconnection method for attaching a cable decoder to a piece of consumer electronics equipment such as a TV or a VCR. Two ports—an audio/video/control port and an IF port—are covered under this interim standard. This specifies the physical characteristics of the interface between the decoder and the receiver.

EIA-23: RF Interface Specification for Television Receiving and Cable Television Systems

This specification is intended to apply to all cable systems and to all receiving devices which may be directly connected to a cable system residential outlet, including, but not limited to, television sets, video cassette recorders, and converters (whether furnished by cable operators or independently acquired by subscribers).

17.1.4 References

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