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Robert R. Weirather, Gregory L. Best, Robert Plonka
Harris Corporation, Quincy, Illinois

INTRODUCTION

Significant advances have been made in TV transmitter technology in recent years. New technology and ideas have been introduced in order to continue to provide high quality TV signal transmission while improving reliability, reducing maintenance, and lowering overall cost of ownership. These new technologies fall primarily into the following areas: solid-state high-power amplifiers, efficiency improvements for UHF transmitters, and transmitter combiners which eliminate off-air switching. Each will be covered in this chapter.

Further, the FCC continued its policy of technical deregulation which has served to allow more flexibility in transmitter design and system operation.

VISUAL EXCITER

The visual exciter has the purpose of receiving a video baseband signal, processing it, and converting it to a fully modulated vestigial sideband on-channel signal. Since intermediate frequency (IF) modulation has become a standard within the industry, most of the signal processing occurs either in the video stages or in the IF stages. The basic block diagram of an IF modulated transmitter is shown in Fig. 1. We will discuss each basic block diagram in greater detail.

Video Processing

For the TV transmitter's goal of becoming transparent to the incoming signal, it is of utmost importance to make sure that the incoming signal is optimized. It is often difficult to define exactly where the station video processing of a signal ends and the transmitter video processing begins. Typically some video processing is done external to the transmitter at the station plant by a *proc amp*. Different manufacturers of transmitters employ various means of accomplishing the same purpose and therefore, may use different approaches to solve the same problem. The main functions of the exciter video processing circuitry are:

- Obtain proper sync to video ratio
- Remove any common mode signal
- Provide overall video level control
- DC restoration
- Overmodulation prevention
- Frequency response correction for the signal applied to the transmitter

Quite often precorrection, used to compensate for some transmitter distortions, is included in the video processing section. Types of transmitter precorrection which are sometimes included are differential gain,

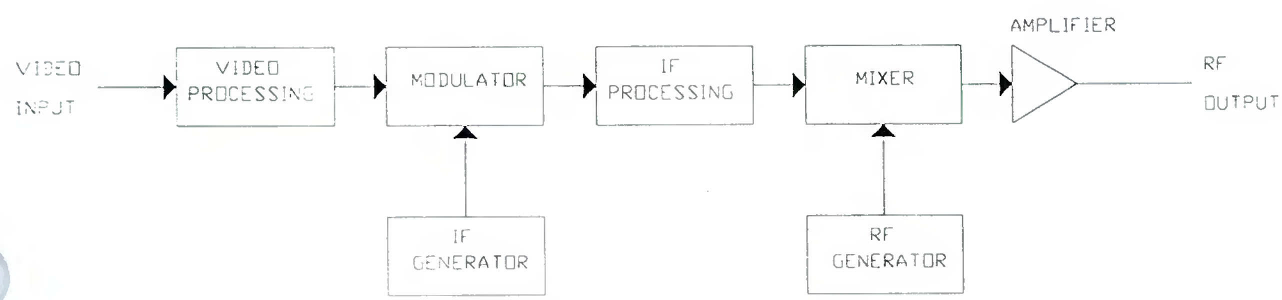


Figure 1. IF modulated transmitter.

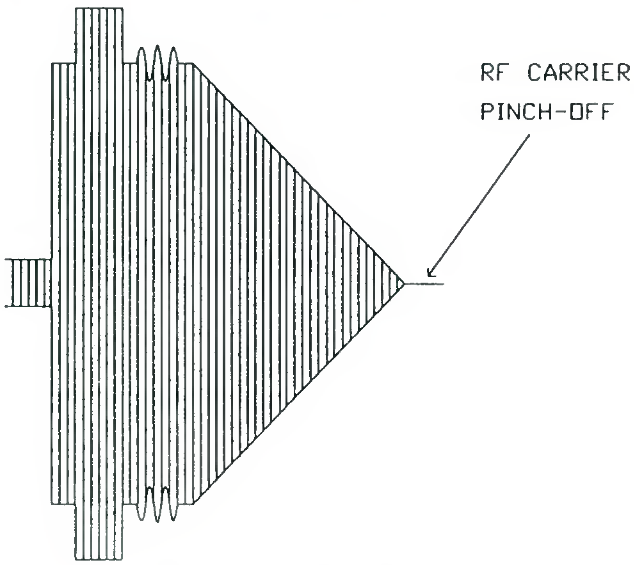


Figure 2. RF carrier pinch-off.

differential phase, and group delay compensation (for transmitter and/or for the notch diplexer, if used).

DC restoration, (clamping) is important because picture brightness information is contained in the DC component of the video signal. If AC coupling is used anywhere in the video circuitry, the DC level will tend to vary as the capacitors in the coupling circuits charge and discharge. AC coupling is convenient since differences in AC and DC grounds can be allowed without introducing waveform distortion. By clamping a consistent part (such as sync tip or "back porch") of a TV waveform during each TV line to a fixed voltage which does not vary, the correct DC level is applied to the modulator.

Common mode signals such as noise or AC hum can be removed by using a differential input mode for the video signal input stage. It is desireable to use an RF choke at the video input stages also to prevent rectification of ambient RF and remodulation of the main signal from that rectified RF.

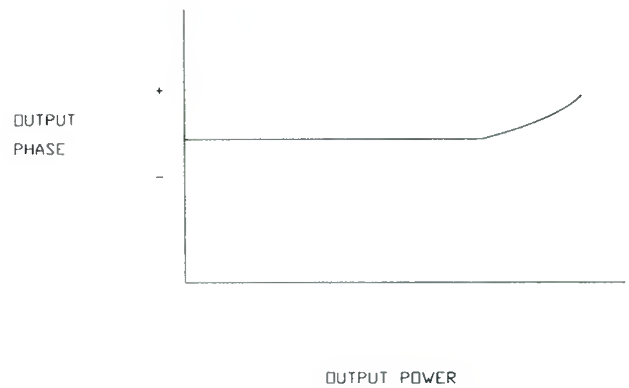


Figure 4. Phase transfer curve.

White peak limiting is used to prevent modulation from reaching 0% or in other words, to keep the carrier from being "pinched off" as shown in Fig. 2. When an intercarrier TV receiver encounters a carrier that has been pinched off, it temporarily has no signal to receive, its automatic gain control circuits increase to maximum gain, and since there is nothing but noise available when the carrier is pinched off, that noise is greatly amplified, transferred to the aural intercarrier and becomes audible as a buzz in the receiver.

High frequency peaking circuits in a transmitter's video processing circuitry are often used to compensate for long video signal runs in a transmitter plant.

In UHF exciters, often the horizontal and vertical sync portions of the TV signal are detected and sent to klystron pulsers.

When a transmitter is adjusted for maximum efficiency, its transfer characteristic is not ideal. Typical transmitter amplitude and phase transfer curves are shown in Figs. 3 and 4. Often the video signal is predistorted in the opposite direction of the errors in amplitude and phase produced by the nonlinear transfer curve of the transmitter. An example of this predistortion is given by Fig. 5.

Nonlinear chroma gain at different luminance levels that exhibits a change in the saturation of the colors, is termed *differential gain*. Nonlinear chroma phase at

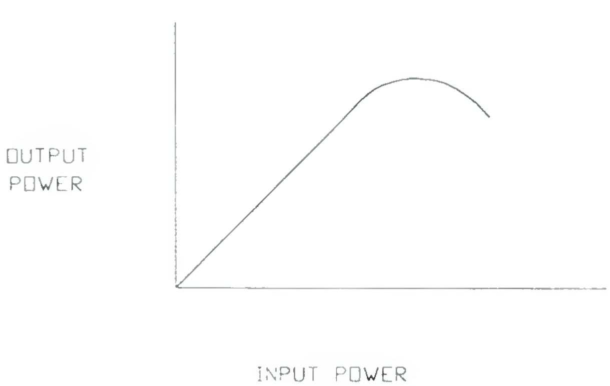


Figure 3. Amplitude transfer curve.

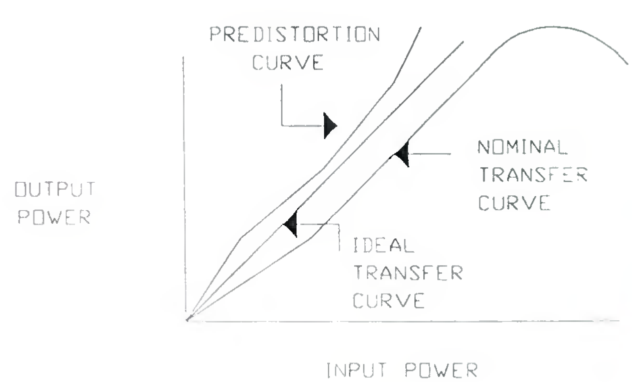


Figure 5. Sample predistortion curve.

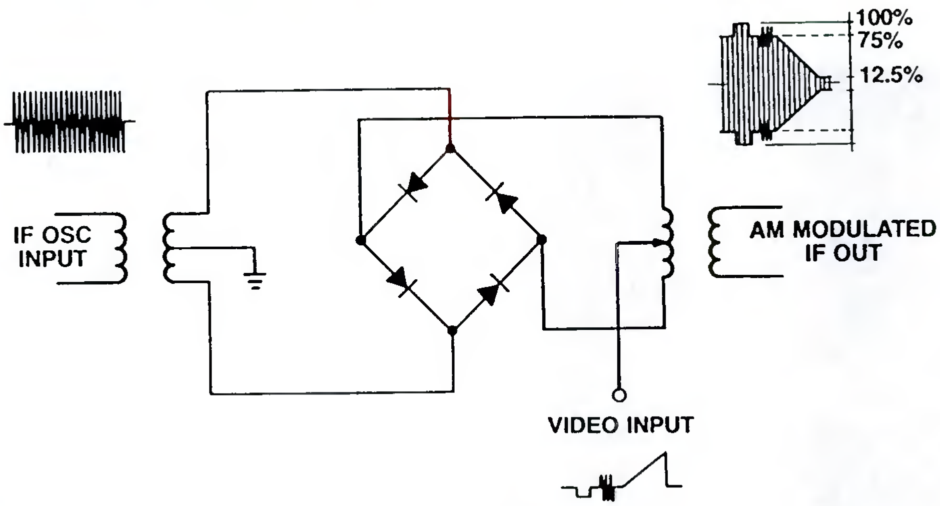


Figure 6. Balanced mixer.

different levels of luminance that exhibits a change in hue of the colors is termed *differential phase*. Differential gain and phase predistortion is often done in the exciter to compensate for nonlinearities in the transmitter.

Luminance nonlinearity is present when there is a change in luminance gain with different brightness levels.

Both luminance nonlinearity and differential gain are amplitude distortions. Precorrection can be accomplished by changing the gain of a video amplifier over a portion of the total video signal. The gain change is usually keyed by the video signal reaching a certain level. Differential phase can be corrected by splitting the video signal into two paths which are 90° different at the color subcarrier. By changing the gain of one path over a portion of the video waveform while not changing the other path phase changes can be obtained. Other methods can also produce the same result.

Modulator

With nearly all IF modulated transmitters, the modulator is a broadband, balanced, diode mixer. It is configured for maximum rejection of the local oscillator signal and biased so as to provide excellent linearity, low noise, and capability to achieve carrier cutoff. A schematic of a typical balanced mixer is shown in Fig. 6. The video signal is DC offset to provide the proper modulation level. The video signal is used to control the attenuation of the diodes. Peak of sync corresponds to maximum IF envelope output and white corresponds to minimum IF output. The output signal of the modulator is a double sideband AM signal having the proper depth of modulation (12.5%).

Group Delay Compensation

Waveform distortion can be caused by group delay inherent in RF amplifiers and tuned RF filters, combiners, and other output systems. Group delay distortion is the nonuniform delay of different frequencies over the bandwidth of the TV signal. Group delay distortion occurs in tube cavity amplifiers and notch diplexer combiners. In general, the closer the amplitude roll-off frequency is to the visual passband, the higher the group delay distortion.

IF and Video Delay Compensators

Group delay impairments show up in a picture as color smear and halo effects on edges. On a waveform monitor the effects may be seen using the 2T and modulated 12.5T contained in the composite test signal shown in Fig. 7. Pulse responses with unacceptable group delay may include exaggerated pre- or post-ringing on 2T and modulated 12.5T base line disturbance.

The multipulse signal shown in Fig. 8 may also be used to analyze group delay. The multipulse signal consists of a gray flag (80 IRE), 2T pulse, and five sine-squared pulses modulated with five discrete frequencies (consisting of one 25T pulse and four 12.5T

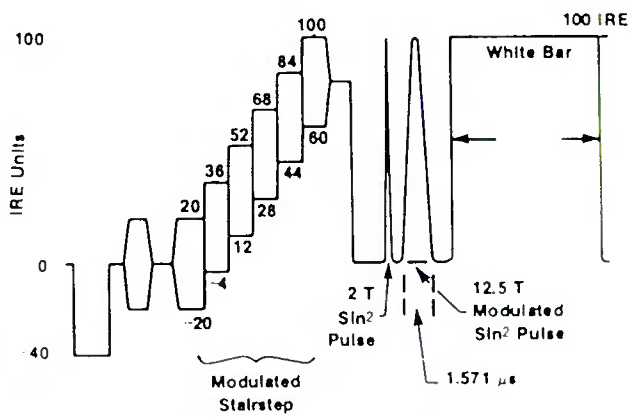


Figure 7. Composite test signal.

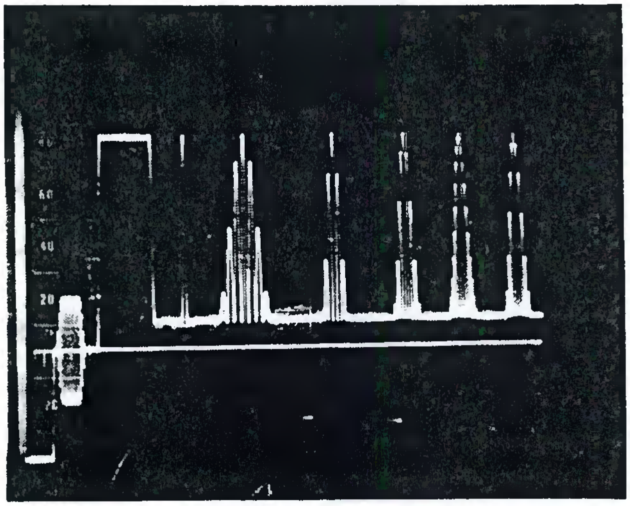
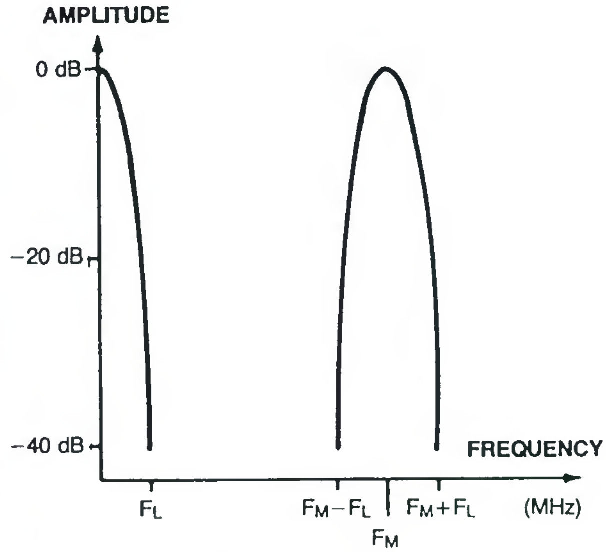


Figure 8. Multipulse.



$$F_L = \frac{1}{\text{H.A.D.}}$$

F_M = MODULATING FREQUENCY

Figure 9. Energy spectrum for modulated pulses.

pulses). This signal may be used to measure linear distortion in TV systems such as amplitude versus frequency response and group delay distortion.

The modulated pulses contain two spectra of information, low frequency and high frequency, as illustrated in Fig. 9. These modulated pulses are useful for measuring group delay errors. If the low-frequency spectra and the high-frequency spectra are delayed equally, the results will be a symmetrical modulated pulse with the same shape as the input. The gain versus frequency distortion will alter the base line flatness but will not change pulse symmetry. Delay errors will result in an asymmetrical pulse baseline. Combinations of group delay and gain errors are shown in Fig. 10.

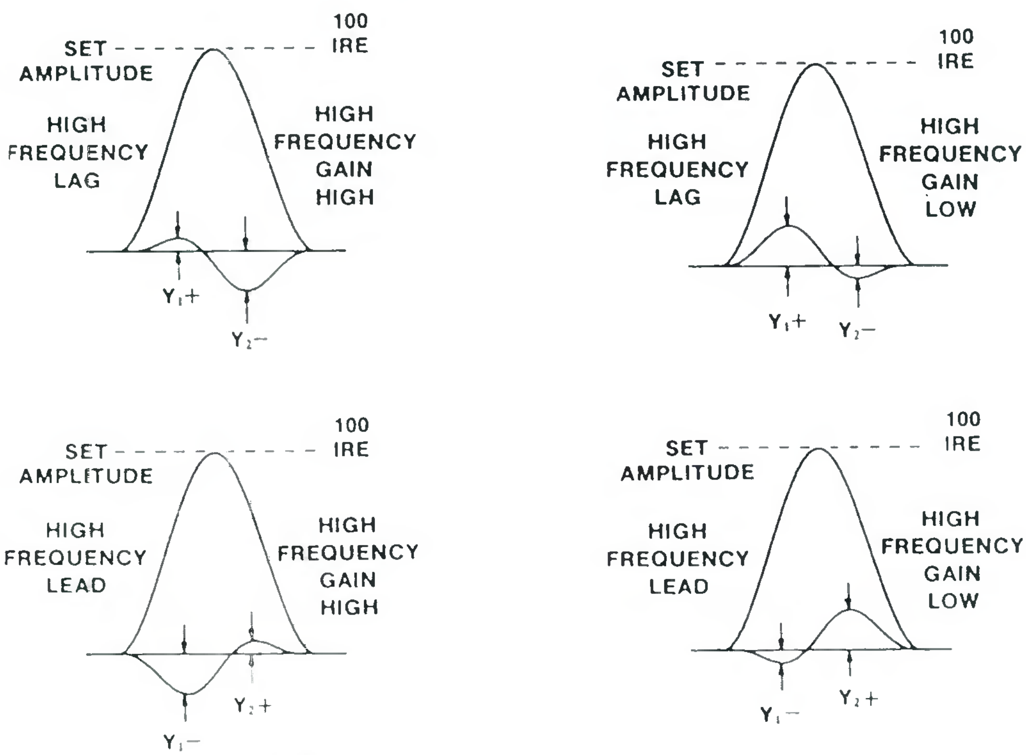


Figure 10. Modulated sine-squared pulses with gain and phase errors.

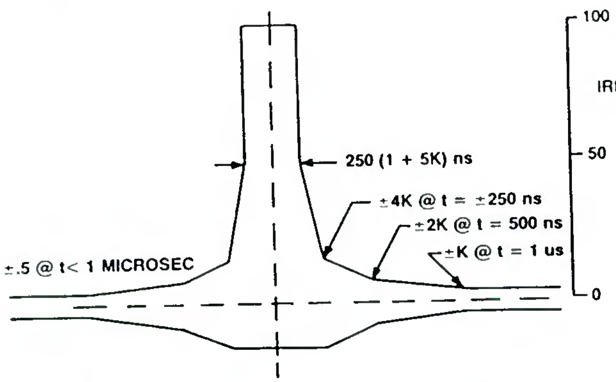


Figure 11. 2T sine-squared pulse graticule.

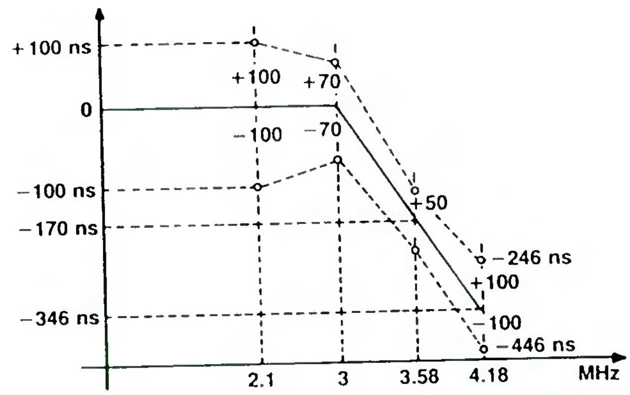


Figure 13. Predistorted group delay curve.

A method of quantifying these distortions uses the waveform graticule shown in Fig. 11. This graticule was arrived at empirically and represents constant perceptible distortion levels. This shows that overshoots closer to the desired pulse are not as perceptible as ringing further away.

The desired K factor graticule, i.e., 2%, is overlapped on a waveform monitor and the group delay corrector is adjusted until the 2T waveform lies entirely within the graticule. This technique is often preferred to swept group delay measurements because the results are in terms of perceptibility.

Low-frequency group delay and amplitude errors are referred to as short time waveform distortions. Fig. 12 shows typical distortions of a 2T pulse.

If the group delay error affects only the high frequency side of the passband, precorrection can be accomplished at video. This is the case for most visual-aural RF notch type diplexers and the FCC receiver equalizer curve. In the case of the notch diplexer, a pair of cavities resonant at the aural frequency are inserted between two 90° hybrid couplers. (Refer to Fig. 87.) The presence of the tuned circuits near the upper edge of the visual passband can cause a significant group delay error.

Another source of group delay is the tuned circuits of the high power tetrode or klystron amplifiers.

Finally the FCC requires the TV transmitter to predistort group delay according to the curve shown in Fig. 13.

The delay shown in Fig. 13 is a characteristic delay curve the FCC requires all visual transmitters to have. The purpose of it is to compensate for the delay error caused by discrete LC aural notch filters in early TV receivers. The reason this was done was that it would be far cheaper to group delay compensate one transmitter than every TV set. The notch filter in TV receivers is necessary to prevent the aural carrier from mixing with the detected video and chroma subcarrier signals and producing visible spurious beats.

Group delay predistortion may be accomplished at video baseband or IF. The techniques used are similar in concept. Both active and passive equalizers may be employed. IF group delay correction is necessary to correct group delay errors below visual carrier while not affecting the signal above visual carrier. Above visual carrier both IF and video correction is effective.

Passive Group Delay Equalizer

A common form of a passive group delay equalizer is shown in Fig. 14. It provides a flat frequency response and a nonlinear group delay which peaks at the resonant frequency of the circuit. This type of circuit is referred to as a passive all-pass network.

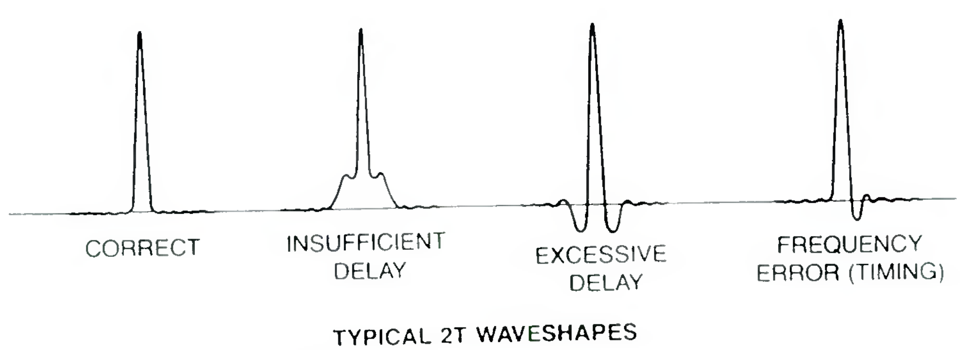


Figure 12. Typical distortion of a 2T pulse.

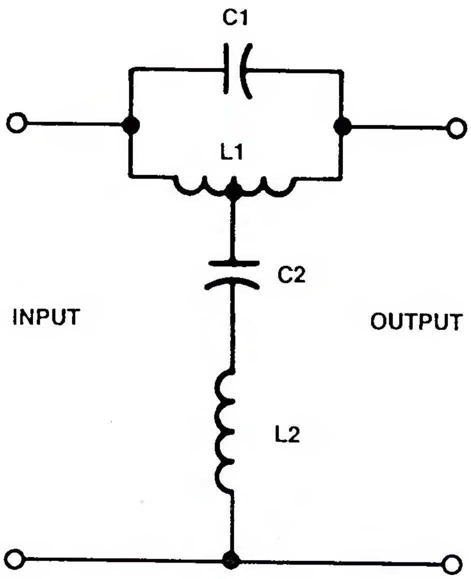


Figure 14. All-pass network.

Without going through a detailed analysis of the circuit, much can be understood by examining the circuit at frequencies well below and above resonance. At low frequency the network can be approximated by a low series inductive reactance due to L1. Here the output voltage leads in phase. At high frequency well above resonance the circuit can be approximated by a series capacitive reactance due to C1. At high frequencies the output voltage lags in phase. The output amplitude, however, is constant across the band. The phase of the network is plotted in Fig. 15. The slope of the phase is defined as group delay and is expressed as

$$\tau = (d\phi/d\omega)$$

where: ϕ = phase angle
ω = angular frequency

The steeper the phase slope the higher the maximum group delay. A plot of the group delay is shown in Fig. 16.

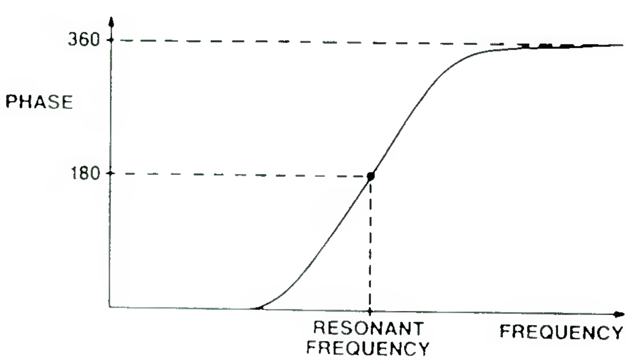


Figure 15. All-pass phase.

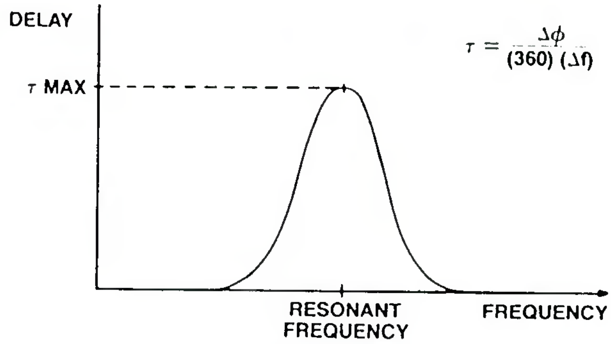


Figure 16. All-pass group delay.

Active Group Delay Equalizer

As in the passive type the network has a constant amplitude frequency response and a nonlinear phase response. The action of an active all-pass network can be best explained using the simplified schematic of Fig. 17. If the phase of voltage e_1 were plotted as a function of frequency, it would trace out a circle with maximum amplitude and zero phase at resonance. Voltage e_2 on the other hand has a constant amplitude and phase.

For the case when e_1 voltage is equal to twice e_2, the output voltage of the summing amplifier, e_3, has the characteristics of an ideal all-pass network. The output amplitude is constant and the resonator tuning determines the frequency of maximum group delay and the resonator Q determines the magnitude of delay.

For the case where voltage e_1 is larger than twice e_2, the output will have an amplitude peak at resonance and conversely if e1 is smaller, the output will have a dip at resonance.

Vestigial Sideband Filter

The FCC requires that the radiated TV signal have a major portion of the lower sideband (vestigial sideband) suppressed. In addition, the upper sideband signal must be contained within 4.75 MHz of the visual carrier. With the advent of IF modulation the filtering is at low-power stages and transmitter manufacturers have selected solid state filters using Surface Acoustic Wave technology to accomplish the stringent filtering requirements.

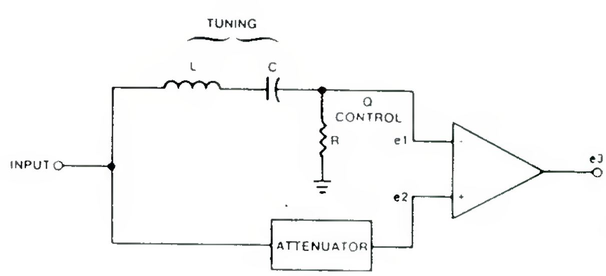


Figure 17. Active all-pass network.

The term *surface acoustic wave* (SAW) refers to the propagation of elastic waves on the surface of a piezoelectric crystal. The wave propagation is roughly the speed of sound, and therefore called acoustic. A time varying voltage on a metalized transducer is used to induce a deformation on the surface and produces a electromechanical wave. The deformations produce local electric fields which travel along with the mechanical wave, and interact with metal electrodes, which convert the mechanical wave back to a time varying voltage. The length, spacing, and the number of the electrodes determine the wave shaping properties of the filter.

The electrodes act like tapped delay lines, as illustrated in Fig. 18. The electrodes are designed to amplitude scale and delay the signal. The output is selectively attenuated depending on the time delay and signal frequency relationship. This type of filter is called transversal because the attenuation is controlled by delay lines rather than resonators. The transversal filter was invented in the early 1940s and used coax cables as delay lines. Today, surface acoustic wave transducers provide the same function. The wavelength of an IF acoustic signal is approximately 0.003″.

This small wavelength allows the filter to be very small and compact. Because the transducers are only on the surface, photographic masks can be used to accurately control the physical dimensions of the transducers. The photographic mask lends itself to modern manufacturing techniques and insures a reproducible filter with a permanent amplitude versus frequency response and group delay.

A characteristic of a transversal filter is that group delay can be set independently of the amplitude characteristic. This is not true of discrete LC filters where basic physical laws couple group delay and amplitude.

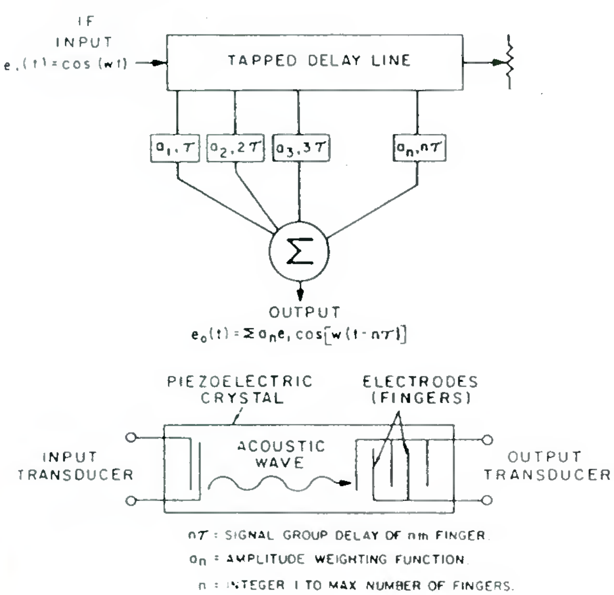


Figure 18. Saw filter.

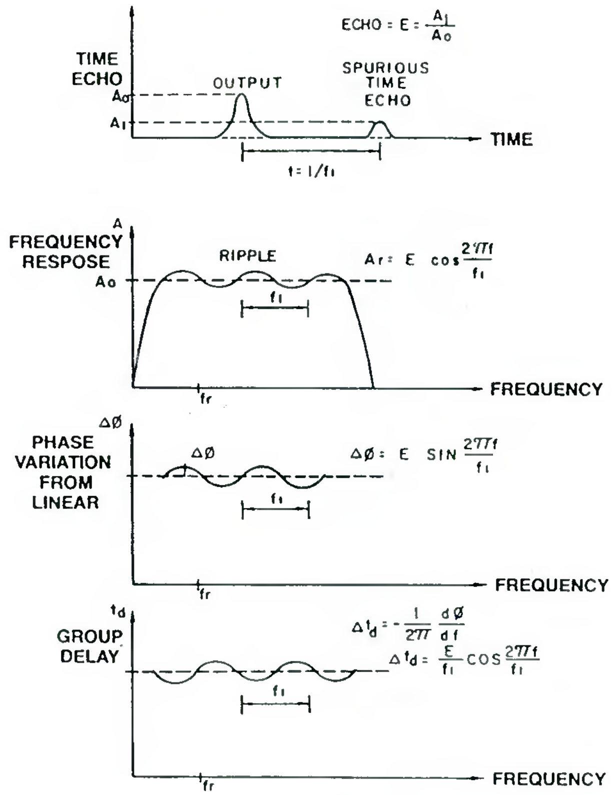


Figure 19. Response, phase, and delay ripples caused by time echoes.

The independent nature of group delay and amplitude allows a manufacturer to provide almost any type of group delay curve (such as a constant delay or the TV receiver group delay precorrection curve) across the passband.

Although the FCC does not require attenuation of visual signals in the aural passband, video signal components can cause interference called visual-to-aural crosstalk. Reduction of visual-to-aural crosstalk is essential to the proper transmission of stereo sound and other subcarrier broadcast services. High resolution cameras and character generators can produce spectral components at 4.5 MHz and beyond. Without sufficient attenuation these signals can cause distortion, especially in aural subcarriers which are more sensitive to visual crosstalk than the main aural signal. SAW filters thus are able to perform many functions simultaneously, i.e., bandwidth shaping, group delay equalization, and video to aural crosstalk reduction.

Distortion products that can be created by SAW filters are due to signal reflections (triple travel) and direct feedthrough. These distortions cause time echoes displaced either before or after the desired responses. In the frequency domain these echoes show up as ripples in the passband. (Fig. 19) A given time echo will contribute uniquely to the amplitude, phase and group delay characteristics by the addition of a ripple component of the waveform in the passband.

The amplitude of the ripple is proportional to the echo level, and its period is proportional to the reciprocal of the SAW time delay.

The SAW group delay characteristic will vary sinusoidally with the same period as the passband ripple. Its magnitude, however, is a function of both echo level and inversely proportional to time delay, and therefore not a sure test of signal distortion. The group delay ripple for long delayed echoes will give peak-to-peak values which have no correlation with conventional signal distortion estimates. For that reason, fast peak-to-peak group delay errors that occur closer than the reciprocal of the filter time delay can usually be ignored.

When pulse waveforms in Figs. 7 and 8 are used to measure group delay distortion, the 2T preshoot, overshoot, and asymmetry can be used to gauge low-frequency group delay error. Modulated pulse baseline "S" curve is a result of group delay and the peak excursions of the "S" curve can be used to quantify the amount of group delay distortion.

IF Linearity Precorrection

Amplitude and phase nonlinearities are sometimes corrected at video baseband. The alternative is to provide the correction at IF frequencies. There are advantages to correcting at IF in that, since most distortions are caused in the high power RF amplifiers after vestigial sideband filtering, a corrector placed after the vestigial sideband filter can more accurately predistort the modulated signal. Any precorrection spectra generated at IF after the VSB filter will produce energy components which can cancel intermodulation products generated in the final amplifier stage. This is particularly important in the case of a pulsed klystron transmitter.

Correction of distortion at IF is particularly helpful for chroma distortions. Chroma spectra at 3.58 MHz has only a single sideband information and thus has less energy than equivalent luminance signals. An ideal diode detector frequency response of the NTSC modulation signal is plotted in Fig. 20. Here it is seen

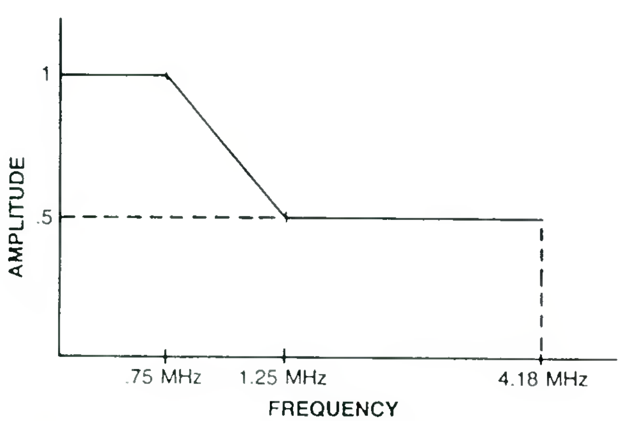
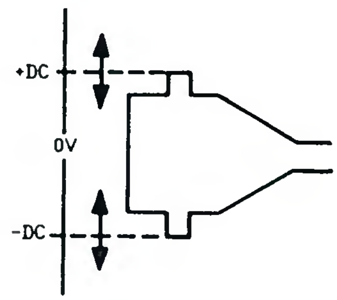


Figure 20. Frequency response using ideal diode detector.



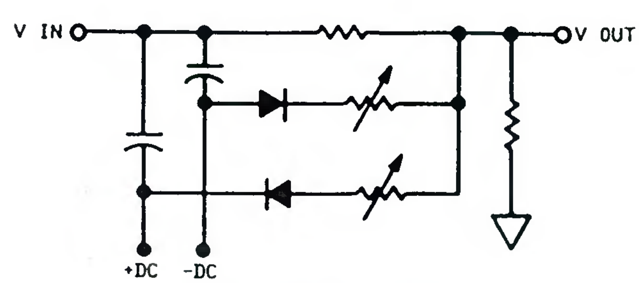


Figure 21. Basic gain expansion circuit.

that beginning at 0.75 MHz the video begins to fall to −6 dB. For video signals lower than 0.75 MHz, the RF spectrum is double sideband and has twice the peak RF voltage.

Intermodulation products are caused in high power amplifiers by nonlinear amplitude transfer characteristics. As the power output increases towards saturation, amplitude compression begins and in general the phase begins to lag as well. The nonlinear transfer characteristics give rise to mixing products which occur at sum and difference frequencies around the visual carrier. This process also creates the frequency spectra of what is called the lower sideband.

Correctors for amplitude distortion are usually similar in concept to video differential gain correctors. Linearity correctors generally use diodes which are keyed by the IF modulated signal. When the diodes conduct, the gain or attenuation is reduced as needed.

An example of a basic gain expansion circuit is shown in Fig. 21. The signal is normally attenuated a fixed amount by using a resistive L-pad. The diodes are normally reverse biased by equal, but opposite polarity, DC voltages. Reducing the DC voltage amplitude permits the diodes to conduct on the signal peaks. This inserts additional resistance in parallel with the series arm of the L-pad attenuator thereby decreasing the attenuation. Varying the resistance in series with the diodes provides for a variable gain expansion.

ICPM

Phase distortions in high power amplifiers produce incidental carrier phase modulation (ICPM), which are spectral components in quadrature with the modulation signal. Fast video amplitude changes such as a step or pulse will cause larger incidental phase spectral components than slow changes. Receivers make this condition worse by attenuating the lower sidebands below 0.75 MHz. The receiver then responds to the extra sidebands created by the phase modulation as if they were amplitude modulated single sidebands and produce spikes. The faster the rise time on the signal the more high frequency energy is present resulting in edge distortions in the displayed picture.

The picture impairment is similar to simultaneous group delay and differential phase errors in that edges are less sharp and color hue changes with brightness. On a waveform monitor, overshoots are visible as trailing edges and as rounding on leading edges. These overshoots vary in severity depending on how close the power amplifier is driven towards saturation.

Audio impairment is produced by ICPM in receivers employing intercarrier conversion. Intercarrier receivers use an AM or synchronous detector to produce a 4.5 MHz aural IF signal from the composite video IF. Any phase modulation present on the visual carrier is then transferred to the aural intercarrier. In the monaural baseband audio the increasing amplitude frequency effect of ICPM is nullified by de-emphasis to some degree. With multichannel sound, however, there is no de-emphasis applied to the baseband stereo signal, and thus the distortion is more pronounced at the stereo subchannel and pilot frequencies. To counteract the effects of ICPM and other noise sources on the stereo subchannel, an audio companding is employed that greatly reduces the potential interference. Although the audio companding process can reduce some of the effects of ICPM, ICPM correction is essential in delivering clear, low-noise audio to intercarrier receivers.

There is no defined level of ICPM for a given stereo performance level since the signal to buzz ratio is highly dependent on the picture spectral components.

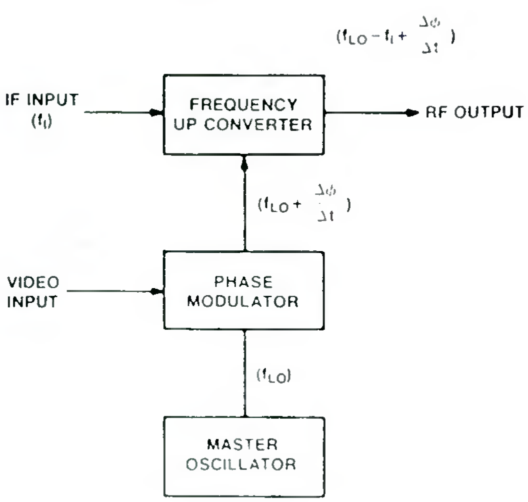


Figure 22. Master oscillator phase modulator block diagram.

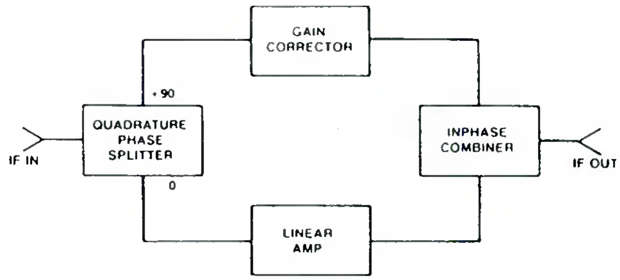


Figure 23. Direct IF ICPM corrector.

Refer to the EIA Recommended Practices for the current recommendations on ICPM limits.

ICPM precorrectors can be grouped into two types: ones using a phase modulator and the other operating on the signal directly. The phase modulator type uses video to vary the IF or master oscillator phase with the opposite phase characteristics of the nonlinear amplifiers. A phase modulator can also operate on the IF signal directly using a video signal to set the amount of modulation. A block diagram of a master oscillator phase modulator is shown in Fig. 22.

ICPM precorrectors operating directly on the IF signal can be implemented several ways. Direct precorrection at IF is similar in concept to baseband differential phase precorrection. See Fig. 23. In both cases, the visual signal is split into two paths which are in phase quadrature. In the IF corrector, the entire channel of frequencies is in quadrature, whereas in the video precorrector only the chroma band is in quadrature. One method of implementation is to modify the quadrature signal gain function with level dependent diode expansion or compression circuits. This can be done using the same techniques as in the linearity corrector.

RF Generator

Transmitters employing IF modulation generate the following frequencies; visual IF, modulated aural IF, and master oscillator signal(s) for translating visual and aural IF to the final carrier frequencies. These oscillators have been implemented with either digital synthesizer techniques or crystal oscillators. An advantage of the synthesizer is that only one crystal is needed and it operates at a single standard frequency for all TV channels. The crystal oscillator approach, however, may involve simpler circuitry.

The two commonly used IF frequencies are 37 MHz and 45.75 MHz. There are many reasons for selecting one IF frequency or the other. One advantage of 37 MHz is that the temperature drift sensitivity of most IF components such as the SAW filter is related directly to carrier frequency. Thus the lower IF has a 12% less drift sensitivity than components at 45.75 MHz. The second harmonic of 37 MHz falls in between channels 4 and 5 so as not to cause interference. On the other hand, 45.75 MHz is a common demodulator IF which can be useful for IF troubleshooting. Temperature drift may be minimized at either IF by maintaining the SAW filter at a stable temperature.

The important performance characteristics of an oscillator are: its phase noise, sensitivity to frequency drift with time and temperature, and susceptibility to mechanically induced phase and frequency shifts called microphonics.

In replacing a crystal, it is important to follow the recommendations of the oscillator manufacturer to insure proper operation. Synthesizer performance should be properly maintained to prevent inadvertent phase noise and spurious frequency generation.

OFFSET FREQUENCY CONTROL

The limited number of available channels for TV Broadcasting makes it necessary to assign the same carrier frequencies to many stations. To avoid interference between stations operating on the same frequency (co-channel interference) geographical separation and radiated powers are carefully selected.

Nevertheless, considerable co-channel interference was encountered in many locations. Investigation has shown that additional means were able to reduce co-channel interference by reducing optical or perceptive interference elements in the system. In other words, steps were taken not to alter the strength of the interfering carrier but to choose operating parameters to reduce the visibility of the interference.

Considerable investigative work was done to lay the scientific and physiological basis and to define the parameters which need to be controlled to reduce the visibility.

It turned out, with horizontal and vertical sweep ratios fixed, the visual carrier frequencies of the interfering stations was the parameter which needed to be controlled.

Co-channel interference between television stations appears to viewers as a horizontal pattern of alternating light and dark bars on the viewing screen—very much like the shadows cast by venetian blinds. It has been demonstrated for many years that the visibility of these bars varies cyclically as a function of the difference in frequency of the interfering carriers (Fig. 24). The interference is most visible when the carriers were offset by multiples of the line frequency (15734 kHz) and least visible when the carriers are offset by odd multiples of one-half the line frequency. In addition to the gross maxima and minima, fine grain maxima and minima occur when the frequency offset is a multiple of the frame frequency (29.97 Hz).

Ideally, stations would be offset by odd multiples of one-half the line frequency to provide minimum interference visibility. However, a third station in the same area would be offset from one of the other stations by an even multiple of the line frequency. Hence, maximum visibility of the interference would occur. Therefore, 10 kHz offsets currently used in the United States were chosen to provide approximately equal reduction of the interference patterns for any number of stations in geographical proximity. (See Fig. 25.)

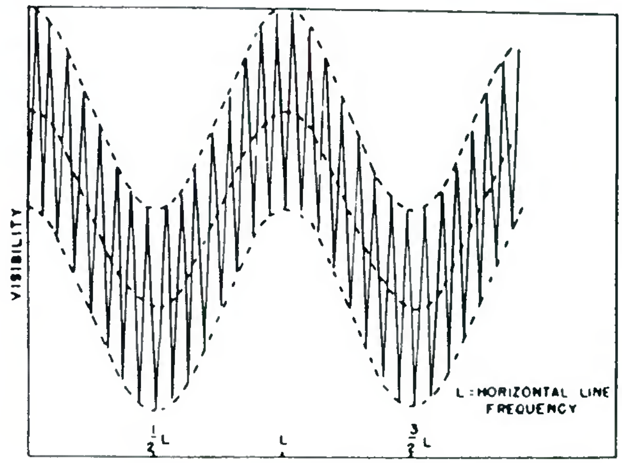


Figure 24. Co-channel interference.

Precise Frequency Offset

Although it is not practical to utilize the gross minima occurring at odd multiples of one-half the line frequency, it was determined experimentally that utilizing the fine grain minima, occurring at even multiples of the frame frequency, would be very advantageous in reducing the visibility of co-channel interference.

The nearest even multiple (334th) of the frame frequency to the 10 kHz offset is 10010 Hz. In a three station arrangement, one station will have zero offset, and the other two stations will be offset by ±10010 Hz. Experiments indicated that changes in the frequency differences of 5 Hz had a negligible effect on the reduction of the interference visibility.

To maintain the precision offset within 5 Hz requires maintaining each visual carrier frequency within 2 Hz or 3 Hz. Maintaining a television transmitter to such tight frequency tolerances requires some type of control system using an extremely stable frequency source.

In exciters using two independent oscillators (one for IF and one for the local oscillator), the visual carrier signal may be derived from mixing the oscillators together and comparing that signal to the reference signal in the comparator. That resultant error signal can be used to adjust one of the oscillators, preferably the local oscillator. For exciters using synthesizers, the sythesizer reference oscillator is compared to the precise frequency standard in the phase detector and the resultant error voltage is used to adjust the synthesizer oscillator.

By phase-locking the visual carrier to a stable reference oscillator, the master oscillator acquires the stability of the reference source. Sources which use an internal WWV Receiver/Comparator to self-correct or an Atomic Frequency Standard can easily provide the stable reference source required. These reference sources allow a transmitter to be maintained within a few hertz of a desired frequency indefinitely. Experimental results have indicated that frequency differences between transmitters can vary as much as 5 Hz

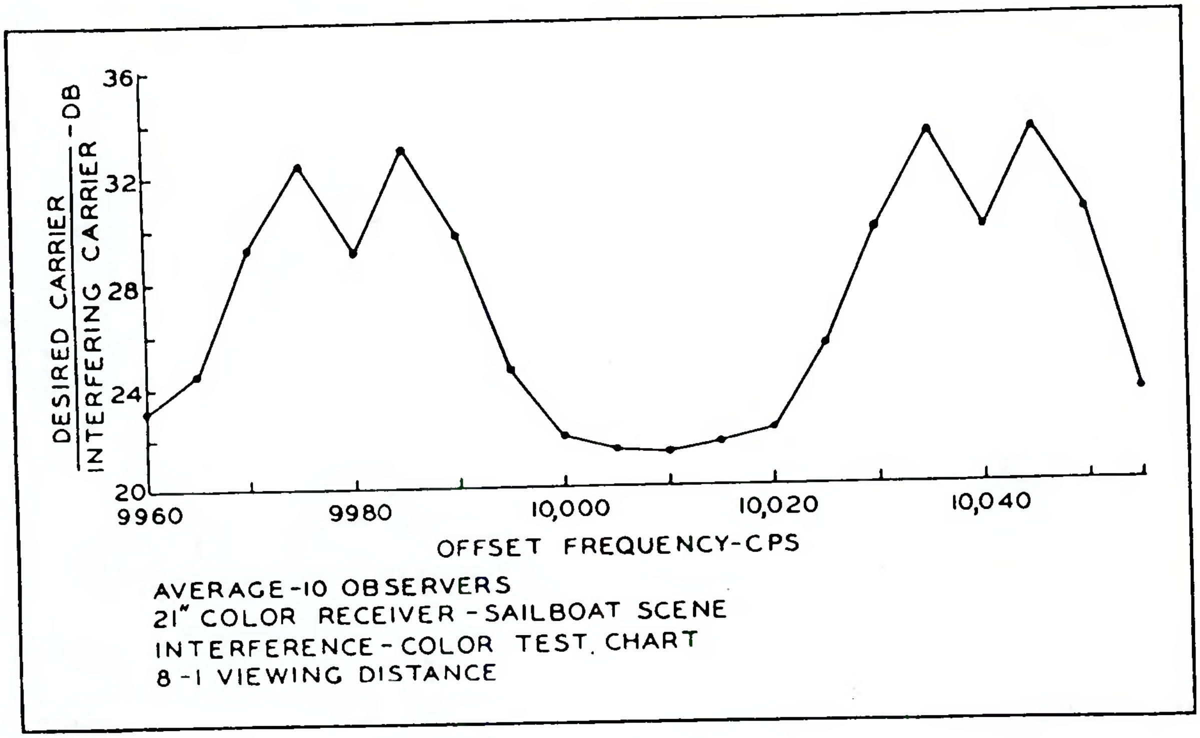


Figure 25. 10 kHz offset pattern.

from the precision offset before co-channel interference becomes noticeably worse. This means that two television stations can minimize their co-channel interference without the requirement to continuously adjust the transmitter frequency.

One word of caution when working with any frequency standard. If power is removed from the standard and no backup power supply is incorporated either internally or externally, ensure that the standard has stabilized before making any adjustments. Also ensure that after any adjustment has been made, a suitable time period has elapsed to allow the standard to settle before making any reading.

To ascertain the carrier frequency of a station operating under precision frequency control, assuming a measuring accuracy ten times better than the quantity to be measured, means measurement equipment must have the following accuracy requirements:

VHF Low Band	$+2.5 \times 10^{-9}$
VHF High Band	$+1 \times 10^{-9}$
UHF	$+3 \times 10^{-10}$

Frequency counters with the above accuracy and with National Institute for Standards and Technology (NIST) broadcasts, traceable calibration may be used for making measurements.

AURAL EXCITER

In its most basic form, the aural exciter consists of some audio processing, an FM modulated IF oscillator, and an up-converter to obtain the desired carrier signal as shown in Fig. 26.

Audio Processing Circuits

To ensure that the transmitter is not the limiting factor in audio (monaural and stereo reproduction), it is desired that the transmitter add as little distortion to the incoming signal as possible.

Baseband audio of the BTSC[1] MTS (Multi-channel Television Sound) system (including the mono, stereo, SAP, and profession channels) include frequency components to 105 kHz. Emphasis must be placed on phase linearity, low distortion, reduction of any amplitude ripples, or roll-off over the stereo (BTSC) passband to achieve good stereo separation and minimum crosstalk between the main stereo and the SAP channels.

While unbalanced coaxial inputs are used for the MTS input to the aural exciter, it is still necessary to use some form of common mode rejection to reduce the possibility of hum and noise from getting directly into the audio stages.

All errors in phase linearity and amplitude response within the audio circuitry contribute to stereo separation degradation. *As a general rule, amplitude roll-off*

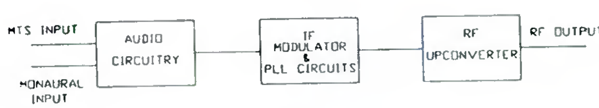


Figure 26. Aural exciter.

should be less than 0.1 dB and departure from phase linearity should be less than 1 degree for good quality stereo.

IF Modulation

All modern transmitters are IF modulated. To ensure low power stages do not contribute any group delay or amplitude roll-off, wideband amplifiers should be used. Modulated oscillator linearity requirements include flat modulation sensitivity versus frequency characteristics up to 47 kHz minimum, and typically out to 120 kHz.

With the advent of the MTS stereo system, intermodulation (IMD) and harmonic distortion (THD) products which, in monaural operation lie above 15 kHz (which would therefore be reduced by deemphasis in the receiver), now lie in the stereo channel or the SAP channel and will degrade stereo separation or will cause crosstalk into the SAP channel. In addition, IMD products generated in the stereo channel may now lie in the mono or SAP channel.

As the MTS signal is up-converted to the aural RF carrier frequency, the residual FM of the local oscillator signal becomes a determining factor. The level of FM produced by the local oscillator should be 10 dB lower than the modulated oscillator. Synthesized sources should be inspected for spurious frequencies which may show up as FM noise.

RF Amplifiers and Diplexers

All RF amplifiers are optimized (for stereo performance) when a flat, symmetric frequency response and a minimized variation in group delay across the modulation passband is achieved. Since FM modulation and demodulation is a nonlinear process there is not a one-to-one correspondence between RF amplitude/phase response and baseband stereo separation and crosstalk. Measurements indicate that a 1.5 MHz, 3 dB bandwidth will provide excellent stereo and SAP performance. RF amplifiers must be swept-frequency analyzed to insure the passband is sufficiently wide to pass the stereo signal and that its passband amplitude and group delay characteristics are symmetrical about the carrier frequency.

The diplexer is the last and possibly the most critical element in the aural chain with regard to stereo separation and crosstalk between stereo channel and SAP. However, hybrid diplexers, since they are broader bandwidth than notch diplexers, normally do not present any degradations to multichannel sound. If correct tuning of the notch diplexer itself and of any associated group delay compensation or amplitude correction is not properly maintained, stereo separation and crosstalk will be degraded. Low insertion loss at 4.18 MHz and proper aural bandwidth are both required for optimum visual and sound performance. Refer to the EIA Recommended Practices (Systems Bulletin #5) for other current recommendations.

Minimizing antenna reflections (VSWR) helps to keep the pilot signal from becoming distorted and the audio channel from losing separation or becoming noisy.

In summary, these characteristics have been incorporated into transmitters designed for MTS:

1. Wideband low distortion audio stages with excellent signal-to-noise ratio (SNR) are employed to insure negligible distortion in the baseband signal.
2. Low distortion wideband modulated oscillators with improved noise performance and phase locked loop (PLL) techniques are used so that the quality of TV multichannel sound will approach that of the FM broadcast service.
3. The bandwidth of IF and RF stages is wide. Power bandwidth tradeoffs may need to be made to achieve optimum stereo performance.

COMPENSATION OF AURAL PASSBAND FOR OPTIMUM STEREO PERFORMANCE

This section describes circuits that offer group delay equalization for a TV aural transmitter when operating through a notch diplexer. These circuits introduce group delay correction in the IF section of the aural exciter and after up-converting to the operating channel, the group delay precorrection effectively corrects the adverse group delay existing in the output diplexer circuit. The end result is improved TV stereo separation. In addition, precorrection of the FM bandpass can allow the use of lower cost, single-cavity notch diplexers. Stagger-tuned dual-cavity notch diplexers also have been utilized to provide a broad bandwidth desired for good stereo separation and negligible crosstalk between the different MTS components. However, dual-cavity notch diplexers introduce more group delay in the visual path, are more expensive, and should be tuned differently than single-cavity diplexers.

Many TV engineers have noticed a significant difference in the level of distortion when making performance measurements on an aural TV transmitter at points before the notch diplexer and after the notch diplexer. With the introduction of TV BTSC stereo broadcasting, the increase in THD distortion and stereo separation errors caused by the diplexer became a major problem which had to be solved.

The notch diplexer is a passive device and under first consideration it may seem strange that it can introduce nonlinear distortion and stereo separation errors. The basic problem is that the FM stereo signal is sensitive to the notch diplexer group delay and amplitude response over the occupied bandwidth of the FM signal.

The group delay and amplitude response of a single-cavity diplexer is shown in Fig. 27. The phrase, "single-cavity diplexer", means a single aural notch cavity is used in each branch of the diplexer. Dual-cavity notch diplexers use two aural notch cavities in each branch of the diplexer. The response curves shown in Fig. 27 are typical of a notch diplexer optimized for minimum aural reject power. The bandpass is somewhat narrow and the group delay is steep. Fortunately, the response

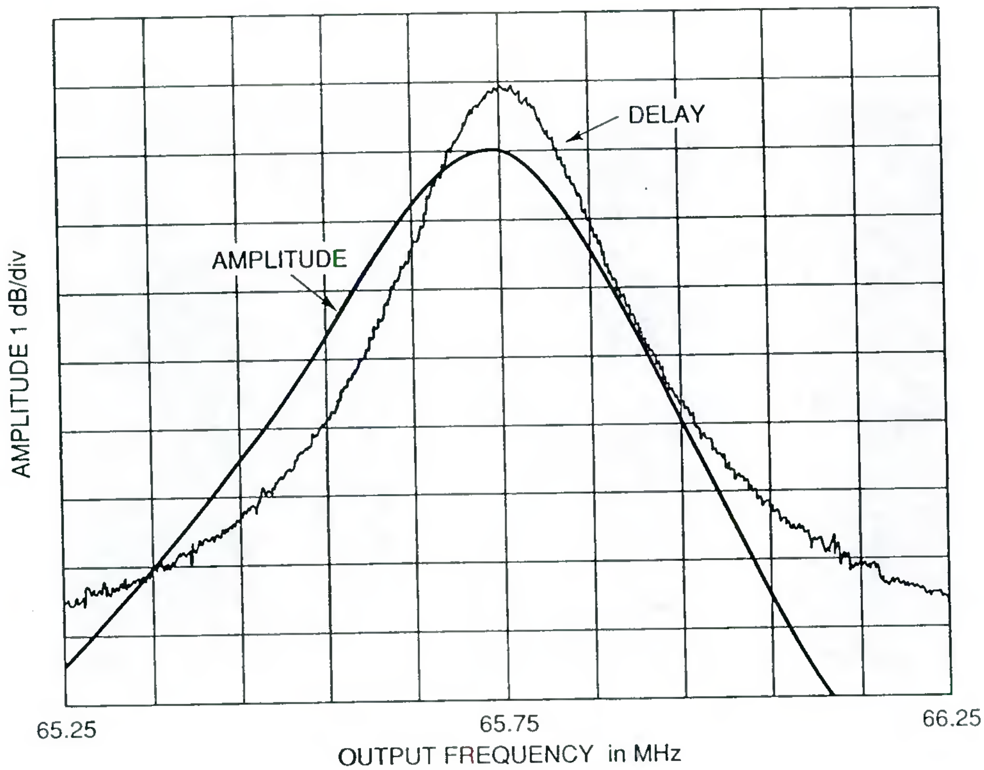


Figure 27. Typical single cavity diplexer amplitude and group delay.

curves in Fig. 27 show a high degree of symmetry which makes precorrection possible.

The group delay equalization concept as applied here to the FM modulated signal is essentially a feed-forward correction scheme. The phrase "feed-forward" is used here to describe the technique of generating correction signals early in the RF line up of a transmitter for the purpose of correcting distortions occurring downstream in the system.

Feed-forward correction signals are operated open loop (without feedback) and are manually adjusted for optimum operation and, once adjusted, left to operate in this manner.

IF Group Delay Correction

The equalization technique of adjusting the circuit to produce an inverse curve of that of the diplexer for correction is the same as that used on the visual transmitter.

There is however, a significant difference between group delay correction on a FM system versus an AM system. The aural transmitter must incorporate group delay correction at RF or IF to compensate for group delay distortions occurring in the output diplexer system. *Group delay correction before the FM modulation process is not effective for equalizing group delay occurring at RF.* The reason for this is that the occupied bandwidth of an FM signal increases or decreases as a function of baseband signal level. Fig. 28 shows the occupied bandwidth when the carrier is deviated 25 kHz and how it compares to a typical notch diplexer group delay curve. (Also note the precorrected group delay curve shown as an additional overlay.) When the baseband signal level is increased, the FM deviation will increase and a number of additional significant sidebands will be generated which will begin to extend beyond the acceptable group delay curvature region. The result is that distortion is observed in the demodulated FM signal.

In addition, as the baseband high frequency content increases, i.e., when switching from a mono to a stereo signal, significant sidebands extend even further outward increasing the demodulated distortion. Fig. 29 shows the spectral content of a TV stereo signal (with notch diplexer group delay curves, standard and precorrected, overlayed).

Simply put, there is not a direct relationship between the amount of group delay correction injected at baseband versus the amount of group delay correction achieved at the output of the RF system for a FM modulated signal.

To avoid the additional cost of extra cavities in the diplexer to achieve better stereo performance, an adjustable group delay circuit designed for the low power IF section of the aural exciter can provide the inverse group delay curve necessary to equalize the output. The corrector location in the aural transmitter system is shown in Fig. 30.

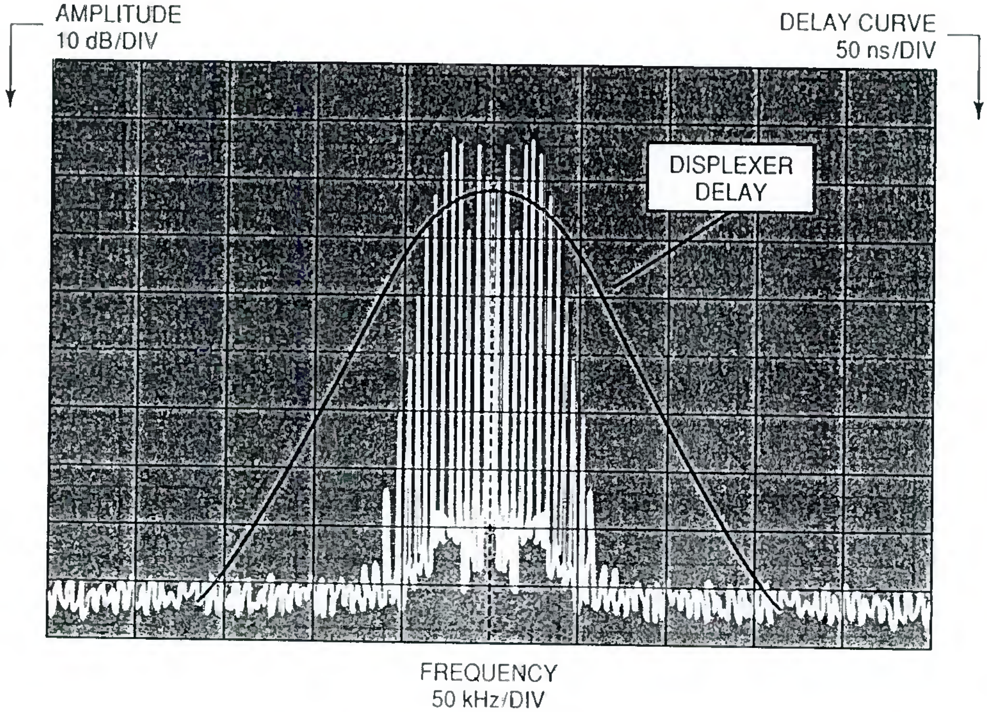


Figure 28. Occupied bandwidth with 25 kHz deviation with overlayed notch diplexer bandwidth.

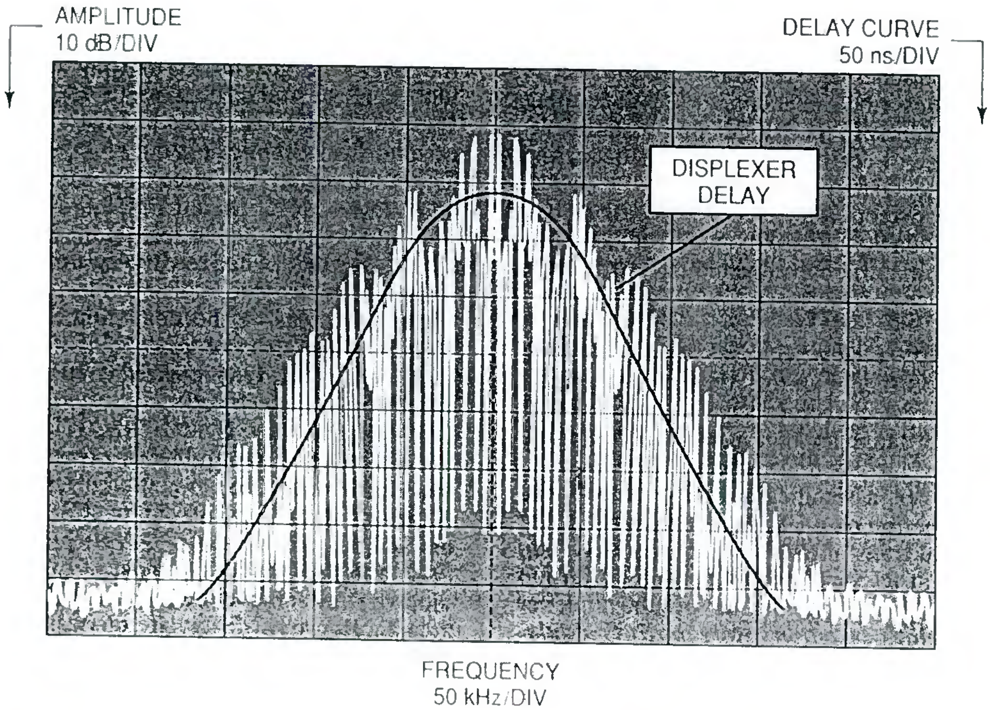


Figure 29. Occupied bandwidth of TV stereo signal with 55 kHz deviation and overlayed notch diplexer bandwidth.

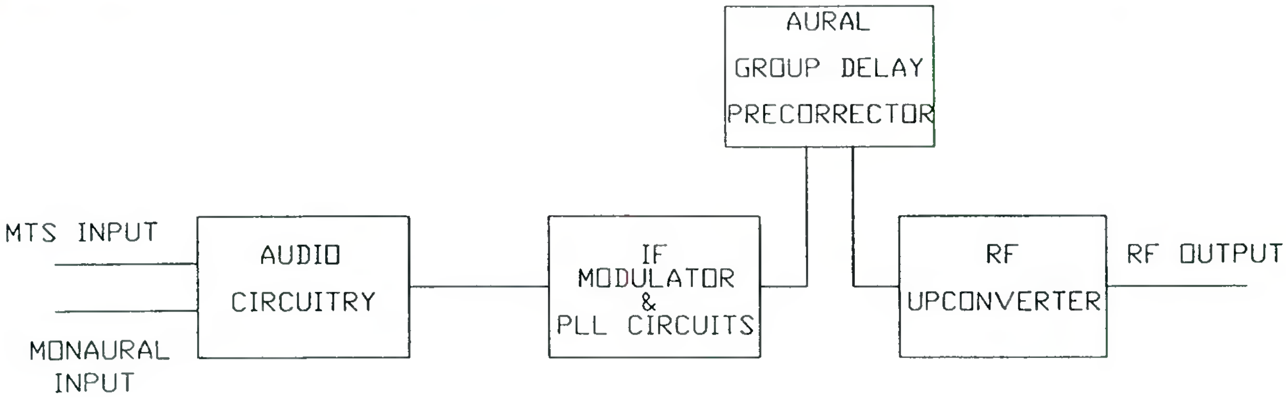


Figure 30. Aural exciter with group delay precorrector.

Equalization at IF frequency is effective because the up-conversion circuit uses a signal mixing process rather than a frequency multiplying process. In this manner, the precorrected sidebands are passed directly to the output for compensation.

Circuit Design

The circuit configuration is shown functionally in Fig. 31.

Results

The measured amplitude and group delay response of the group delay corrector is shown in Fig. 31. An ideal precorrection circuit without any circuit losses would have a flat response without a dip. The response dip shown in Fig. 32, however, is very useful because it also provides a first order correction to the notch diplexer amplitude response.

Fig. 33 shows the overall system equalization when the delay corrector is switched in and out of the circuit. The diplexer delay response contributes nearly all of the delay errors. The lower curve shows that a significant amount of equalization has been achieved over the occupied bandwidth of a stereo signal.

The effect on stereo separation is shown in Fig. 34 with the corrector switched in and out of the circuit. The curves generated in Fig. 34 are from measured data through the system with the response characteristics

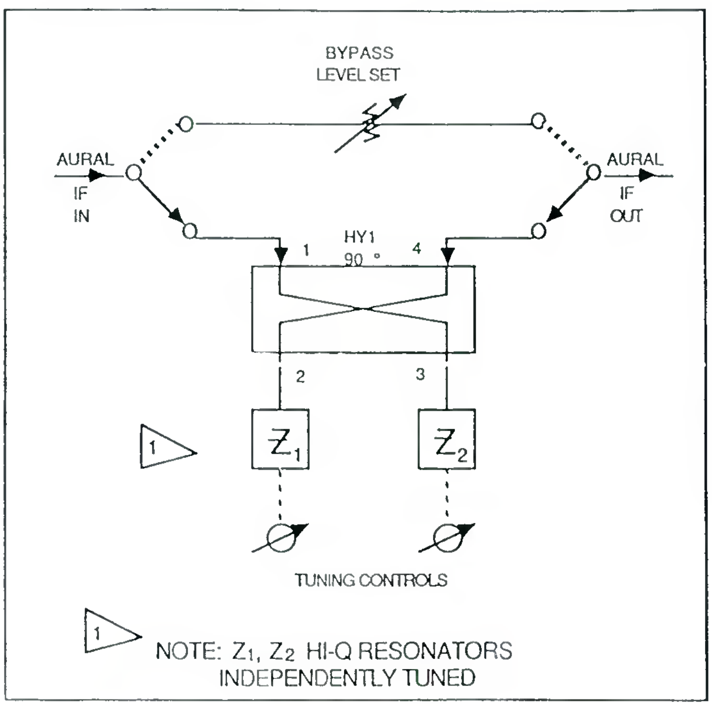


Figure 31. Aural group delay corrector functional block diagram.

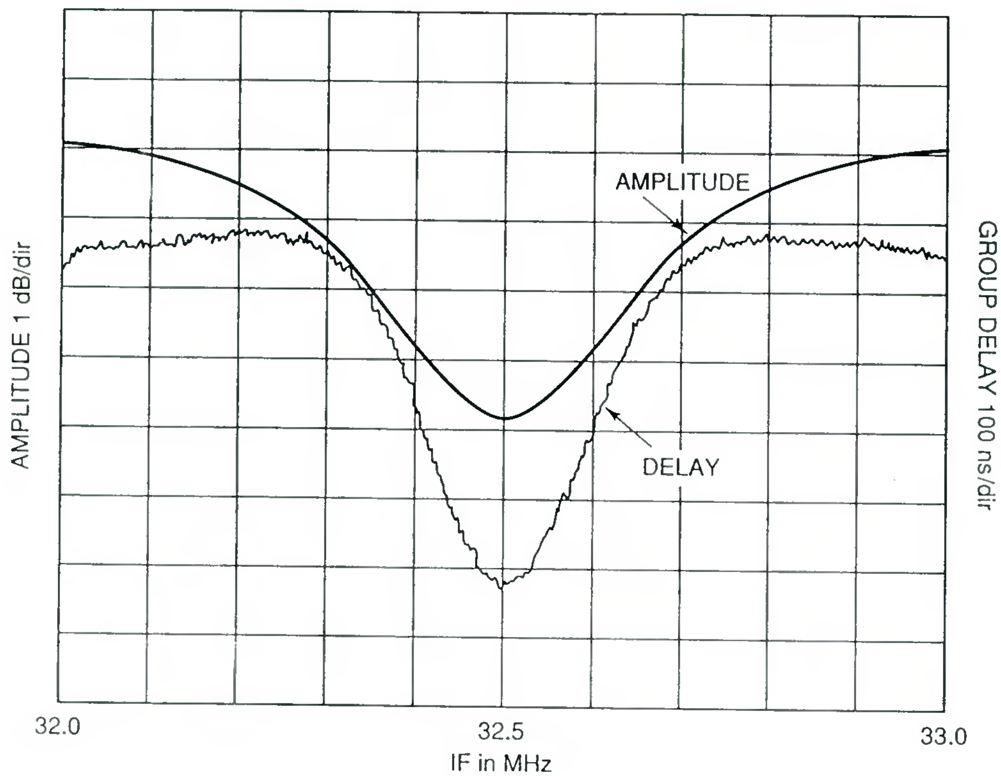


Figure 32. Aural group delay corrector measured results.

shown previously in Fig. 33. Fig. 34 shows that more than 10 dB stereo separation improvement can be obtained over midband audio frequencies.

SOLID-STATE TRANSMITTERS

More and more emphasis has been placed on maximizing the on-air time of TV transmitters. Recent technological advances in field effect transistors (FET) have made the development of solid state high power, linear amplifiers for TV applications both practical and cost effective.

Designing For Reliability

Because there are many factors which can affect the reliability of a TV transmitter, it must be understood that the application of solid-state high power technology in itself does not necessarily ensure a high reliability, low maintenance transmitter. Overall design philosophy, device technology, module design, control architecture, power supplies, cooling systems and cabinet design are some of the critical areas which also must be considered. By paying careful attention to the overall architecture and all of the subassemblies in the entire transmitter, a very reliable system results.

In a transmitter design which uses circuits in series with no system redundancy, it is clear that if one device fails then the entire transmitter will fail. Fig. 35 shows a system of three series devices with no redundancy.

If each device (a, b, c) has a probability of survival over a given period of time (P) of 0.5, then the overall probability of the system surviving P(s) over the same period of time is given by the formula:

$$\begin{aligned} P(s) &= P(a) \times P(b) \times P(c) \\ &= 0.5 \times 0.5 \times 0.5 \\ &= 0.125 \end{aligned}$$

If three identical devices are operated in parallel and only one is required for adequate operation of the system, the probability of the system surviving over the same time period is greatly enhanced. Fig. 36 shows this configuration.

The overall system reliability now becomes:

$$\begin{aligned} P(s) &= P(a) + P(b) + P(c) - P(a)P(b) - \\ &\quad P(a)P(c) - P(b)P(c) + P(a)P(b)P(c) \\ &= 1.5 - 0.25 - 0.25 - 0.25 + 0.125 \\ &= 0.875 \end{aligned}$$

System On-Air Availability

Related to reliability, but perhaps even more important to the broadcaster, is on-air availability. On-air availability is the percentage of time the transmitter

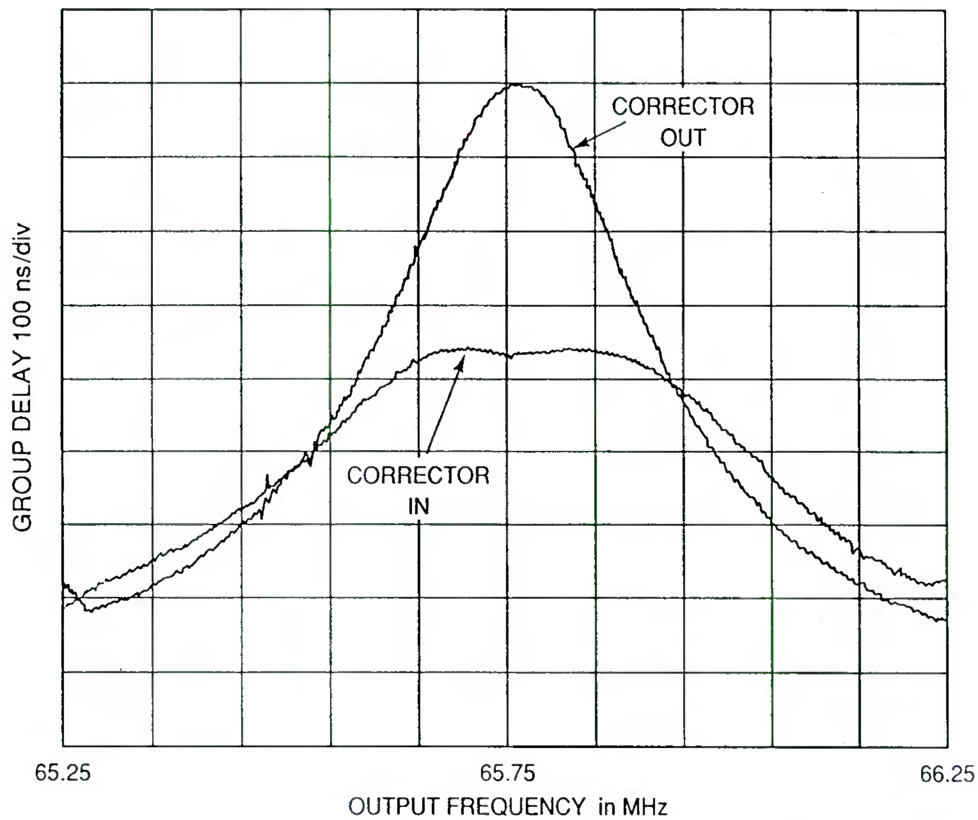


Figure 33. Overall transmitter delay with aural delay corrector switched in and out.

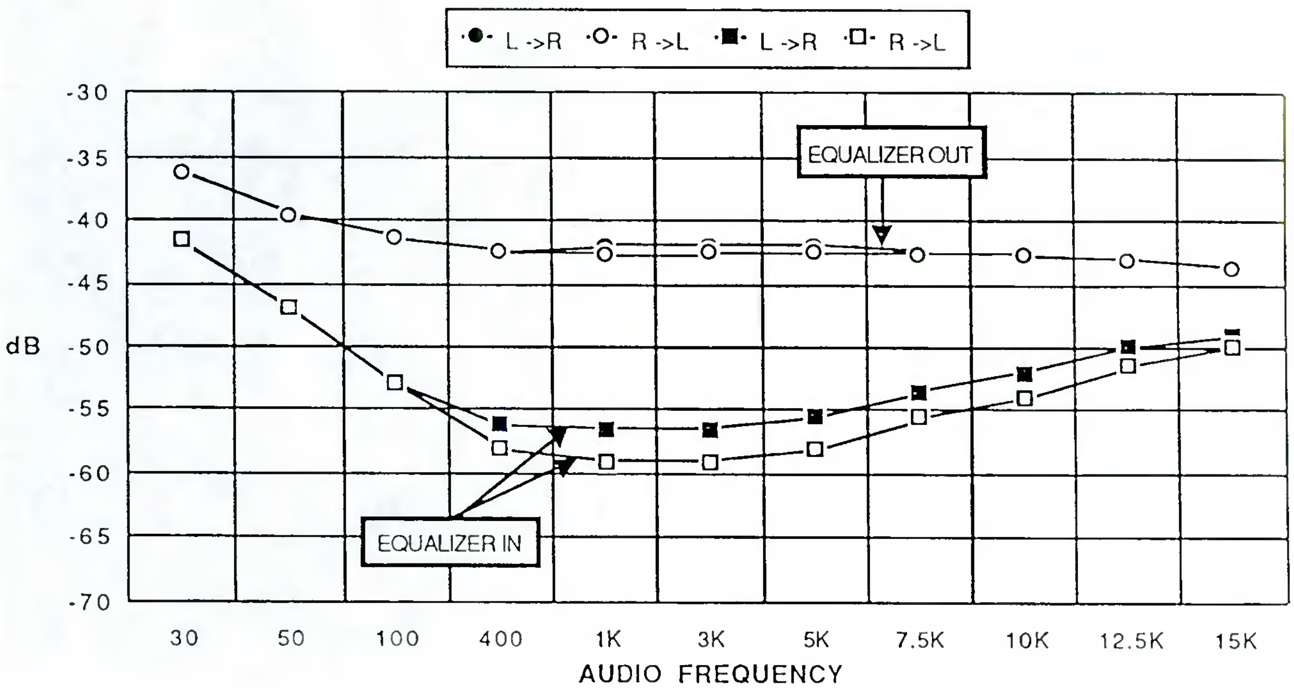


Figure 34. TV stereo separation with and without group delay equalizer.

Figure 35. Circuits in series.

is in service, or could be in service. Availability is defined by the following equation:

$$\text{On-Air Availability} = \frac{\text{MTBF}}{\text{MTBF} + \text{MTTR} + \text{MPMT}} \times 100\%$$

where:
MTBF = Mean Time Between Failures (hours)
MTTR = Mean Time To Repair (hours)
MPMT = Mean Preventative Maintenance Time (hours)

From this equation, one can see that there is little point in designing a transmitter that has an extremely high MTBF figure if, due to poor design and mechanical packaging, it takes an inordinate length of time to make repairs, or the transmitter has to be shut down frequently for routine preventative maintenance.

Many stations have very short sign-off windows or operate 24 hours a day. This often results in a less-than-optimum maintenance schedule which can lead to premature failure or out-of-tolerance operation. One way to reduce the amount of off-air maintenance time is by making provisions for on-air maintenance or to have redundant transmissions. This significantly reduces the MPMT factor.

Several design factors should be considered for a transmitter for optimum on-air availability:

1. Very high reliability for the fundamental circuits.
2. Provision for fast and easy access to all modules and subassemblies.

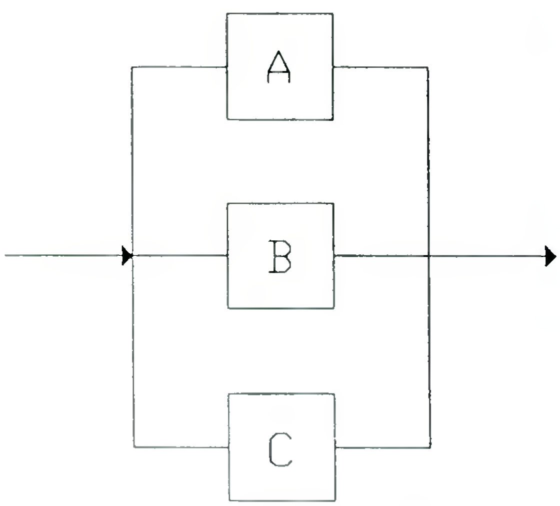


Figure 36. Circuits in parallel.

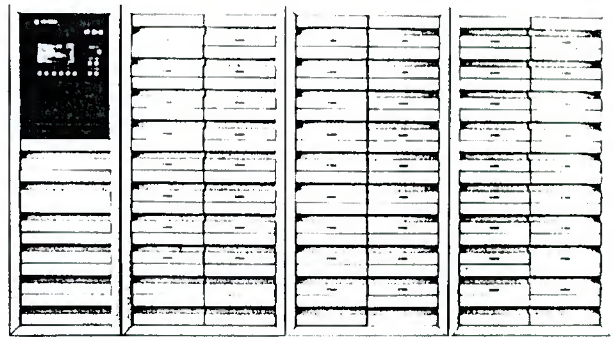


Figure 37. Modular VHF-TV transmitter (Model HT-30HS). (Photo courtesy of Harris corporation)

3. Maximum use of like part and subassemblies. Because only a few items would be needed, this allows most TV stations to maintain a full inventory of spares. If spares are on hand, it follows that the repair time will be much shorter.
4. Repair of transmitter at module or subassembly level. Modules which have been removed can then be repaired by station personnel or returned to the manufacturer for exchange.

Figs. 37 and 38 show two different modular solid state VHF-TV transmitters.

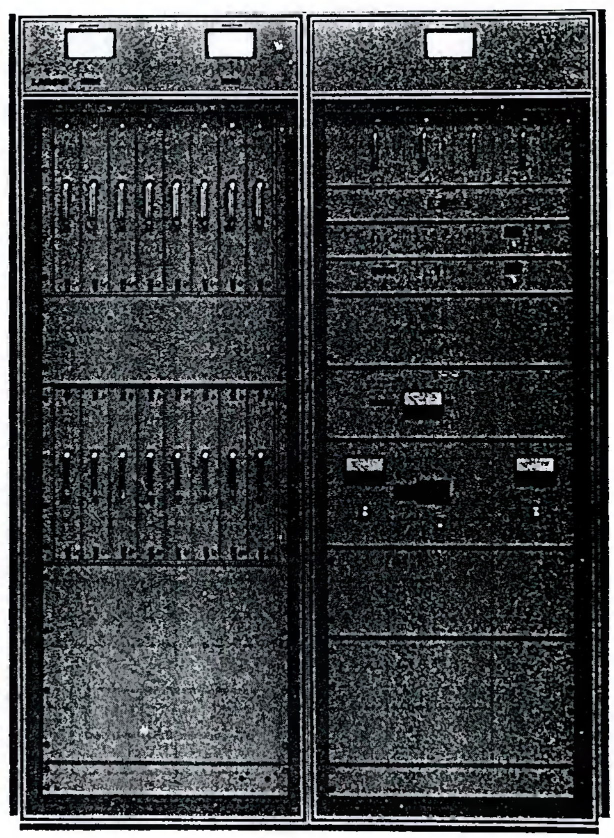


Figure 38. Modular VHF-TV transmitter (Model TTS-12M). (Photo courtesy of Larcan)

By combining a few or many RF modules, it is practical to create transmitters at any power range up to 60 kW. Solid state transmitters maintain their performance over extended periods of time due primarily to the fact that they have no tuning controls nor filament emission degradation with time. Another significant advantage is that no warm-up time is required. Solid-state transmitters can be producing full rated power within seconds of activation. Since solid state transmitters are of a different structure than tube transmitters, it is worthwhile to review some of the new transmitter architecture.

AGC

An automatic gain control (AGC) system is used to maintain constant power output from the transmitter. Ambient temperature changes will cause gain changes in a solid state amplifier or a faulty RF power module will cause a reduction in final output power. Then, RF drive power must be boosted to maintain constant power output. In most cases a detected RF sample of the PA output is fed to an input of a comparator. The exciter output sample or a voltage proportional to the exciter output is applied to the other input of a comparator. The DC output is then integrated and fed to an attenuator which varies the RF drive level at a low power level.

Combining Multiple Amplifier Cabinets

When combining RF power amplifiers, they must be matched in phase and gain for maximum power to the antenna and minimum power to the reject load. Electronic phase shifters and attenuators must have the capability for remembering their settings in case of AC power failure.

RF Amplifiers

Combining several RF power modules to achieve the desired transmitter output power increases the parallel redundancy and the on-air availability described earlier. Choosing the optimum power level for the PA modules of a solid-state transmitter is not an easy task. This point is illustrated in Fig. 39.

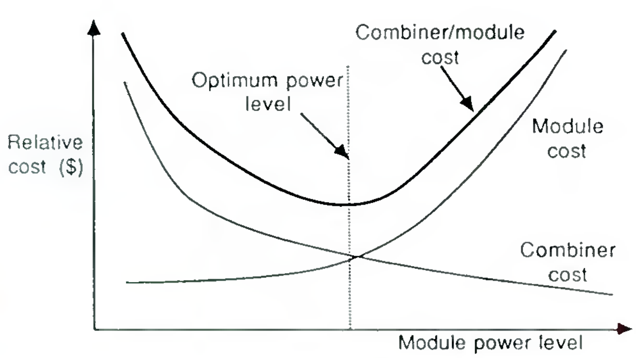


Figure 39. Module cost versus module power level.

A 1 kW output power module has been adopted in nearly all manufacturers' PA modules based on overall cost, practical weight, and size limitations.

Self-protection of each PA module against various fault conditions is good engineering practice. It is desirable that one sub-assembly failure not cause another sub-assembly failure. By using self-protecting modules, the cabinet control logic and overall transmitter control logic can be kept very simple, thus improving overall reliability. Self-diagnostics for the module aid in minimizing the time to repair. Protection from over-voltage, RF overdrive, VSWR, overtemperature, and ensuring proper load sharing among devices is essential to maintaining amplifiers for long life.

Modular amplifiers, which can be removed while "hot" can help in improving overall on-air availability. If an amplifier module fails, the transmitter can continue to function properly indefinitely without disrupting transmitter operation. If a spare PA amplifier is kept at the station, it can be used while the failed unit is repaired or returned to the manufacturer.

Temperature compensated, regulated bias supplies for the amplifiers are essential. Otherwise, performance and power output would vary roughly in direct proportion to temperature or supply voltage changes.

Amplifier level faults are most easily verified by swapping modules in different slots. If the fault follows the module it is an internal module problem but if the fault remains at the same slot after substituting the module, then the problem may be somewhere else in the system.

Solid-State Devices

Both bipolar and field effect transistor (FET) technology exist today as suitable RF amplification devices. Power amps are operated in class AB for the best trade-off of efficiency and linearity. The driver amplifiers usually contain class A operated RF amplifiers. Although both types have their merits, FETs have some advantages over bipolar devices.

FETs have a higher amplification factor than bipolar transistors, helping to reduce the number of devices required as driver stages. Higher supply voltages help to reduce the current capacity of the power supply. Simpler bias circuitry minimizes parts count.

Cooling System

Proper cooling of the solid-state modules is extremely important in obtaining a high MTBF. The MTBF of a FET or transistor essentially doubles for every 10°C drop in the actual FET junction temperature.

Distributed cooling systems employing more than one fan offer good redundancy. Current motor/fan technology has matured to the point where a few larger direct drive fans can be equally as reliable as, or more reliable than many smaller fans. Since many RF power amplifiers may be employed, a large volume of air is needed to adequately cool the heatsinks. Low pressure fans or blowers may be used if heatsink fin density is not high. This aids in reducing audible noise generation from the transmitters. The heat is distributed over a large volume of air and thus the temperature rise is relatively low; on the order of five to seven degrees Celsius.

Power Supplies

Power supply design is another area which is critical to the reliability of a solid-state transmitter. Since FET and bipolar devices are low voltage devices, the power supplies which serve them provide low voltage and high current. Therefore, high reliability connections must be guaranteed in the DC distribution. Since available power output from a FET or bipolar transistor varies roughly as

$$P_o = \frac{(Vcc)^2}{2R_L}$$

it is desirable that the supply remain very tightly controlled over incoming AC line variation.

Also, any anomaly present such as voltage or current transients or voltage sags at the AC input to the power supply should be significantly suppressed before reaching the amplifier transistor device. Transmitters should successfully pass the applicable portions of the ANSI/IEEE C62.41 transient testing standard (also referred to as the IEEE-587 standard).

Since the amplifier's current demand will vary with picture level, the power supply output voltage must remain stable from very little load (i.e., white picture) to very large loads (i.e., producing sync output power.)

Efficiency of the power supply is important since the lost power appears as waste heat energy.

Control Systems

If individual amplifier modules and power supplies are self-protecting, control and monitoring functions can be kept simple and straightforward.

One approach for the control system is to use a single controller to control and monitor all the functions of the transmitter.

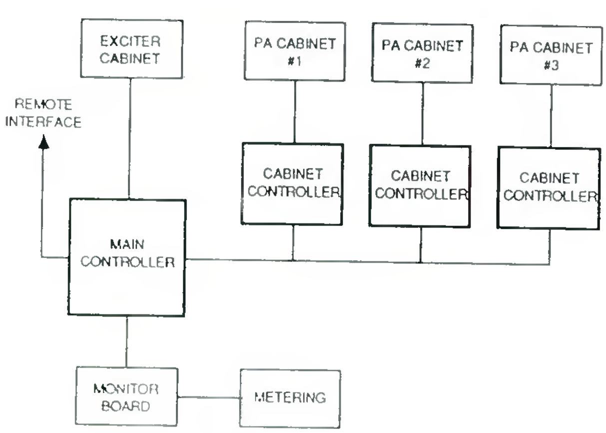


Figure 40. Distributed control and monitoring.

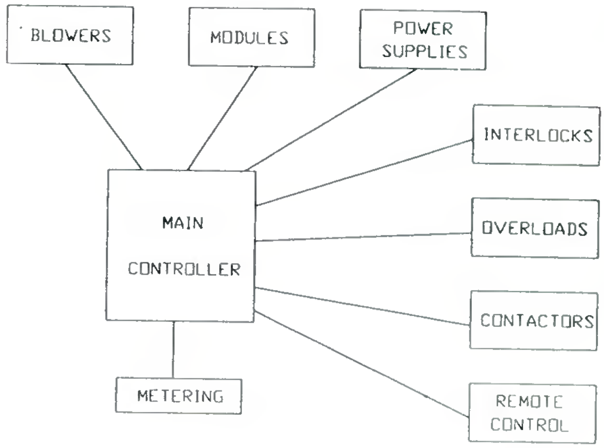


Figure 41. Centralized control system.

Another approach is to distribute the control system throughout the transmitter. The distributed control system can be designed so that the failure of any individual controller does not affect the operation of the others.

Both distributed control system architecture and centralized control systems are shown in Figs. 40 and 41.

After AC power failures, the controller should have back-up memory to restore the transmitter to the same operating condition as before power was interrupted. This would apply to all the control systems within the transmitter.

A system of indicators is essential to quick fault diagnosis. Some typical status conditions which may be displayed are: exciter fault, VSWR fault, VSWR foldback, power supply fault, controller fault, air loss, door open, failsafe interlock, phase loss, module fault, visual drive fault, aural drive fault, and external interlock(s).

VSWR foldback reduces power during high VSWR operation, such as antenna icing, and restores RF power back to normal when difficulties are removed. This is a technique used to keep transmitters on the air. Other options used to enhance the on-air capability of solid state transmitters may include dual exciters, 20% aural power, and redundant drive chains. An example of a VSWR foldback block diagram is shown in Fig. 42.

AC Distribution

Tube-type transmitters typically have one AC service connection to the transmitter. A more reliable method is to provide power to the modular RF amplifier cabinets through a distributed AC power feed system in which each cabinet is protected by a separate AC breaker that is external to the transmitter, as shown in Fig. 43. This concept also allows a cabinet to be safely serviced while the remaining cabinets are operational. Phase monitors guard against low voltage, loss of one phase, or reversal of the phase sequencing.

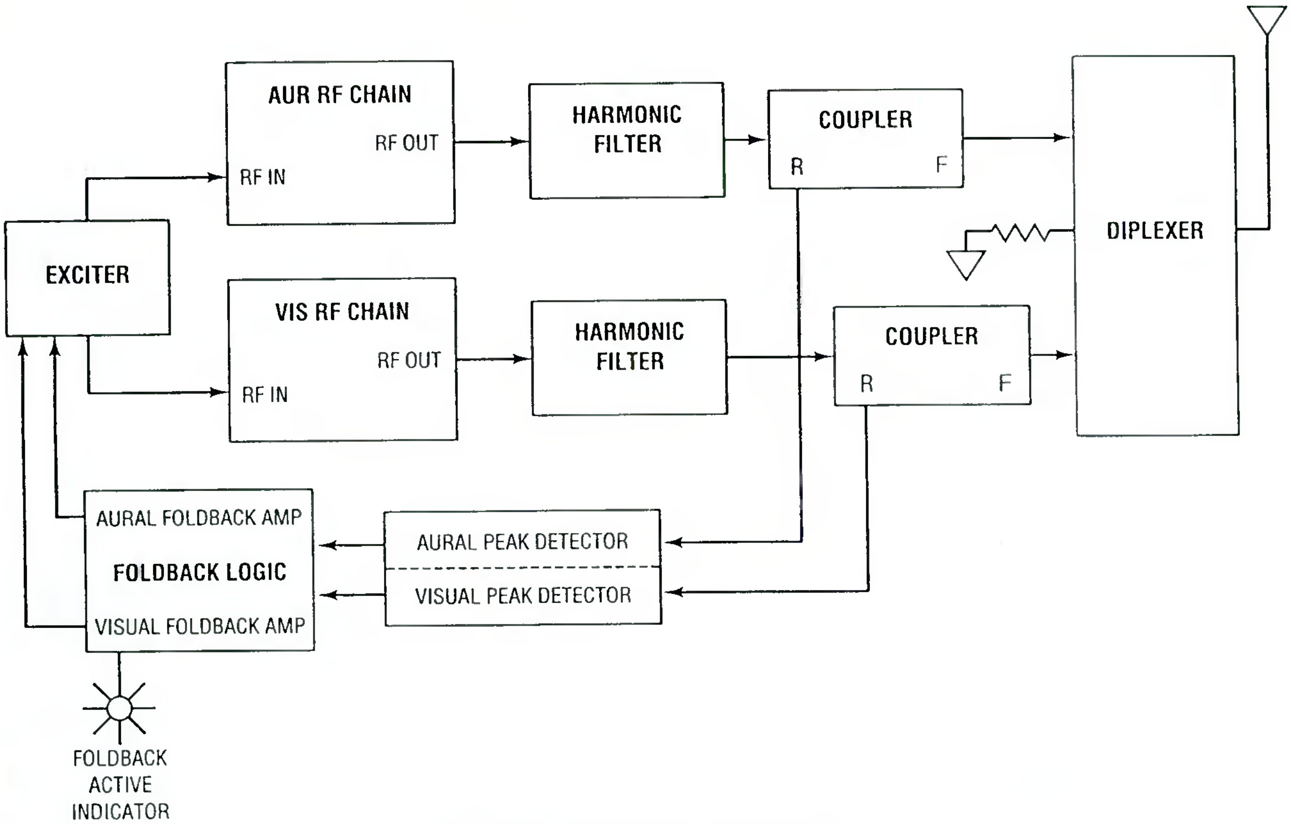


Figure 42. VSWR foldback block diagram.

Combiners and Dividers

There are several choices as to the method used for dividing and combining RF power for the solid-state visual and sound amplifier modules. The most predominant method used has been with in-phase N-way ring combiners. Two common examples are described below.

Microstrip Wilkinson Combiner

Fig. 44 shows an example of this type of combiner. Microstrip is used as the transmission line to carry the RF power. When all amplifiers are operating, equal voltages are presented to each side of the load resistor so that no power is dissipated. When an amplifier failure occurs, the power is distributed between the loads and the output. In this type of combiner, the impedance of the transmission lines and the length of the lines may be varied to achieve the desired impedance transformation.

Balanced reject loads are used to absorb RF power in case of an amplifier failure and to provide isolation for the other amplifiers.

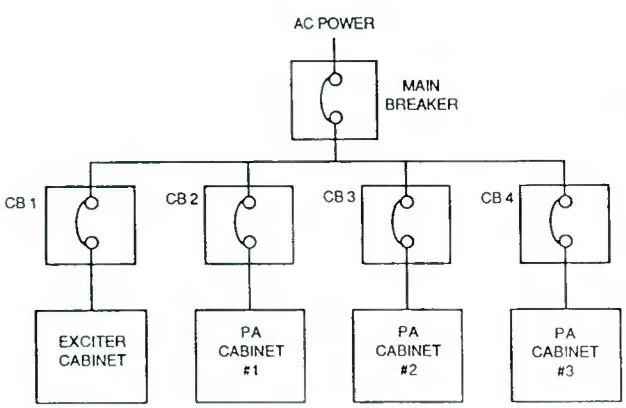


Figure 43. Distributed AC power system.

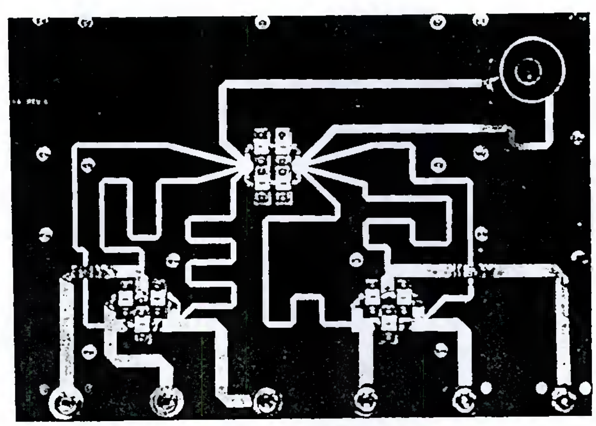


Figure 44. Microstrip Wilkinson combiner.

The combiner is housed in a shielded casing so that outside fields cannot alter the balanced configuration. In case of an amplifier failure, the reject loads conduct heat through the flange to the heatsink where the heat is exchanged to the moving air stream. This type of combiner is used generally for lower numbers of amplifiers (i.e., two to six).

Ring Combiner

Fig. 45 shows an example of this type of combiner. The higher power handling capability of the coax lines used in this combiner will also allow a large range of amplifiers to be combined, (i.e., two to 20). It also provides isolation from one amplifier to another using reject loads which are not in the direct RF path to the output. The operation of the multiport combiner is easily understood by first understanding a two-way combiner.

Refer to Fig. 46 for a simplified version of the combiner. Each of the transmission lines is a quarter wavelength. When equal voltages are applied to both input ports (both amplifiers operating) the combined signal arrives at the output. This is true for three reasons: (1) the distance from each input port to the output is electrically equal whether the signal follows the shorter path or the longer path, (2) the signal from one amplifier arrives at the load port out of phase with that from the other amplifiers so no power is dissipated, and (3) the signals from amplifier 1 arriving at input port 2 via the short and long paths are out of phase and vice versa. *Thus under normal conditions, all of the power appears at the output, none is absorbed in the loads, and there is complete isolation between amplifiers.*

Power is absorbed in the load resistors only when an amplifier is not operating. Assume that only amplifier number 1 is operating. The signal path is electrically equal not only for the long and short paths to the output but also for the right and left paths to either load resistor. The power from the operating amplifier

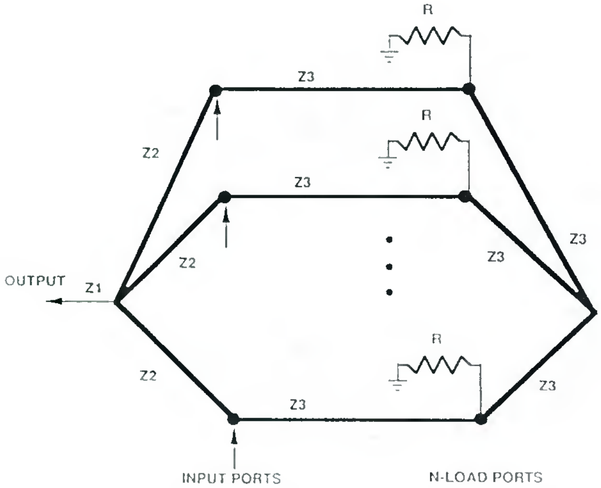


Figure 45. N-way ring combiner.

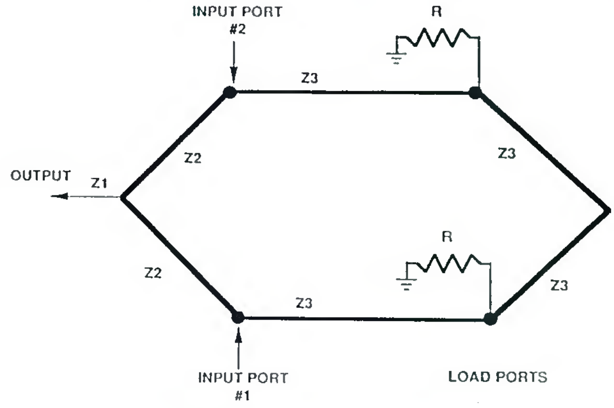


Figure 46. Two way ring combiner.

is split equally between the output and the isolation loads. Due to the isolation inherent in the network, the input ports remain matched even when one or more amplifiers are removed.

Since the transmission line that connects the amplifier port to the output port is a quarter wavelength long, and 75 ohm transmission line is used, the 50 ohm amplifier impedance is transformed to: $(75/50) \times 75 = 112.5$ ohms. For combining N amplifiers, the impedance at the output combiner junction is 112.5/N. This impedance is then matched to 50 ohms.

In this type of combiner, the reject loads may also be mounted to a grounded structure. Removing the heat from this structure should be done as quickly and efficiently as possible. One method to accomplish this is to use a heat pipe for the mounting structure. In case of an amplifier failure or removal, the reject load temperature rises, the fluid in the lower section of the heatpipe heats up until it vaporizes. The vapor rises to the finned area where the heat is exchanged to the moving air stream. The vapor condenses as it releases its heat and returns to the bottom of the pipe to absorb more heat.

VHF TUBE TRANSMITTERS

Triodes and tetrodes are still used in the RF power stages of television transmitters. This section will highlight some of the facts on high power gridded transmitter tubes. The parts of the power tube will be reviewed.

General

A vacuum tube uses grids to control the flow of current through the tube. The cathode emits the electrons which travel to the plate. The plate is more positive than the cathode. The grids control this flow of electrons, thereby controlling the plate current. They also can modulate the electron stream causing the plate current to have the same waveform as the grid voltage, depending on class of operation.

A triode is a three element tube consisting of:

1. A heated cathode which emits electrons.
2. A control grid which modulates the electron stream in accordance with the DC and AC voltages impressed between the control grid and cathode.
3. A plate which accepts the electron stream. The plate is positive with respect to the cathode.

Input signals, DC and AC, are applied between control grid and cathode. Output signals, DC and AC, flow to the tube output load impedance by way of the plate and cathode.

A *tetrode* is a four-element tube which has a *screen grid* added between the control grid and plate. The screen grid acts as an electrostatic shield which helps to isolate the plate output signal from the control grid input signal, and is operated at a positive DC potential with respect to the cathode. This potential is much lower than the DC plate voltage and is usually operated at AC (RF) ground potential.

The tetrode has higher gain than the triode. The plate voltage has minimum effect on the plate current as long as the instantaneous plate voltage is greater than the screen grid DC voltage. The plate voltage is normally allowed to swing close to, but not below, the value of DC screen grid voltage. If the tetrode screen grid is not maintained at AC (RF) ground potential, the tube can take on the characteristics of a triode. The tetrode input bias and signal are applied between the control grid and cathode and are similar in performance to the triode. The output AC (RF) signal current path in the tetrode is from the plate, through the load impedance, and returns through the screen grid.

The Cathode

A power grid tube will have high peak and average current levels. Therefore, the cathode must be hot to emit many electrons. It normally requires large heater power at low voltage and high currents. DC current is often used to reduce AC noise levels. It is usually directly heated (the heater in the cathode) to provide quick warm up with lower heater power. Some type of cooling to remove excess cathode heat from the tube filament/cathode contacts and the tube socket may be required.

The Grids

The voltages (DC, AC, and RF) of all grids are measured with respect to the cathode. Grids can have current flow which is positive when the grid accepts electrons from the electron stream or negative when the grid emits electrons.

Grids can emit electrons and negative grid current can occur because they are located close to the hot cathode. The grids can become very hot and have small amounts of thermionic emission (primary emission) or electrons emitted by being bumped off by other electrons (secondary emission).

As tubes age, some of the cathode electron emitting coating may be deposited on the relatively cooler grids thereby increasing their tendency to emit electrons.

The control grid has a negative DC bias and an AC driving signal superimposed upon it. The instantaneous grid current can be negative when the grid voltage is swinging maximum negative, zero when the grid is only slightly negative, or positive when the grid is positive.

The screen grid has a positive DC bias and is at AC ground potential. The instantaneous screen grid current is negative when the plate voltage swings maximum positive, and positive when the plate voltage swings close to or below the screen grid DC voltage, and zero when the plate voltage is between these extremes.

Both grids have instantaneous positive, zero, or negative currents that follow the voltage swing. The average of these currents is the DC grid current. Since both DC voltage and current are present in both grids, the grids can dissipate power. The control grid also has AC (RF) voltages and currents present so that AC power dissipation in grids must also be considered.

The DC plate power that is applied to the tube is either converted to AC (RF) output power or dissipated as heat. This heat is developed in the plate itself and must be removed by air or water cooling.

Below 30 MHz, RF amplifiers use lumped components to implement the matching circuits for the tube input and output. At VHF and UHF frequencies, several problems make the use of lumped components (L&C in the same purchase) impractical.

The Transmission Line Cavity

As frequency increases, resonant circuits are smaller to reduce inductance and capacitance, larger in diameter to reduce skin effect, and closer to the tube to reduce the effects of stray inductance, and there is difficulty in predicting exactly what values of resistance, inductance, and capacitance a component or circuit may have.

These problems can be managed in low-power circuits, but with high power circuits, arcs and shorts due to high DC and RF voltages become a problem. Larger size and spacing of components are a good start towards arc and short prevention, but this is in opposition to the smaller size and spacing needs dictated by the high frequency operation. Also in high power circuits, the unpredictability of the circuit values of R, L, and C make it difficult to control the vitally important parameters of dissipation, efficiency, and reliability of operation.

One solution to the above problem is the resonant transmission line cavity amplifier. In this type of amplifier the tube becomes part of a resonant transmission line. The elements of these tubes themselves are arranged to look like concentric coaxial transmission lines. The design of these power tubes stresses low inter-electrode capacity and low distributed inductance. The stray inter-electrode and distributed capacity and inductance of the tube becomes part of the resonant transmission line. The resonant transmission line is physically larger than the equivalent lumped constant LRC resonant circuit operating in the same frequency. This larger physical size aids in solving the

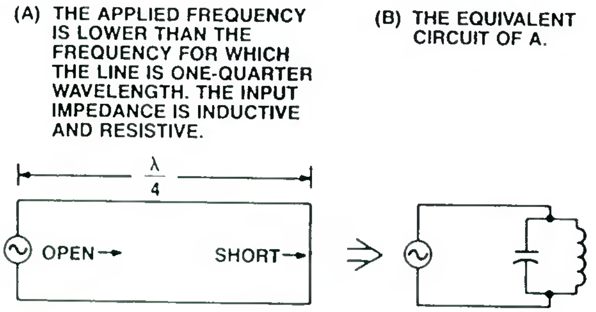


Figure 47. Shorted quarter wavelength line.

high power operation problems of skin effect losses, prevention of arcs and shorts, and yields reliable and predictable operation. A commonly used transmission line cavity amplifier uses a quarter wavelength transmission line as its resonant element.

The Shorted Quarter Wavelength Line

A shorted quarter wavelength transmission line has a very high, purely resistive input impedance. Electrically, it looks like a parallel resonant circuit as shown in Fig. 47).

The Shorted Transmission Line Less Than a Quarter Wavelength Long

When the physical length of the line is less than one quarter wavelength, the impedance will be lower and the line will look inductive. Refer to Fig. 48. This inductance will be used to resonate with capacitive reactance in the tube and surrounding circuit.

Quarter Wavelength Cavity and Amplifier

In Fig. 49, shorted transmission lines are used to resonate the inputs and outputs of this amplifier. Notice that the length of the lines is less than a quarter wavelength but the tube's shunt input and output capacity and its series lead inductance will electrically lengthen and resonate the transmission lines. The input is shown inductively coupled, but it could just as easily have been capacitively coupled to the cathode. The input could also have a lumped constant resonant circuit or a transmission line resonant circuit since its

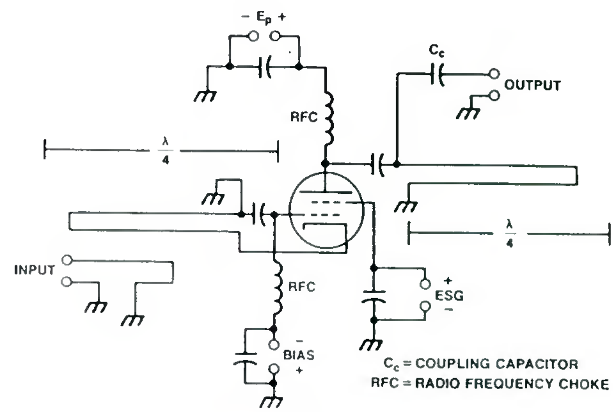


Figure 49. A shorted quarter wavelength transmission line amplifier.

power level is low. The output coupling is capacitive, but it also could have been inductive. The construction of the tube lends itself to transmission line tuning techniques.

Coaxial Construction of a Tetrode RF Power Amplifier Tube

The Anode (Plate)

The plate resembles a copper cup with the plate contact ring welded to the mouth and the cooling fins silver soldered or welded to the outside of the cup as shown in Figs. 50 and 51.

The contact ring is bonded to the base ceramic spacer through a strain isolation ring. This ceramic spacer is the same ceramic that is shown above the screen contact ring shown in Fig. 52.

The Screen Structure

The screen grid consists of many vertical supports fastened to a metal base cone. The other end of the metal base cone fastens to the screen contact ring. The inductance of the individual vertical supports is reduced by building the screen grid of many of vertical

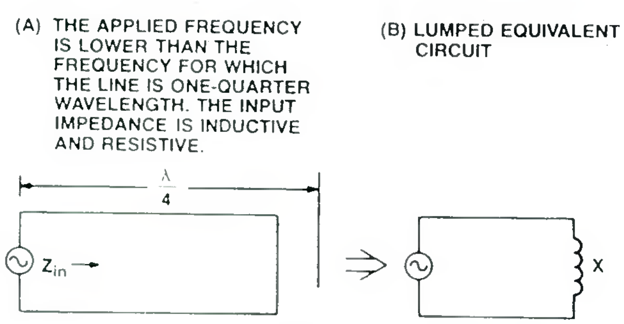


Figure 48. Shorted line less than one-quarter wavelength.

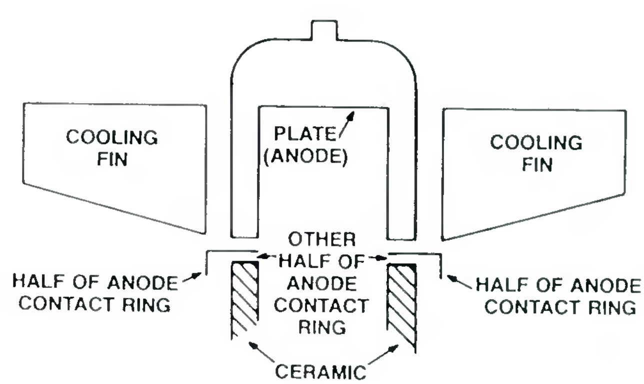


Figure 50. Cutaway view of the anode structure.

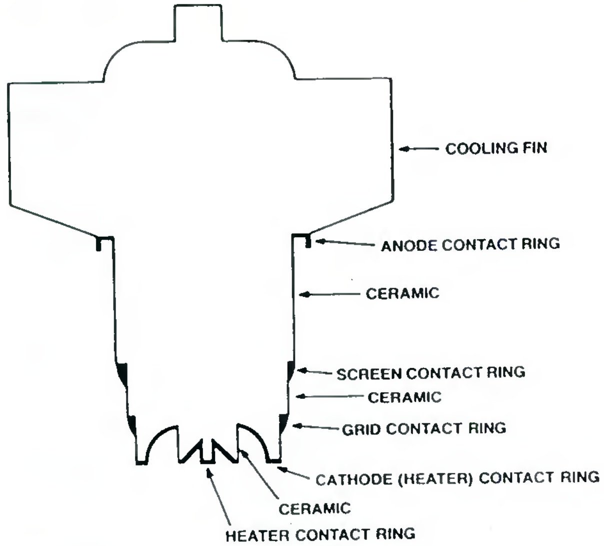


Figure 51. Cutaway view of the exterior of a RF tetrode.

supports in parallel. The vertical supports are held rigid by horizontal rings welded to them and a metal cap on the top of the assembly. The screen contact ring, metal base, and metal base cone also function to reduce lead inductance and RF resistance due to skin effect (refer to Fig. 52).

A cutaway view of the plate circuit and the screen circuit in Fig. 53 shows a concentric construction that resembles a coaxial transmission line.

Consider that the output RF current is generated by an hypothetical current generator between the plate and screen grid. The RF current travels along the inside of the plate structure on its surface (skin effect), through the ceramic at the bottom of the anode contact ring, around the anode contact ring, across the bottom of the fins, and to the band around the outside of the fins. From here it flows through the plate bypass capacitor to the RF tuned circuit and load, and returns to the screen grid. The return current travels through the screen contact ring, up the cone, and up the screen bypass capacitor, then through the screen grid to return to the hypothetical generator. The screen grid has RF

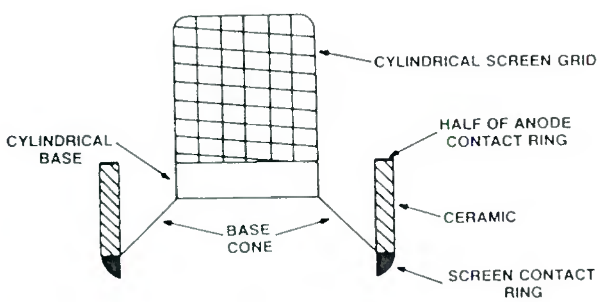


Figure 52. The screen grid assembly.

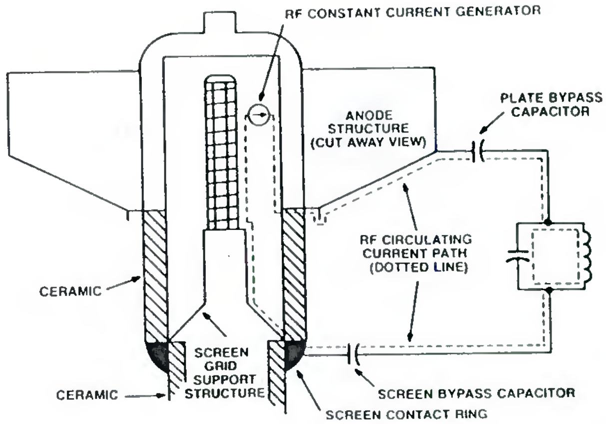


Figure 53. Showing the plate and screen assembly and RF circulating current path (dotted line).

current returning to it but due to its low impedance, the screen grid is at RF ground potential. The RF current generator appears to be feeding an open ended transmission line consisting of the anode (plate) assembly, and the screen assembly. The RF voltage developed by the anode is due to the plate impedance presented to the anode by the resonant circuit and its load.

The control grid assembly and the cathode assembly are also cylindrically constructed and concentric. The control grid assembly is constructed similarly to the screen grid but slightly smaller. Fig. 54 shows the screen grid, control grid, and the cathode assemblies as they are placed in the tube.

Fig. 54 also shows the current path of an RF generator feeding a signal into the grid/cathode assembly. It resembles a transmission line terminated by the RF resistance of the tube's electron stream. The outer contact ring for the cathode heater is the inner conductor of this transmission line.

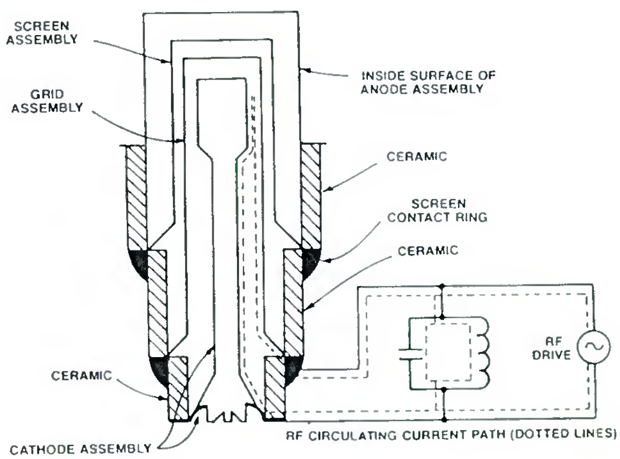


Figure 54. Showing details of assembly of grids and cathode components and the simplified RF input circuit (bias circuit not shown).

Double-Tuned Tube Power Amplifiers For TV Transmitter Applications

Television RF power amplifiers use tetrode tubes because of their high gain, good linearity, good efficiency, and high power advantages. A double-tuned output circuit is used to achieve proper bandwidth and efficiency.

As an example, the double-tuned overcoupled amplifier shown in Fig. 55 should not be thought of as an amplifier and a socket into which a tube is placed. The tube is an integral part of the cavity. The internal electrical properties of the tube will determine the amplifier gain and power handling capabilities. It will also dictate the dimensions and functions of the circuitry.

Effects of Tuning an Overcoupled P.A.

All double-tuned overcoupled visual power amplifiers have four controls to accomplish output tuning.

Plate tune (primary tune) resonates the plate circuit and tends to tilt the response and slide it up and down the bandpass (shown as A on Fig. 56, also Fig. 57 and Fig. 58).

Coupling sets the bandwidth of the PA. Increased coupling increases bandwidth and lowers the PA efficiency. When the coupling is adjusted, it can tilt the response and change the center of the bandpass necessitating the readjustment of the plate tune control (shown as B on Fig. 55, also Fig. 56 and Fig. 58).

Secondary tune resonates the secondary cavity and will tilt the response if adjusted. Generally it will not slide the response up and down the bandpass as will the primary tune (shown as C on Fig. 55, also Fig. 56 and Fig. 59).

Loading (secondary load or output load) determines the value of ripple in the response. Heavier loading (maximum C or minimum L) creates a haystacked response and light loading creates excessive ripple. Adjustment of the loading control usually tilts the response and changes bandwidth. This necessitates readjustment of secondary tune and coupling. In some cavities, primary tune may also have to be readjusted. The coupling control will also effect the value of ripple, but its greatest effect will be on the bandwidth (shown as D on Fig. 55, also Fig. 56 and Fig. 60).

Tuning For Power

Sweep the entire transmitter into the antenna or dummy load at 100% power. If tuning of any part or

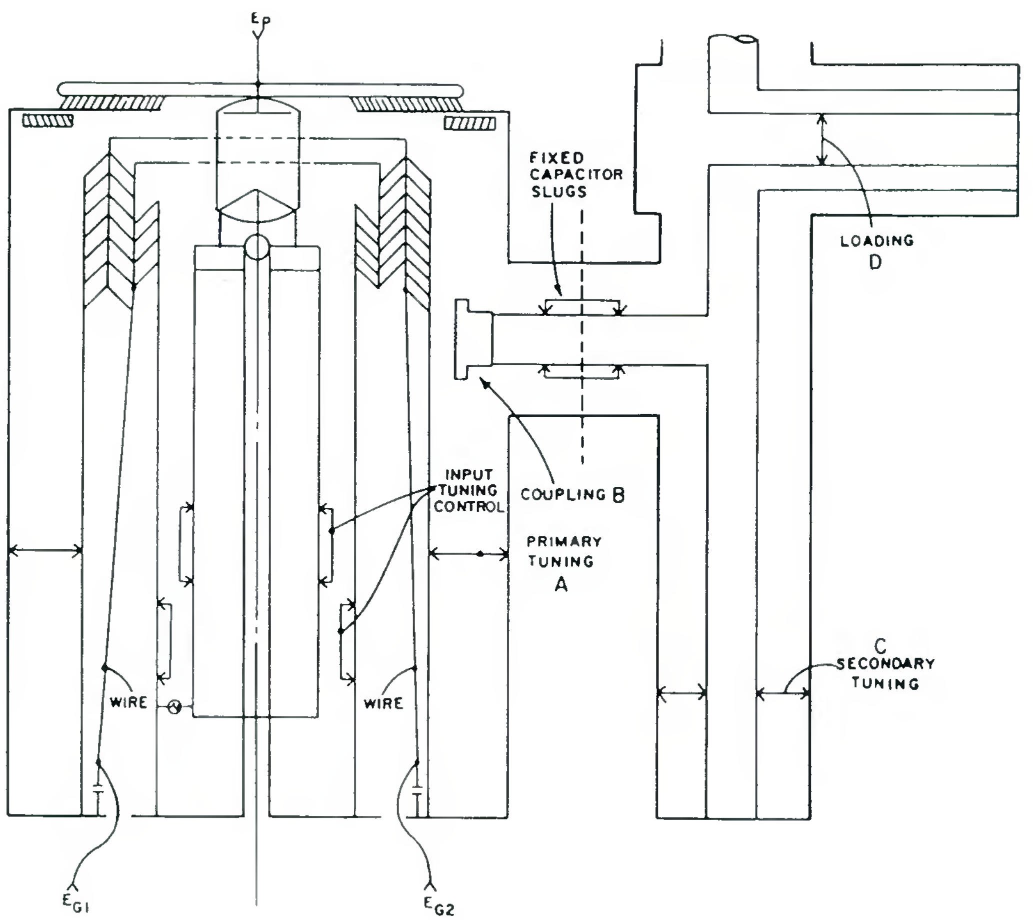


Figure 55. High band coaxial transmission line cavity amplifier for TV use.

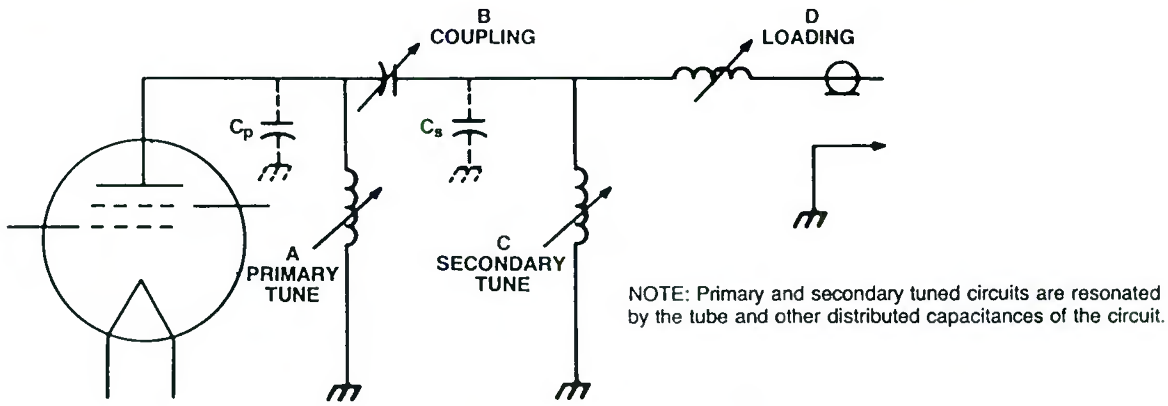


Figure 56. Double-tuned power amplifier.

all of the transmitter is in doubt, check the following. Observe the frequency response, reflected power (VSWR), plate current, screen grid current, and grid current. If the output response is not right, check each stage, starting at the exciter, to be sure each stage is flat and has the correct bandwidth. The final amplifier should be the narrowest of all stages.

After the transmitter sweep is completed and the overall response has the proper bandwidth, transmit a black picture, (video set at blanking, 75% modulation), operate into the station load and check for:

1. Proper sync level at the transmitter output. Adjust the visual exciter linearity corrector as necessary.
2. Excessive plate dissipation.

Plate dissipation = (Plate voltage × Plate current) − (Average RF output power)

3. Excessive screen current. The screen current will increase rapidly at higher power if the PA is lightly loaded. Very light loading can cause sync compression.
4. Excessive grid current, if PA is loaded too heavy, can cause sync compression.
5. Amplifier output efficiency. It can also be a guide to proper operation of the stage. The efficiency at

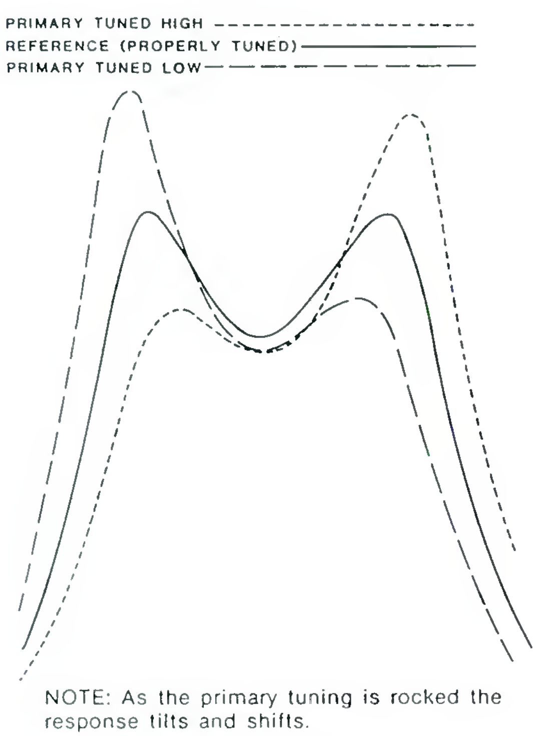


Figure 57. Primary tuning.

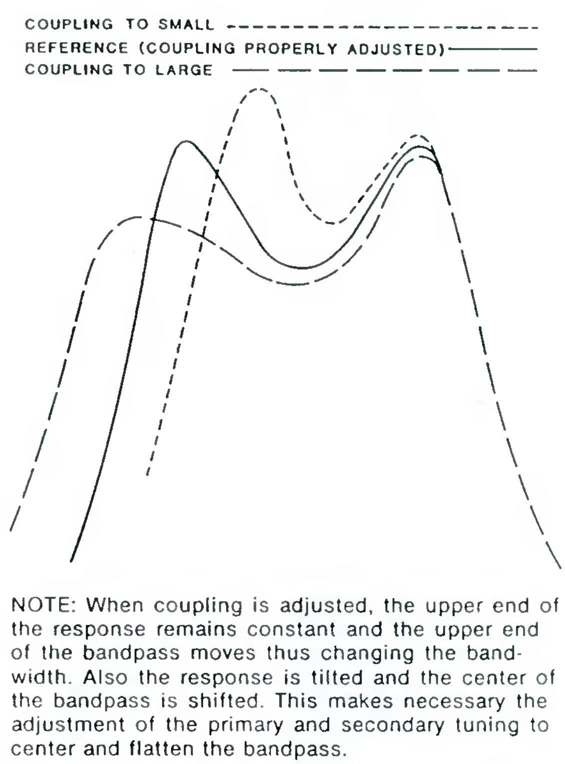
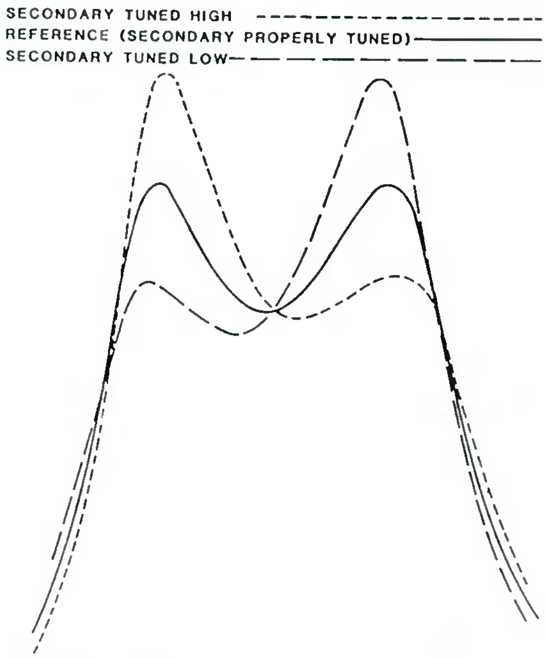


Figure 58. Coupling plate loading.



NOTE: As secondary tuning is rocked, the response tilts, but it tilts in the opposite direction to the primary. Thus, the secondary and primary tuning can be changed together to shift the response up or down in frequency.

Figure 59. Secondary tuning.

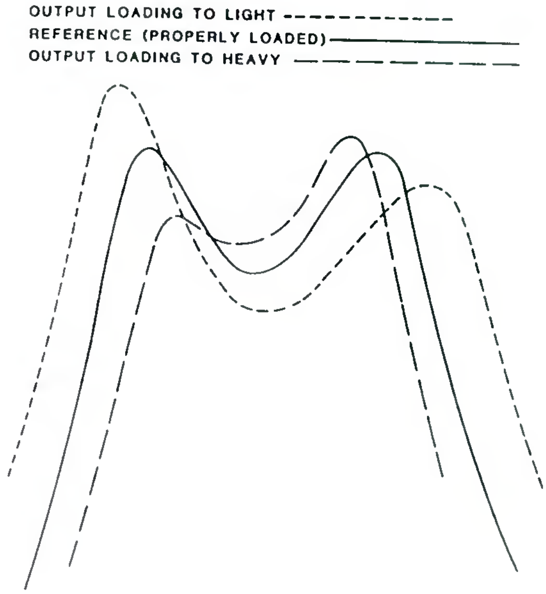


Figure 60. Loading adjustments.

VHF typically will range from 41% to 45%. At UHF the range is somewhat lower.

$$\text{Efficiency} = \frac{\text{Average power output}}{(\text{Plate voltage} \times \text{Plate current})}$$

Heavy Loading	Smaller Ripple	Lower Plate Impedance	Higher I_p and Positive I_g Lower Gain	Lower Positive I_{sg}
Light Loading	Larger Ripple	Higher Plate Impedance	Lower I_p and Positive I_g Higher Gain	Higher Positive I_{sg}

It is assumed that the RF output power is measured by an accurately calibrated RF wattmeter or by a calorimeter. If the wattmeter is not accurate, the amplifier dissipation and efficiency may be in doubt and the tube life may be shortened. If the wattmeter reads low, 100% output power will appear difficult to obtain. The amplifier might be trying to produce more than its full rated power. The symptoms would be high plate dissipation, and sync compression. These are the same symptoms that improper tuning could yield.

If the wattmeter reads high, power will appear easy to obtain and efficiency will appear high. Plate dissipation might appear low and in reality be high. This will become apparent over a long time by a blackened anode.

A calorimeter is the most accurate way of measuring power but is not always practical. If the wattmeter is to be calibrated, send the wattmeter transmission line section, the sampling slugs, and the meter (with the meter cable) together with information giving your frequency and power to the manufacturer for calibration.

The power measured on a calorimeter or on the wattmeter is average power. To find peak power (assuming a black picture with no setup) use this formula:

$$\text{Peak power} = \frac{\text{Average power}}{0.595}$$

If a large value of plate current is required to make power, it indicates that the plate voltage is not swinging very far.

The low plate voltage swing is also indicated by the low screen current. Remember that positive screen current flows only during the time that the plate voltage swings close to the screen voltage.

The low swing of plate voltage along with the high swing of plate current indicates a low plate impedance (heavy loading).

$$\text{Plate impedance } (Z_p) = \frac{\text{The swing of plate voltage } (\delta e_p)}{\text{The swing of plate current } (\delta i_p)}$$

To make the amplifier more efficient and bring plate dissipation down, the plate impedance must be increased. To increase the plate impedance, the amplifier loading must be decreased by performing the following corrections:

Solid line —Transmitter overall response
Dotted line—The part of PA response outside of the transmitters bandpass.

Figure 61. Nonsymmetrical response caused by the amplifiers outside the transmitters bandpass. Solid line is transmitter overall response. Dotted line is the part of PA response outside of the transmitters bandpass.

1. Sweep the transmitter and decrease the PA loading.
2. This will cause the response to tilt. Correct this tilt by adjusting the secondary tuning.
3. Decreasing the loading will also cause the bandwidth (and ripple) to increase. This can be counteracted by decreasing coupling.
4. Changing coupling will cause the response to tilt. This can be corrected by adjusting the primary and/or secondary tuning.
5. The above procedure may have caused PA response to slide up or down out of the bandpass. It will show up as an asymmetrical bandpass (as shown in Fig. 61). It can be corrected by adjusting both primary and secondary tune simultaneously to center the response.

The transmitter's overall response will have the same bandpass but will now have slightly more ripple. The ripple content in the transmitter's overall response should still be within the 0.25 dB to 1.25 dB limits.

CAUTION: When changing loading, coupling, and primary and secondary tuning, it is possible to get the PA bandwidth too wide. The excessive bandwidth of the PA may be masked by narrow driver response. This excessive bandwidth will cause PA overdissipation. Correct bandwidth is shown in Fig. 62 and improper bandwidth is shown in Fig. 63.

Once again, transmit a black picture into the dummy station load. Switch the vestigial sideband filter and the linearity corrector in and check sync level at the transmitter's output. Plate current and grid current should be lower and screen grid current should be higher. The amplifier efficiency should fall between 41% to 45%.

This is a general tuning procedure given to illustrate tuning methods, control interactions and tube operating characteristics. It will work well with most double-tuned overcoupled power amplifiers. For specific tuning information on a given transmitter, the manufacturer's instructions should be consulted.

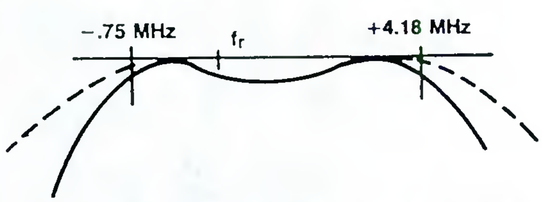
UHF TRANSMITTERS

There has been a significant amount of new technology introduced recently regarding UHF amplifiers used in TV transmitters. Most of this has been concentrated in enhancing the efficiency of the UHF amplifiers.

The introduction of the multiple depressed collector klystron and klystrode® to UHF TV transmitters has dramatically reduced transmitter power consumption. Those new technologies will be addressed in this section. However, a basic background is helpful to the understanding of the new devices.

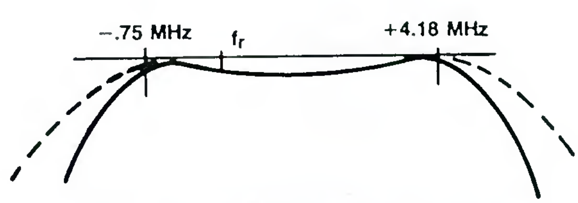
Basic Klystron Theory and Practice

The klystron uses velocity modulation to serve as an amplifying device. The electron beam emitted from the cathode is accelerated to high velocity by the



NOTE: The driver bandwidth is the dotted line, and the PA bandwidth is the solid line. The solid line also represents the overall transmitter bandpass.

Figure 62. Proper bandwidth.



NOTE: The PA bandwidth is the dotted line, and the driver bandwidth is the solid line. The solid line also represents the overall transmitter bandpass.

Figure 63. Improper PA bandwidth.

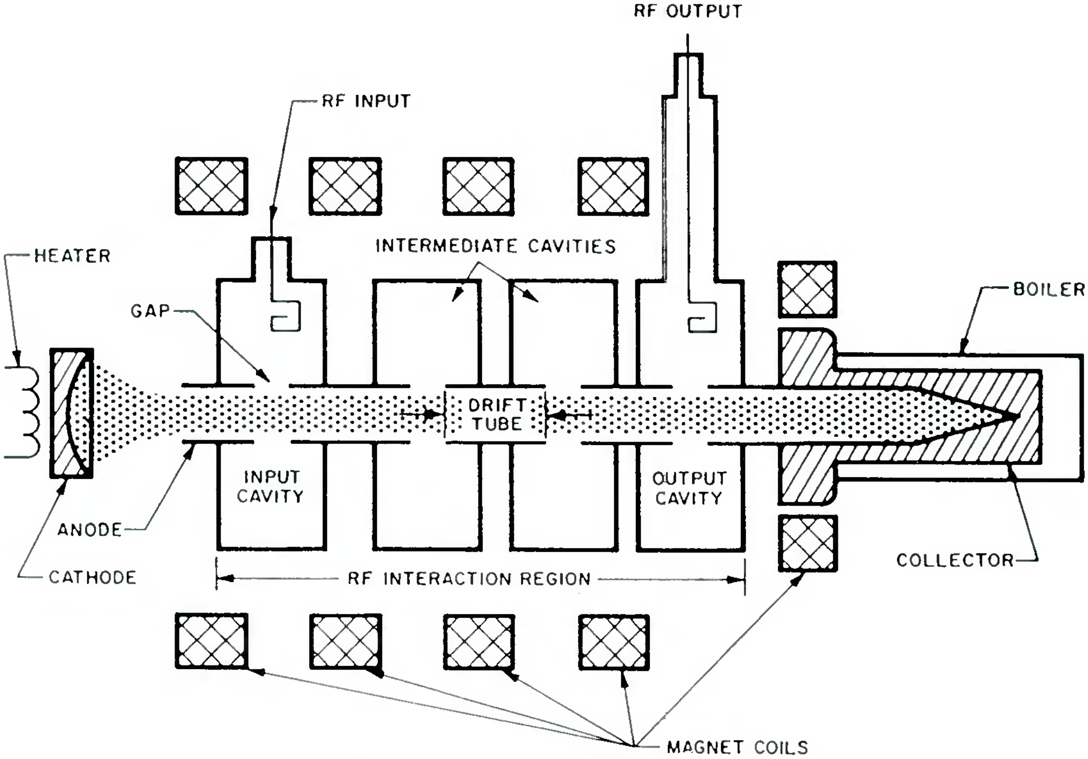


Figure 64. Principal elements of a klystron.

electric field between the cathode and anode and is directed into the RF interaction region, as shown in Fig. 64. An external magnetic field is employed to prevent the beam from spreading as it passes through the tube. At the other end of the tube, the electron beam impinges on the collector electrode, which dissipates the beam energy and returns the electron current to the beam power supply.

The RF interaction region, where the amplification occurs, contains resonant cavities and field-free drift spaces. The first resonant cavity encountered by an electron in the beam (the input cavity) is excited by the UHF signal to be amplified, and a RF voltage is developed across the gap. Since electrons approach the input-cavity gap with equal velocities and emerge with different velocities, the electron beam is said to be *velocity modulated*. As the electrons travel down the drift tube, bunching develops, and thus the density of electrons passing a given point varies cyclically with time.

The RF energy produced by this interaction with the beam is extracted from the beam and fed into a coaxial or waveguide transmission line by means of a coupling loop in the output cavity. The DC beam input power not converted to RF energy is dissipated in the collector.

The cavities can be mounted external to the klystron as shown in Fig. 65 or can be included in the vacuum envelope as shown in Fig. 66.

All cathodes have optimum ranges of operating temperature. The operating temperature of the cathode

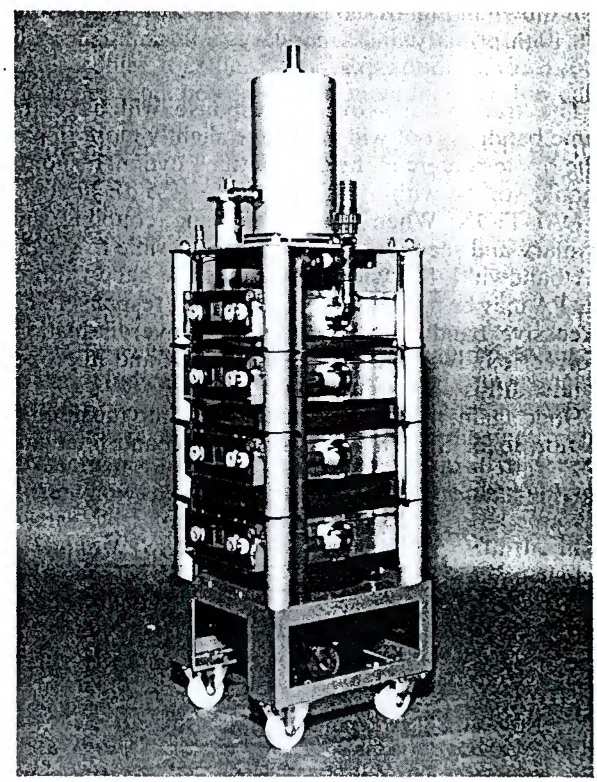


Figure 65. External cavity klystron.

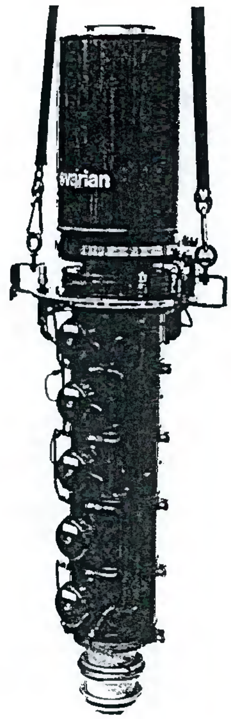


Figure 66. Integral cavity klystron.

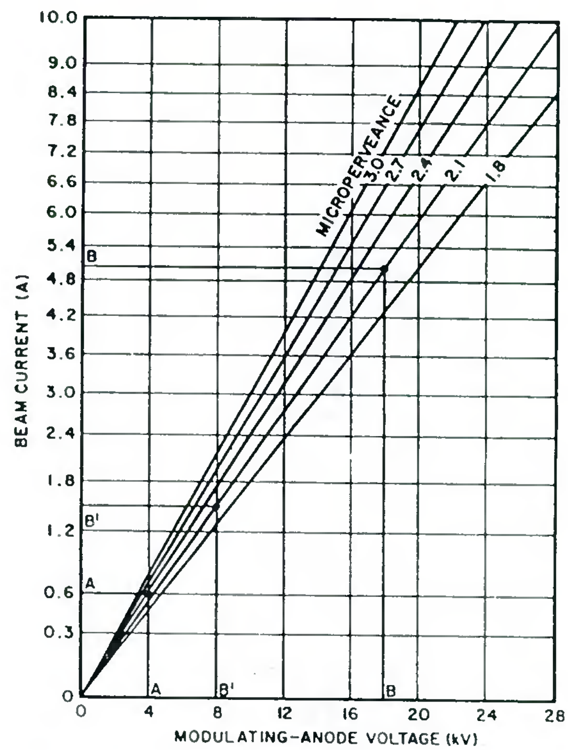


Figure 67. Beam current variation with modulating-anode voltage.

must be high enough to prevent variations in heater power from affecting the electron emission current (beam current) in the klystron. However, the temperature of the emitting surface must not be higher than necessary, since excessive temperature can reduce cathode life.

Perveance and the Modulating Anode

Perveance is a function of the geometry of the cathode-anode structure. In klystrons manufactured today, there are two electrodes which may control the beam current: The modulating anode, and a lower voltage (0 to 1,400 volts) electrode typically used for pulsing the beam current. (Examples are: Beam Control Device (BCD), Annular Control Electrode (ACE), and Annular Beam Control (ABC). If the low voltage electrode is connected to the cathode, the modulating anode voltage controls beam current which can be calculated using the following equation:

$$I_b = K \times E^{3/2}$$

where: K = Perveance constant of the klystron
I_b = Beam current in amperes
E = Beam voltage

Fig. 67 shows the relationship between beam current and voltage described in the above equation. Two examples for using the graph are given. In example A, if a modulating anode of 4,000 volts with respect to the cathode beam voltage produces a beam current of 0.6 amp, the intersection point lies on the 2.4 microperveance line. The perveance is expressed as 2.4×10^{-6} or 2.4 micropervs. Operating condition B illustrates a practical television transmitter situation in which a common beam supply of 18 kilovolts is used to power both the visual and aural klystron. At 18 kV, the visual tube operates at a beam current of approximately 5.0 amperes if the modulating anode is connected (through an isolating resistance) to the body of the tube, and the perveance is 2.1 micropervs. Since the aural output power required is much less, the DC input power is less than that required to operate the visual tube. Points B' indicate that if the modulating anode is supplied with only 8 kV (through a voltage divider) then the intersection with the 2.1 microperv line yields a beam current of only 1.5 amperes, thus accomplishing the necessary reduction of input power for aural service.

Magnetic Field

Electromagnetic coils are placed around the klystron to develop a magnetic field along the axis of the electron beam which controls the size of the electron beam and keep it aligned with the drift tubes. If the magnetic field is interrupted or insufficiently controlled the electron beam will land on surfaces other than the collector and may destroy the tube.

Cavity Tuning

The resonant frequency of each of the cavities of a klystron can be adjusted in two ways to the operating frequency of the transmitter. The inductance can be changed by changing the volume of the cavity in external cavity klystrons, or the capacitance of the drift-tube gaps can be changed in integral cavity klystrons.

Cavity/Transmission Line Coupling

Fig. 68 illustrates magnetic-loop coupling, where the RF energy is fed through a coaxial line with its center conductor inserted into the klystron cavity. The end of the center conductor is formed into a loop. This forms a simple one-turn transformer which couples RF energy into or out of the cavity through a coaxial transmission line. Intermediate cavities may have their loops coupled into RF loads to vary the "Q" of the cavities to change the overall bandpass characteristics of the klystron.

Effect of RF Drive Power on RF Output Power

Fig. 69 shows RF output power as a function of RF drive power applied to the tube. From this curve, we see that when the RF drive power level is low, the RF output power is low. As the level of RF drive power increases, RF output power increases until an optimum point is reached. Beyond this point, further increases in RF drive power result in less RF output power. Because of these effects, two zones and one point have been labeled on the curve. In the zone labeled "Underdriven", RF output power increases when the RF input power is increased. The point labeled "Optimum" represents the maximum RF output power obtainable. Klystrons are said to be saturated at this point, since any further increase in RF drive only decreases the RF output power. The zone formed at the right side of saturation is labeled "Overdriven". To obtain maximum RF output power from a klystron, sufficient RF drive power must be applied to the tube to reach the point of saturation on the curve. Operating at RF drive levels beyond the saturation point will only overdrive the klystron, decrease RF output power, and increase the amount of beam interception at the drift tubes (body current). Klystrons tuned for TV service are operated within the underdriven zone of Fig. 69.

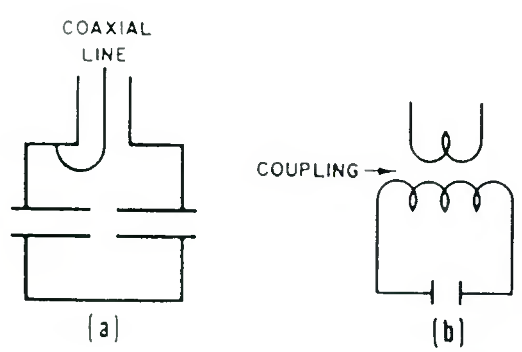


Figure 68. Loop coupling and equivalent circuit.

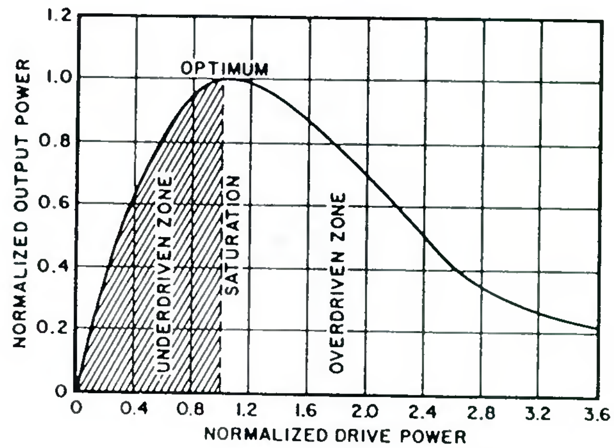


Figure 69. RF output power as a function of RF drive power.

Fig. 70 shows how RF output power changes with various levels of RF drive power applied to a klystron under different tuning conditions. Point A represents the drive saturation point for a synchronously-tuned tube. Point B shows a new point of saturation that is reached by tuning the penultimate (closest to the output cavity) cavity to a somewhat higher frequency. By tuning the penultimate (next to last) cavity still further, Point C is reached. There is a point, Point D, where increasing the penultimate cavity frequency no longer increases RF output power. Instead, it reduces the output power as shown at point E.

Klystron Efficiency

Klystron "efficiency" has often been measured by dividing Peak RF output power by the DC power input. Since klystrons have been compared using "peak RF output power" the term efficiency is not valid because it is possible to have greater than 100% efficiency with some amplifiers.

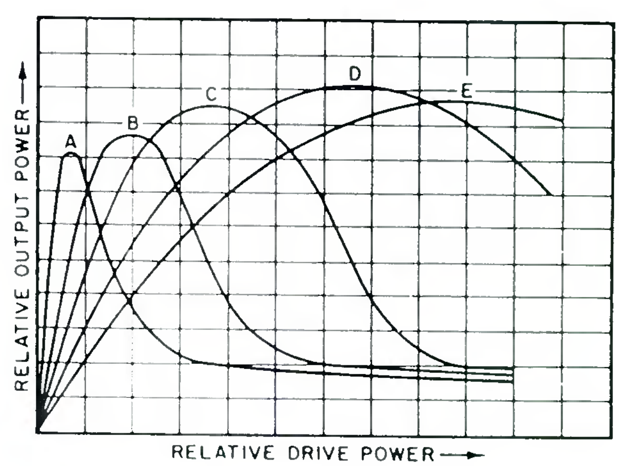


Figure 70. Output power variation with drive power under different tuning conditions.

The term "Figure of Merit" is a more valid terminology and is calculated as follows:

$$\text{Figure of Merit} = \frac{\text{RF power output}}{\text{DC input power}}$$

where DC input power is measured using a 50% APL signal and RF power output is the peak of sync value for visual.

This is a measure of the klystron stage only. Total transmitter or plant efficiency would include the power consumed by the magnets, heat exchanger blower, pumps, driver, and control circuits.

Tuning

There are a number of methods of tuning klystrons. They are:

- High gain tuning for monophonic aural service
- Broadband for MTS aural service
- Visual service with fixed beam current
- Visual service for pulsed beam current
- Tuning for integral cavity klystrons employing the variable visual coupler

It is best to consult the transmitter manufacturer for specific information regarding tuning and mode of operation desired. However, some basic information can be presented here:

AMPLIFIER DEVICE	FIGURE OF MERIT
TETRODE	.9 – 1.0
INTEGRAL CAVITY KLYSTRON	.65 – .75
EXTERNAL CAVITY KLYSTRON	.65 – .75
KLYSTRODE OR IOT	1.1 – 1.3
DEPRESSED COLECTOR KLYSTRON	1.2 – 1.3

General Klystron Tuning Considerations

The output cavity is generally tuned to the carrier frequency. It is essential to operate with the coupling loop adjusted so the output cavity slightly over coupled. Fig. 71 shows the relationship of output power to proper coupling loop adjustment. If the coupling loop is adjusted so that the cavity is under-coupled, arcing and ceramic fracture resulting in klystron failure may occur.

The input coupling is adjusted for the best tradeoff of minimum reflected power and best overall bandpass.

Intermediate cavities may be externally loaded to lower Q and increase bandwidth.

The penultimate or next to last cavity is tuned above the passband. The integral cavity klystron penultimate cavity tuning location is generally 10 MHz to 15 MHz above the passband, while the tuning location for the four cavity external cavity klystron is 6 MHz to 8 MHz above the passband.

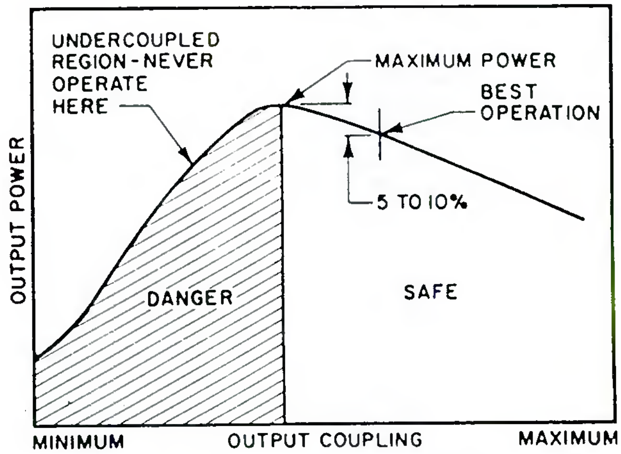


Figure 71. Adjustment of output coupling control.

Variable Coupler for Integral Cavity Tubes

The purpose of the variable load coupler is to improve the klystron Figure of Merit. This is accomplished by raising the impedance presented to the klystron. The variable load coupler as shown in Fig. 72 has a shorted transmission line stub which is tuned beyond a quarter wavelength to present a capacitive susceptance load to the output transmission line. This capacitance susceptance (or inductive reactance) in parallel with the load impedance is transformed back through an approximate ⅜-wavelength to present a substantially resistive load to the output coupling loop of the final klystron cavity.

Transmission line formulas aid in the analysis of how the variable coupler functions.

For example, assume that the transmitter and variable load coupler are terminated with a 50 ohm impedance. If the tuned short is exactly 0.25 wavelength long, the equivalent load impedance at the junction of the tuned short can be represented by:

This impedance, transfored back to the klystron ⅜-wavelength away is given by the equation:

$$Z_{in} = Z_O \frac{[Z_L + jZ_O \tan\beta l]}{[Z_O + jZ_L \tan\beta l]}$$

In this case Z_L = 50 ohms and Z_O = 50 ohms.

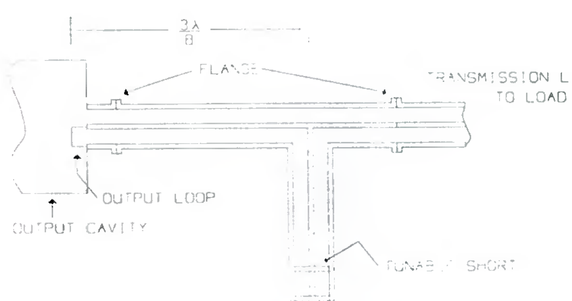


Figure 72. Variable visual coupler.

However, when the tuned short is longer than 0.25 wavelength, an inductive reactance is added in parallel with the 50 ohm load impedance.

Assume for example that the tuned short was lengthened such that the resultant load impedance was now 50 − j10 ohms.

Using the same equation, where $\beta = 2\pi/$(one electrical wavelength) and l = the electrical wavelength between the equivalent load impedance and the klystron.

In this case, l = ⅜-wavelength, $Z_O = 50$, and $Z_L = 50 + j10$.

Solving this equation yields a new impedance at the klystron of about 61 ohms. Thus the impedance has been raised via the tuning stub. To raise the impedance further, the tuning short must be lengthened again.

The iterative procedure of small adjustments which lengthen the tuning stub allows a suitable compromise between efficient operation and klystron safe operation.

Beam Current Pulsing

For a number of years klystrons were operated at maximum beam current. The mod anode was tied to ground through a resistor. Early model klystrons would draw approximately 7.5 amps at 24 kV for 55 kW peak-of-sync visual operation. The development of more efficient klystrons allowed the reduction of beam currents to near 6.3 amps for 55 kW. With saturation set at 115% of needed power, the klystron was operated in the more linear part of the curve, but excess beam current was being consumed.

Operating the klystron at saturation will improve efficiency but requires more linearity and phase compensation. By using a pulser to switch to a higher beam current during sync and back to a lesser current during video the average beam current is significantly reduced.

Fig. 73 shows the horizontal line timing. Observe that sync is 8% of the duty cycle and video is the remaining 92% of the transmitter signal.

The practical limit of reduction of the video beam current is the point at which tip of burst and back porch signal distortion is not correctable.

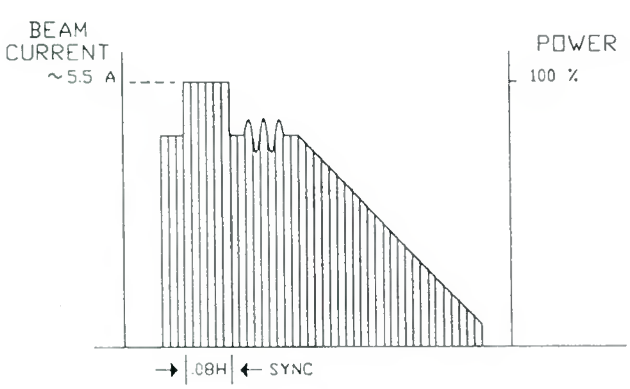


Figure 73. RF envelope vs. beam current.

With beam current pulsing, the effective "Q" of the electron beam and cavity combination is altered. This results as a passband tilt from visual carrier down to the upper edge of the passband. This requires readjustment of the cavities to obtain a flat response. Tilt of as much as 7 dB has been observed. Tuning for high efficiency with beam current pulsing can reduce the gain of a klystron.

Consider the following example of a typical 60 kW external cavity klystron.

In static (nonpulsed) operation, the tube is supplied with enough beam current to saturate at 100% power and a small amount of headroom added for changes in beam voltage. In pulsed operation only enough beam current is supplied to saturate at 100% power during sync. This would normally correspond to about 5.5 amps for a wideband external cavity klystron. The amount of beam current reduction during the video picture is dependent upon operating channel, klystron perveance, available drive power, and precorrection capability.

The optimum value of beam current must be experimentally determined. However, figures of merit of 77% have been achieved. It is important to note that the absolute value of the pulsed efficiency is directly proportional to, and therefore limited by, the efficiency obtained for nonpulsed operation. In order to prevent severe burst and back porch distortion, a small amount of amplifier headroom is needed, approximately 2% to 5%. Typical values of beam current during the video portion of the waveform are 3.4 amps to 3.9 amps.

Pulsed operation for klystrons is accomplished by connecting the beam control electrode to a voltage source of 0 to −1,400 volts with respect to the cathode voltage. During sync, the pulser operates at zero volts. During the video portion of the signal, values of −400 volts to −800 volts are used to achieve reliable high efficiency operation.

In some transmitters, sync is actually reduced or removed from the input video. As the beam current is pulsed, the klystron gain change increases the RF power to produce the proper sync level.

For pulsed operation, DC input power is calculated as follows:

$$\text{DC input power} = \text{Beam voltage} \times [(\text{Beam } I_{sync} \times \text{Sync duty cycle}) + (\text{Beam } I_{video} \times \text{Video duty cycle})]$$

In this example,

$$\text{DC input power} = 24\,[(5.5 \times .08) + (3.7 \times .92)] = 24 \times 3.84 = 92.2\text{ kW}$$

$$\text{Figure of merit} = 60/92.2 = 0.65$$

Effect of Pulsing on Transmitter Precorrection

With increasing RF drive level, klystron amplifiers exhibit a RF phase change in addition to amplitude compression. This phase change is called incidental phase modulation (ICPM). Very little can be done to reduce ICPM by klystron tuning or selection of magnet current. Also, incidental phase distortion increases rapidly near saturation of the klystron.

The RF phase shift through the klystron will also change as beam current is varied. When the klystron is pulsed during sync the change in beam current causes the phase of the signal to change to a new value. When the klystron switches back to the video current level the previous value of phase returns. To combat this problem, ICPM correctors have been developed. These correctors generally operate at the exciter intermediate frequency (IF) and introduce a correction signal equal and opposite to the distortion produced by the klystron. A phase modulation stage in the exciter may be keyed by sync and adjusted to pre-correct for incidental phase distortions caused in the klystron during pulsing. This is most commonly done at the IF level because of the ease of implementation.

When amplifiers are operating very close to saturation during the color burst, more differential gain and differential phase correction may be needed. However, modern exciters can fully precorrect these conditions.

For maximum efficiency, operate the tube at saturation at sync tip. In a pulsed transmitter, the color burst and black picture content are near saturation as well.

The vector diagram of Fig. 74 illustrates first order klystron nonlinearities and the operation of the ICPM and linearity correctors. The desired TV output signal is represented by an amplified version of the desired instantaneous phasor. However, the transmitter output signal is phase shifted by a phase error and compressed in amplitude. To compensate for this, the signal in the exciter is precorrected by an amplitude expansion, and a correction in quadrature. When the resultant signal is amplified, the output signal will be a replica of the desired TV signal.

Sync pulse oscillations may appear as ringing on the sync pulse when the klystron is operated at saturation. This may exhibit itself as a tearing of the picture or sync. This ringing is believed to be caused by secondary electron feedback enhanced by the reverse gain of the klystron cavities. Rebiasing the tube slightly out of saturation will eliminate the ringing.

Multi-stage Depressed Collector (MSDC) Klystrons

Klystron amplifiers operate by converting energy from a beam of electrons to RF output power. At full output power, about half of the beam power is converted to RF output power. Correspondingly, half the DC input power remains on the beam as it exits the cavity region. In a standard klystron this "spent electron beam" energy is dissipated as heat. The MSDC klystron, however, operates on the spent electron beam, recovering its energy to reduce the dissipated heat.

After the electron beam has completed its job by providing the desired output power, significant energy still remains in the electron beam. The interaction process has produced a wide range of velocities for individual electrons, from nearly stopped to up to twice the initial energy.

In the early 1970s, researchers at NASA investigated collector designs. In 1984, NASA provided the depressed collector technology for development of UHF-TV klystrons. Power recovery in the collector region is accomplished by decelerating the electrons in the spent beam. This can be done by providing an electric field in the collector such that the electrons are slowed before they strike the collector wall. A collector composed of multiple elements is utilized, with each element operated at a negative potential with respect to ground potential.

Consequently, the collector potentials are referred to as being *depressed* below ground. The collector element geometry was carefully selected to provide an electric field shape which would sort the impinging electrons according to their velocity, reducing their energy as much as possible, yet ensuring that all electrons would strike one of the elements and not

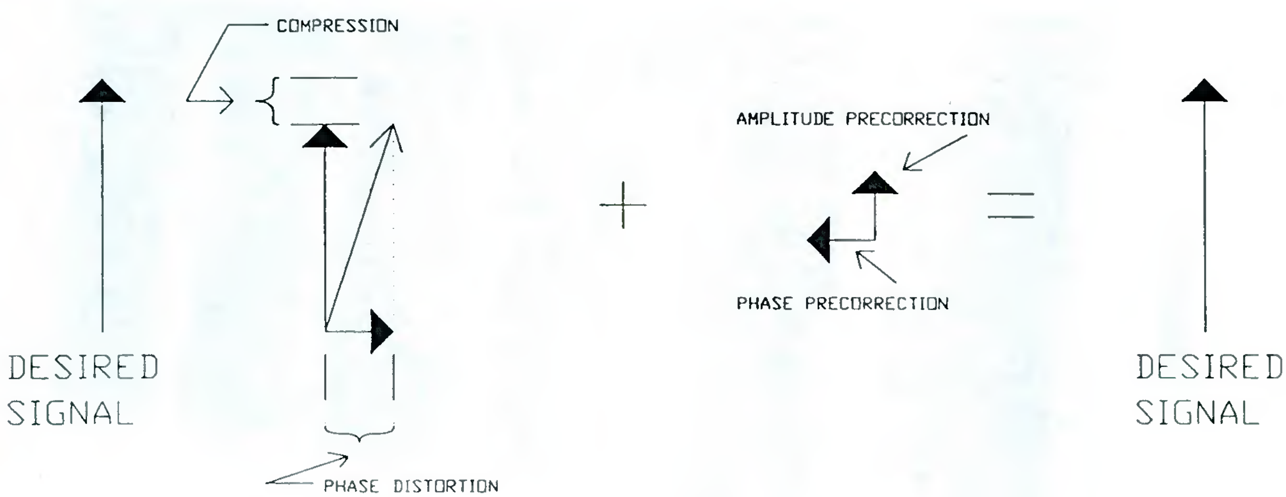


Figure 74. Vector representation of precorrection.

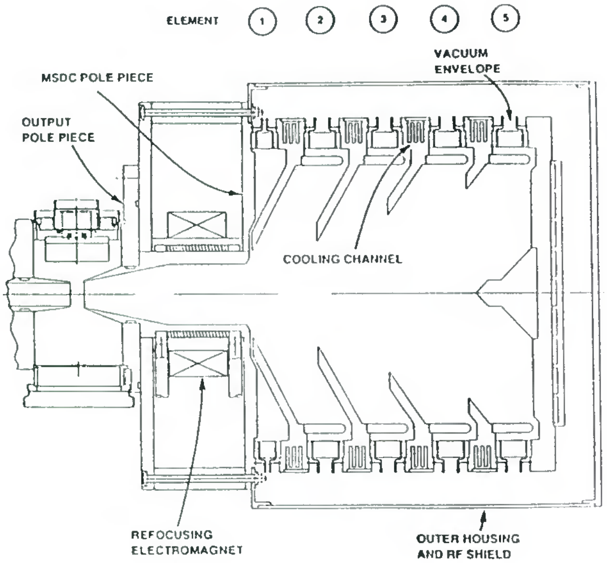


Figure 75. UHF-TV MSDC design.

be reflected back into the electron beam. Modern computer aided techniques facilitated this design effort.

Fig. 75 shows the resulting collector configuration. The collector is composed of five elements and is designed to operate with equal voltages between elements. For the 60 kW klystron, the voltage per collector stage is typically 6,250 volts. This means that element 1 is at ground potential, element 2 is at −6.25 kV, element 3 is at −12.5 kV, element 4 is at −18.75 kV, and element 5 is at full cathode potential of −25 kV.

To make sure that the electron beam is optimized before entering the collector region, a refocus coil is provided just ahead of the collector.

The significant benefit of the depressed collector klystron is the reduction in power consumption for a given power output. The individual collector beam currents will vary depending on the picture level being transmitted. Since the recovering of the spent beam takes place in the collector, it has essentially no impact on the tuning and normal precorrection of the klystron. Therefore, a multi-stage depressed collector klystron tunes the same as a standard klystron and requires the same precorrection as a standard klystron. The depressed collector technique can be applied to either external or integral cavity klystrons. The cooling of the cavities is not affected by the addition of the collector.

To date, these klystrons require the use of a vacuum ion pump. Since the power dissipated in the collector is reduced, only six gallons per minute of high purity water coolant is needed. Although the efficiency performance of the MSDC is dependent upon the transmitter configuration, figures of merit of 1.2 to 1.6 have been obtained in transmitter installations in the field.

Transmitter Design Using Multiple Depressed Collector Klystrons

The primary differences in this transmitter from a standard klystron transmitter are in the beam supply, the cooling system. Fig. 77 shows the power connections made to a MSDC klystron. Since the RF performance of the MSDC klystron is the same as the standard klystron, there are no differences in the RF driver chain.

The cooling system of a MSDC klystron transmitter uses a two-stage heat transfer system as shown in the cooling system block diagram of Fig. 78. The reason a two-stage system is chosen is to allow outside heat exchangers to be used. The cooling system consists of a high purity water loop and a glycol-water mixture

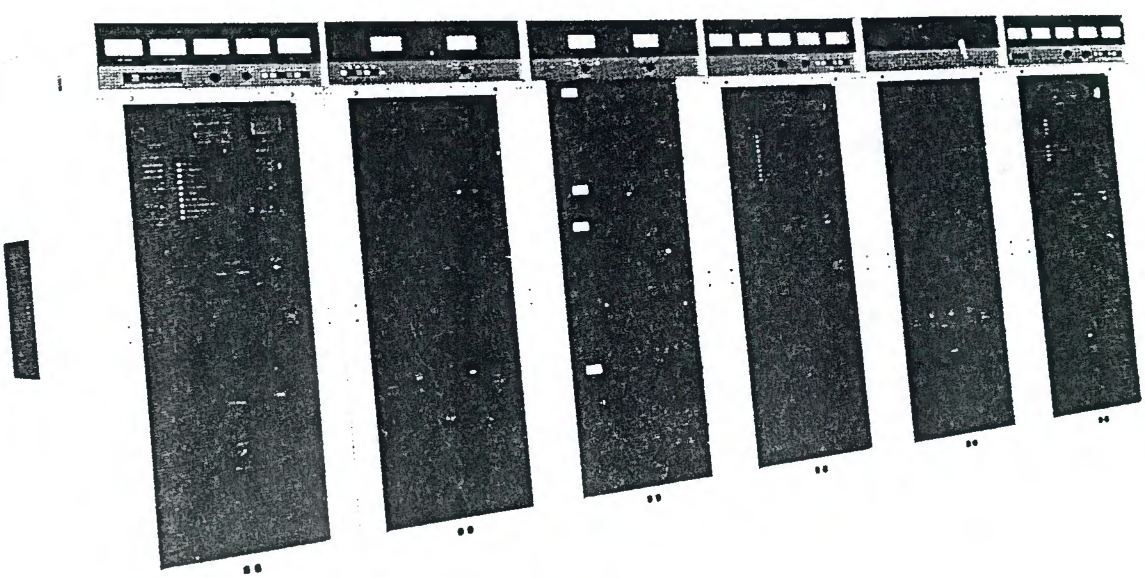


Figure 76. MSDC transmitter.

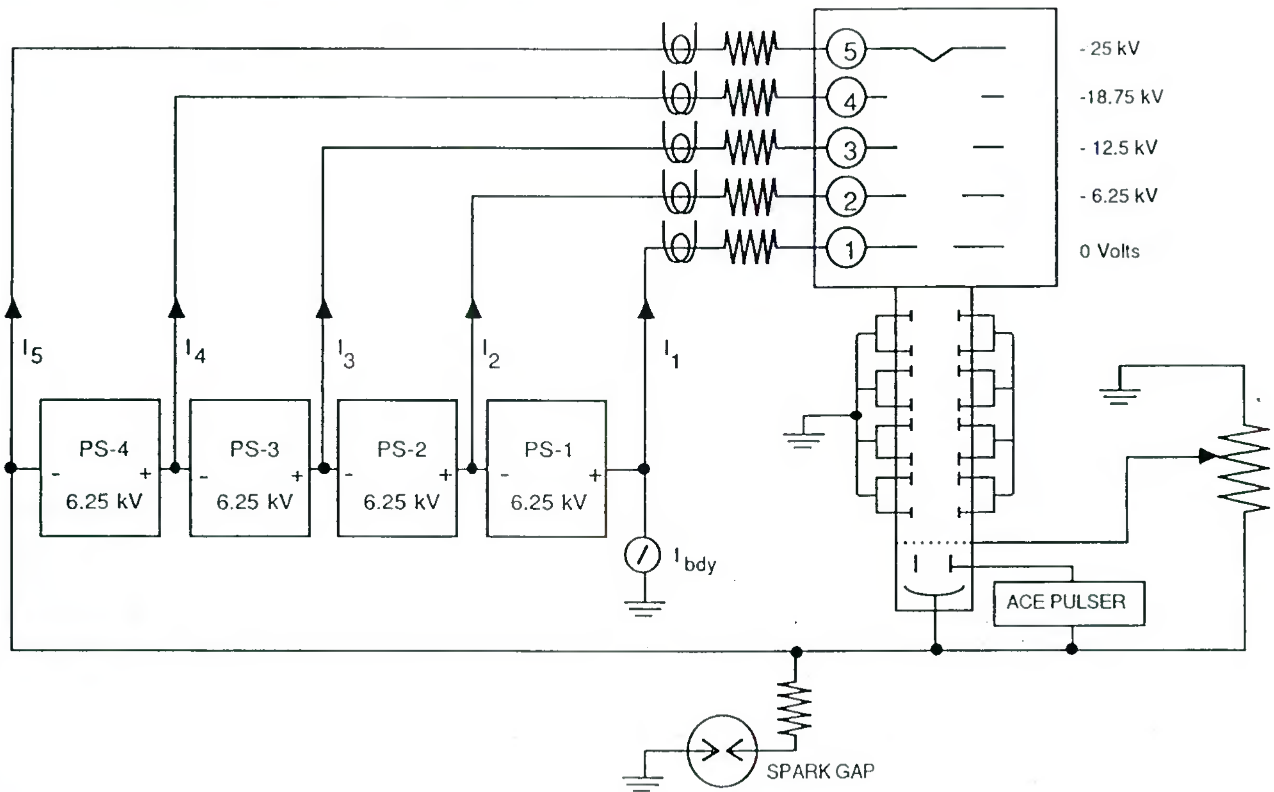


Figure 77. MSDC high voltage connections.

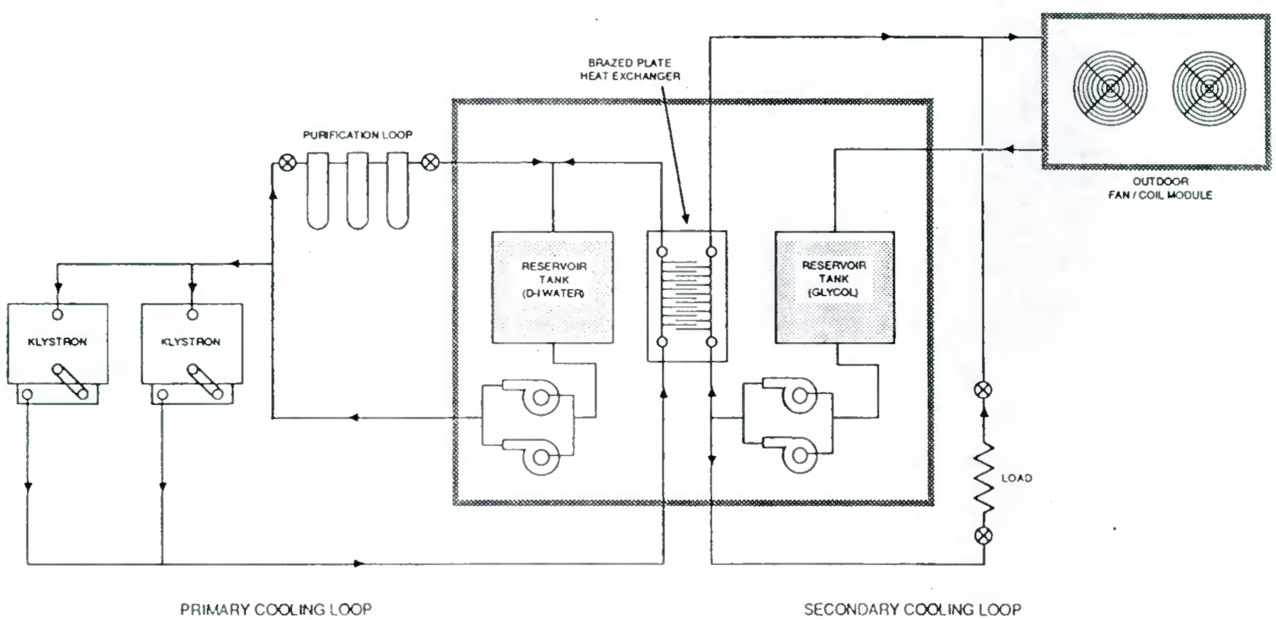


Figure 78. TV-60UM heat exchanger system.

loop. High purity water (resistivity of 200,000 ohm-cm or more) must be used with the cooling system to prevent current flow between collectors. To remove ions, free oxygen and other possible contaminants from the water, a three-stage filter system (purification loop) is used. The filter cartridges sample part of the water flow so the whole system is continuously cleaned. The filter cartridges can be replaced without taking the transmitter off the air.

Separate beam supplies for each klystron are frequently used in TV transmitters.

Since the currents from the beam supplies will change dependent on the power level, there will be video frequency currents present on the power supply leads. Therefore, sufficient bypassing and power supply high voltage wire shielding is required.

Monitoring each section of the beam supply current is required to obtain the currents for power dissipation calculations.

Protection circuitry needs to include magnet overcurrent and undercurrent trip points, beam supply overcurrent and overvoltage sensors (for each collector), water flow sensors, arc detectors within the 3rd and 4th cavities, and sufficient interlocks to prevent personnel from accidentally coming in contact with high voltage.

KLYSTRODE®

. The Klystrode® basically combines features of a tetrode and a klystron. Hence, the name Klystrode® was created.

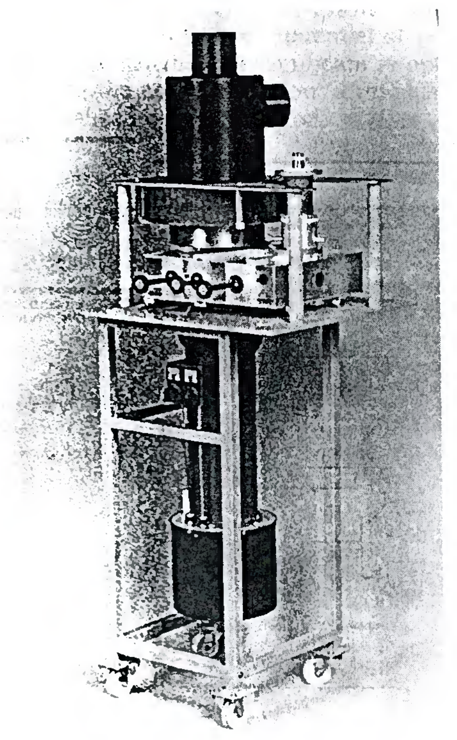


Figure 79A. Klystrode.

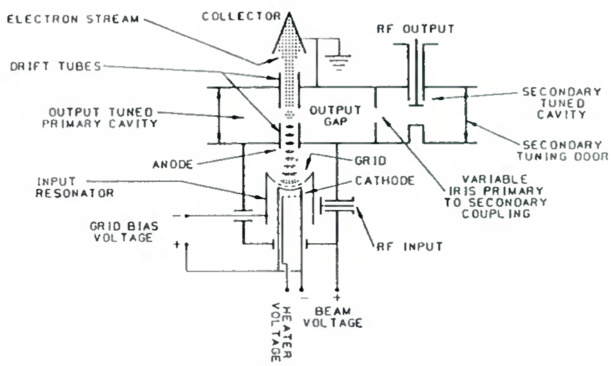


Figure 79B. Klystrode schematic.

A Klystrode® is shown in Fig. 79A. It is composed of an electron gun very similar to a klystron, a control grid, an input cavity, accelerating anode, drift tube, output cavity, and collector. It is physically smaller than a klystron and weighs approximately 90 pounds including its circuit assembly. The tube itself weighs about 45 pounds.

Refer to Fig. 79B. The electron beam is formed at the cathode, density modulated with the input RF signal applied to a grid, and then accelerated through the anode aperture. In its bunched form, the beam drifts through a field-free region and then interacts with the RF field at the drift tube gap in the output cavity. Power is extracted from the beam in the output cavity in the same manner as a standard klystron.

RF input power is applied to the control grid via a resonant cavity. The grid is usually biased negatively. The first part of the tube may be thought of as a triode with a perforated anode through which the electron beam is guided by electric and magnetic fields. The beam is bunched, or chopped, at the radio frequency and is accelerated by the high anode potential. It passes through the anode extension cylinder, which is an electrostatic shield, and then interacts with the RF field in the output gap. The spent beam is dissipated in the collector, separate from the output RF interaction circuit.

The tuned input and output circuits are external. The RF input circuit is a 4 stub tuner which matches the drive source to the high impedance grid. Also included is a circuit which provides a DC block for high voltage. The output circuit cavity is clamped to the body of the tube with techniques similar to those used for external cavity klystrons. The output circuit is double-tuned to achieve the bandwidth required for visual service. The double-tuned output circuit consists of a primary cavity clamped to the body and an iris-coupled secondary cavity. The output transmission line is probe coupled to the electric field in the secondary cavity. Variable controls are used to adjust primary and secondary cavity resonant frequency and the iris coupling.

The grid structure intercepts some electrons causing grid current to flow. On 60 kW models, some material from the cathode migrates to the grid and must be periodically boiled off. A design goal of future Klystrodes® is to eliminate this periodic procedure.

Because the tube only has two cavities it is much shorter than a klystron. The magnetic field requirements of the tube are about 7 volts at 30 amps.

The high voltage circuitry is contained within a shielded compartment on top of the input circuit. This circuitry consists of various filters to contain the UHF fields and prevent instabilities at video frequencies. High voltage, grid bias, and filament power enter this section via high voltage cables.

The fundamental benefit of the Klystrode® is that it may operated as a Class B amplifier. Thus the beam current is proportional to the RF drive signal. Although the efficiency performance of the Klystrode® is dependent upon the transmitter configuration, figures of merit of 1.2 to 1.4 have been obtained for 60 kW visual service from transmitter installations in the field.

In aural service, the Klystrode® is tuned the same as for visual service.

Power gain in either type of service is about 23 dB. Thus drive power is about 300 watts for the visual and 30 watts for the aural (assuming 10% nominal aural power).

The transfer characteristics of the Klystrode® is also a combination of a klystron and a tetrode. A typical transfer curve is shown in Fig. 80. The nonlinearity at white picture power levels resembles that of a tetrode and the nonlinearity approaching peak output power also resembles a tetrode. The amount of nonlinearity is similar to a tetrode and the shape of the precorrection curve is "S" shaped.

Klystrodes® at 15 kW, 40 kW, and 60 kW visual peak sync power levels have been constructed and are in service.

The Klystrode® may also be used as a multiplexed amplifier. This means the diplexer may be eliminated. For example, a tube may be used as a 30 kW visual with a 10% aural signal. Intermodulation products at ±920 kHz from visual carrier generated by the tube can be precorrected by low level IF circuitry.

Cooling of the Klystrode® at the 60 kW power level requires about 25 gallons per minute of water for the collector. The body of the tube is also water cooled.

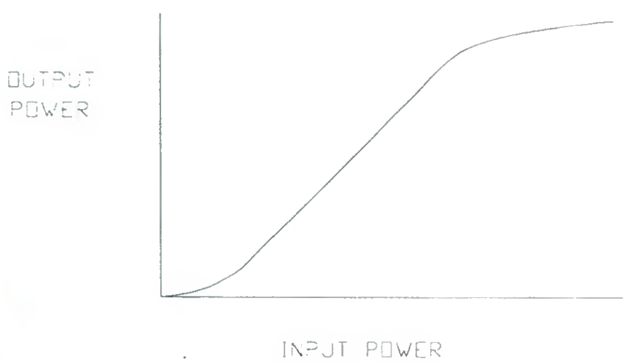


Figure 80. Klystrode transfer curve.

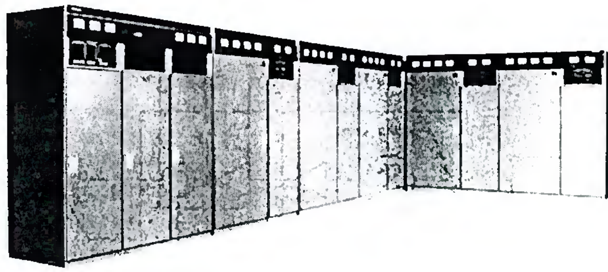


Figure 81. Klystrode transmitter.

A 50/50 water/glycol solution is typically used without any special water purification. The RF input and output cavities are air cooled.

The 15 kW and 40 kW Klystrodes® may be air-cooled. The volume of air and air pressure needed are comparable to a tetrode of similar power.

Two versions of Klystrodes® are needed to cover the entire UHF operating band.

To date, these devices require the use of a vacuum ion pump.

Transmitter Design Using Klystrode®.

Support circuitry for the Klystrode® consists of providing the necessary drive power, precorrection, different types of power supplies necessary, and the protection circuitry. A Klystrode® transmitter is shown in Fig. 81. A Klystrode® transmitter block diagram is shown in Fig. 82.

The Klystrode® uses a beam voltage of 32 kV. Since the beam current changes with modulation, there will be video frequency currents required from the beam supply. The beam supply must be designed to provide excellent regulation from no load to full load and to also provide a low source impedance for all video frequencies. In the event of a high voltage failure, the conventional beam supply should limit the energy dumped into an arc. The Klystrode® supply, being stiffer, requires a triggered "crowbar circuit" to limit the beam supply arc energy. A block diagram of a crowbar circuit is shown in Fig. 83.

The grid bias supply of −10 to −70 volts, with respect to the cathode, typically should "float" with the beam voltage. A simple method of developing the grid voltage is to use zener diodes connected between the power supply and the tube cathode connection and tap the grid to the appropriate zener to obtain the desired bias current. As with klystrons, the Klystrode® magnet power supply must have sufficient energy storage that the beam remains focused until the beam decays. Also, a power supply for the ion vacuum pump is needed with appropriate sensors for the protection circuitry.

Other protection circuitry provided in the Klystrode® transmitter is similar to that needed in the klystron transmitter.

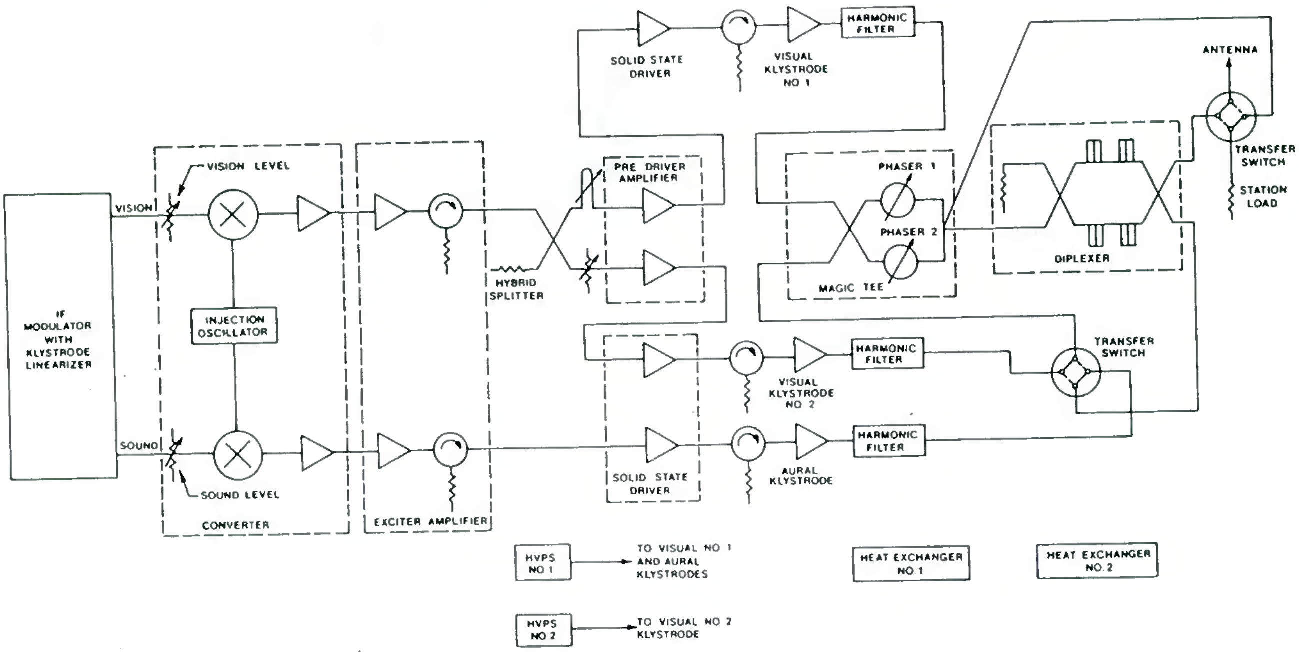


Figure 82. Klystrode transmitter block diagram.

TRANSMITTER COMBINING CIRCUITS AND RF SYSTEMS

A typical RF system for a transmitter will consist of hybrid combiners, (if more than one amplifier cabinet is used), harmonic filters, −3.58 MHz notch filter, and diplexer to combine visual and aural.

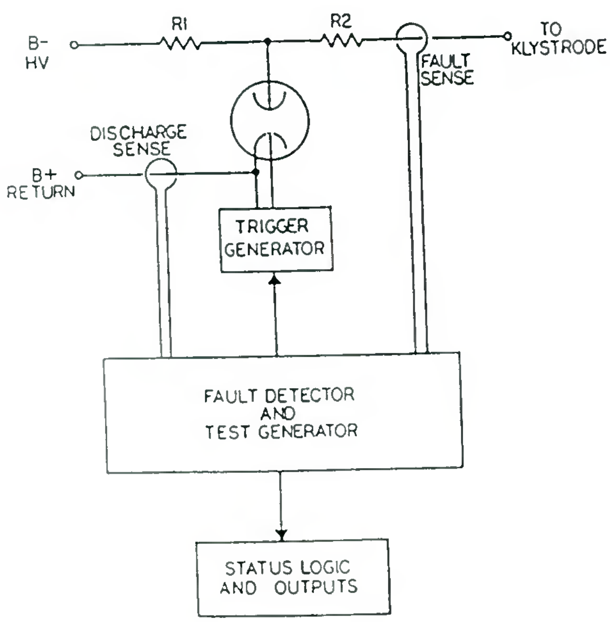


Figure 83. Triggered crowbar circuit.

At VHF frequencies the −3.58 MHz notch filter, harmonic filters, and diplexers are all quite large and are typically supplied as individual units. In some cases, the −3.58 MHz filter is incorporated with the notch diplexer. However, at UHF frequencies, components are smaller and waveguide technology is typically used so the −3.58 MHz notch filter, the notch diplexer aural cavities, and associated hybrids are often integrated into one assembly. Multiple aural notch cavities may be used at UHF frequencies to optimize multichannel TV sound performance and for power sharing when very large powers are used.

This section will review hybrid combiners, notch diplexers, switchless combiners, and Magic Tee RF systems.

Two different types of combining circuits must be considered:

1. Combining of sources with the same frequency (i.e., multiple power amplifiers).
2. Combining of sources with different frequencies (i.e., visual and aural).

For combining of sources of the same frequency, 90° 3 dB hybrid couplers have found almost universal acceptance. Fig. 84 shows the physical model of a 90° 3 dB hybrid coupler.

A 3 dB hybrid coupler consists of two identical parallel transmission lines mounted in a common outer conductor and coupled over a length equal or approximately equal to the quarter wavelength of the operating frequency.

The construction is symmetrical, i.e., both inner conductors have the same physical dimensions with

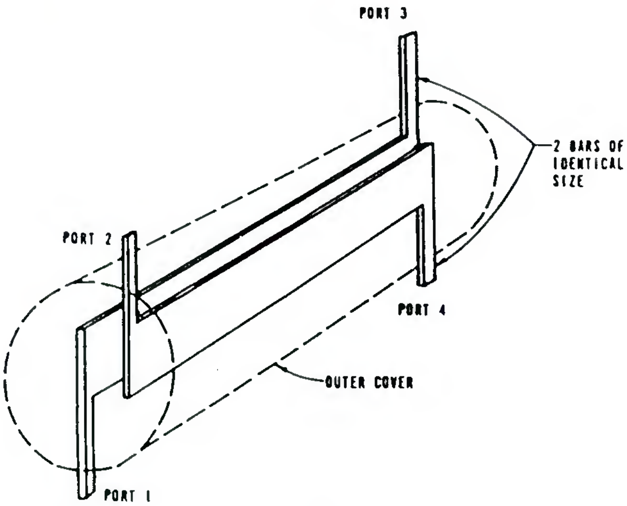


Figure 84. 90° 3 dB hybrid coupler.

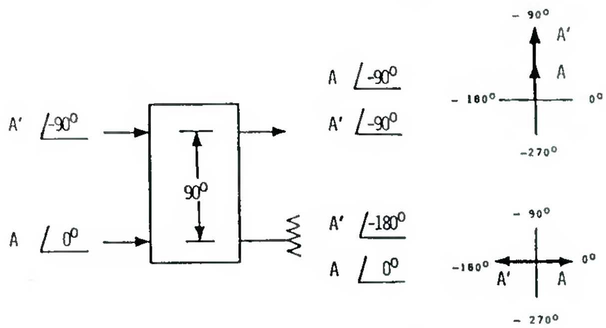


Figure 86. 3 dB hybrid as a power combiner.

respect to the outer conductor. The 3 dB hybrid coupler can be used as a power splitter or as a power combiner. Fig. 85 shows the 3 dB hybrid coupler as a power splitter, and Fig. 86 shows the 3 dB hybrid coupler as a power combiner.

One of the common uses of the 3 dB hybrid is to combine two frequencies that are relatively close (visual and aural) in a back-to-back hybrid configuration called a notch diplexer. The first step in building a notch diplexer can be shown by combining two 3 dB hybrids back to back as shown in Fig. 87A.

The input hybrid will act as a splitter while the hybrid at the output will act as a combiner. The visual signal is inserted into the left hybrid and will appear at the antenna port as shown in Fig. 87B.

If we now add the aural signal as in Fig. 88, we can see that it is passed through the 3 dB hybrids but does not appear at the antenna port as required.

If we now add two very high Q notch cavities tuned to the aural frequency, as depicted in Fig. 89, we can introduce a short circuit to the aural frequency. These shorts circuits cause the aural signal to be reflected back into the 3 dB hybrid where the same 3 dB hybrid now effectively acts as a power combiner and passes the combined aural signals to the antenna port along with the visual signal.

The aural notch cavities are theoretically not seen by the visual signal since they are high Q cavities tuned to the aural frequency. In reality, the aural cavities do cause some amplitude roll-off and group delay of the high end of the visual passband.

Another use of the back-to-back hybrids can be seen in Fig. 90 which shows a single signal source driving a power splitter (Hy1) and dual amplifiers (A1 and A2). Real world amplifiers do not have exactly the same gain and phase shift through the amplifier especially if the amplifier has tuning controls. This hybrid arrangement can be used for visual or aural signals. Gain (AT1 and AT2) and phase adjustments are used to keep the amplifiers' output signals (A1 and A2) at equal output power and 90° phase difference so that they can be properly combined by the power combiner (HY2) with a minimum of reject load power.

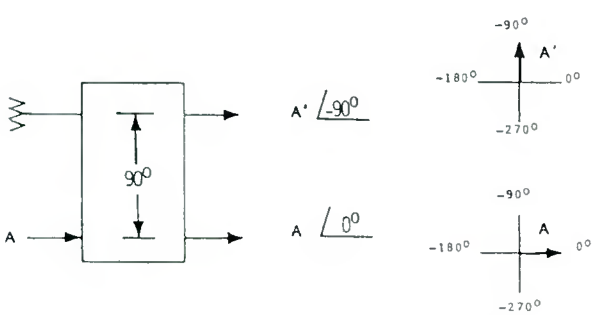


Figure 85. 3 dB hybrid as a power splitter.

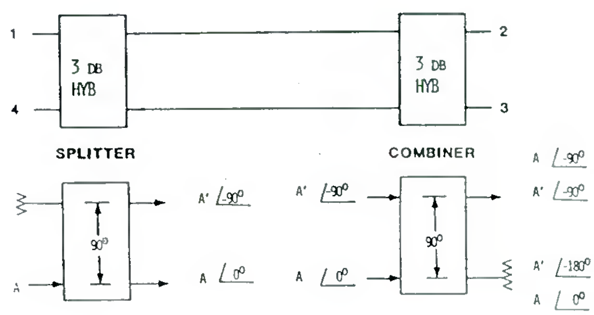


Figure 87A. Back-to-back hybrids.

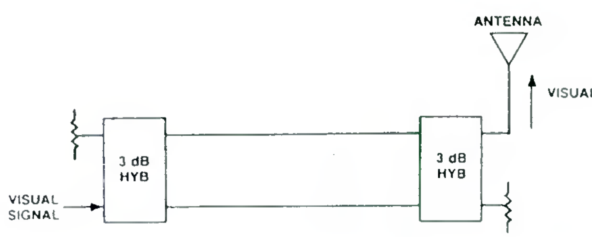


Figure 87B. Visual signal applied to back-to-back hybrids.

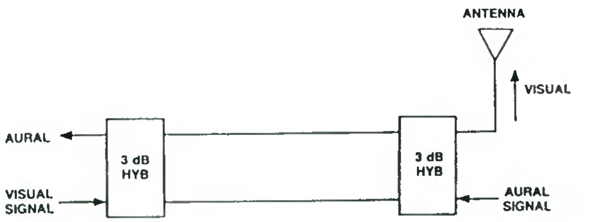


Figure 88. Aural signal applied to back-to-back hybrids.

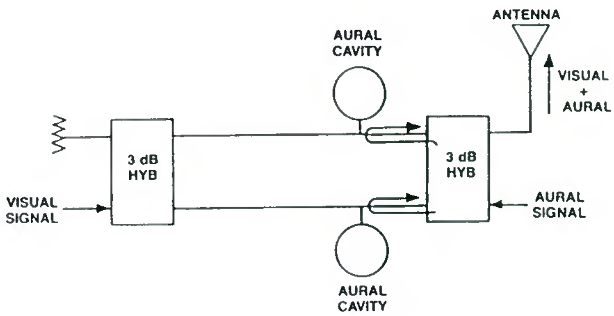


Figure 89. Aural and visual signals applied to back-to-back hybrids with aural notch cavities to reflect aural signals back to antenna port.

A typical use of this system would be the driving of two power amplifier chains in parallel with a single exciter. HY1 would be a low power splitter while HY2 would be a high power combiner. A1 and A2 would be the complete amplification chain consisting of multiple stages of solid state and/or tube amplifiers.

Differences in amplifier chain gain are compensated for by attenuators AT1 and AT2 which vary the amount of input signal applied to the amplifier chains. Differences in phase are compensated for by the length of cable between the attenuators and the amplifier chain. Optimum phase and gain adjustments are determined by minimum power being dissipated by the reject load and maximum power into output load as shown in Fig. 90.

Effective power transferred to antenna or load with respect to differences in phase or gain of amplifier chains can be seen in Figures 91 and 92 respectively. As can be seen from these figures, a 60° error in-phase will cause only a 25% reduction in output power. If one amplifier has only half the output power of its counterpart, a 3% reduction in power (in reference to the combined input powers) will result. For example, let transmitter A = 5 kW, transmitter B = 10 kW, K^2 = Pa/Pb = 0.5 which gives a K of 0.707 which equals approximately 3% in Fig. 92 or approximately 14.55 kW of useful power out of the hybrid. Worst case for amplifier gain differences occurs when one transmitter is not producing any output. In this case half of the remaining transmitter's output (25%) will be dissipated in the reject load and the other half (25%) will be applied to the antenna.

A typical dual transmitter with notch diplexer is shown in Fig. 93.

Switchless Combiners

As discussed in the previous section, dual transmitters may be combined using hybrids to obtain maximum power. However, if one of the transmitters were to be disabled, the combined power output would drop to 25% of the nominal value. In order to continue operating at a reasonable power level many stations have employed coax switches to bypass the hybrid as shown in Fig. 94 to boost the transmitter power back to 50% of nominal operating power. During the switching process, however, it is necessary to take the transmitter off the air.

An alternative which allows the transmitter to stay-on-the-air while changing power from 25% to 50% is to use a "switchless combiner," A switchless combiner uses phase shifting with back-to-back 3 dB hybrids to accomplish the correct direction of power output.

A diagram of a transmitter using a switchless combiner is shown in Fig. 95. As identified in Table 1, nominally both transmitters are producing equal power and

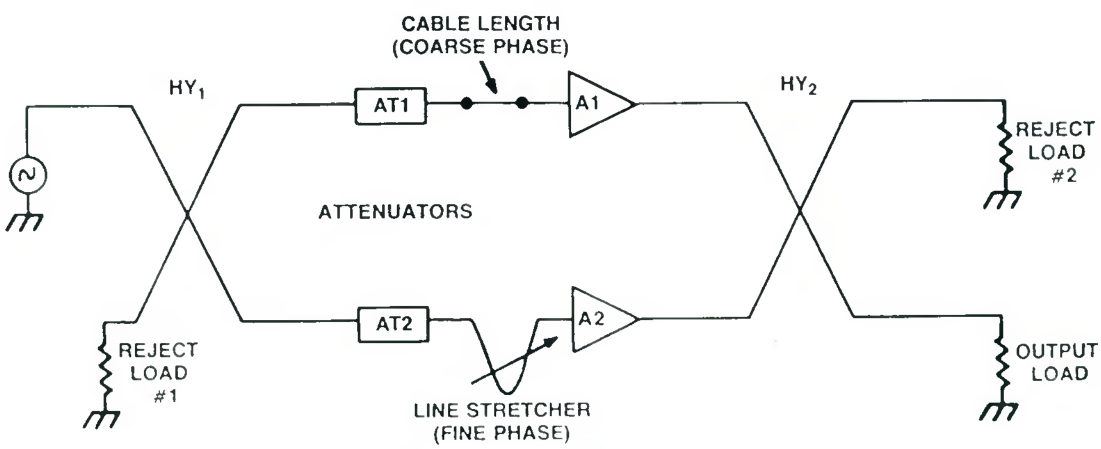


Figure 90. Typical parallel amplifier with single signal source.

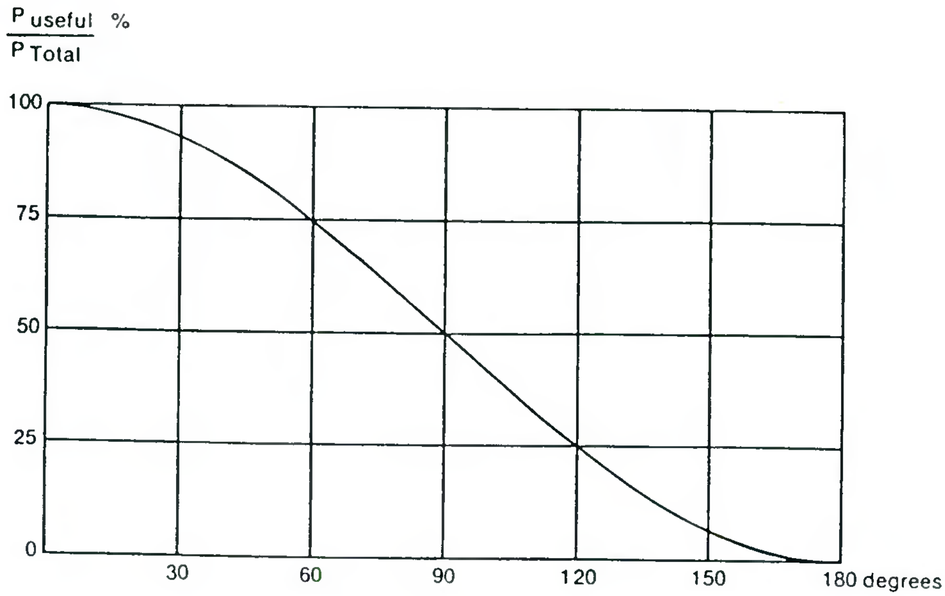


Figure 91. Hybrid coupler phase sensitivity.

are 90° out of phase. When transmitter A is disabled, Phase Shifter 1 (PS1) is energized changing the phase relationship as shown in Table 1. Transmitter A is routed to the switchless combiner load. When transmitter B is disabled (and assuming transmitter A is enabled) Phase Shifter 2 (PS2) is energized changing the phase relationship as indicated in Table 1. Now, transmitter B is routed to the switchless combiner load.

With one transmitter routed to the antenna and one transmitter routed to the combiner load, it is possible to perform adjustments on the transmitter connected to the load without impacting the on-air signal provided a separate exciter is used. (For more detail on this subject see "Subjective Tests on a Switchless Combiner TV Transmitter System" noted in the bibliography.)

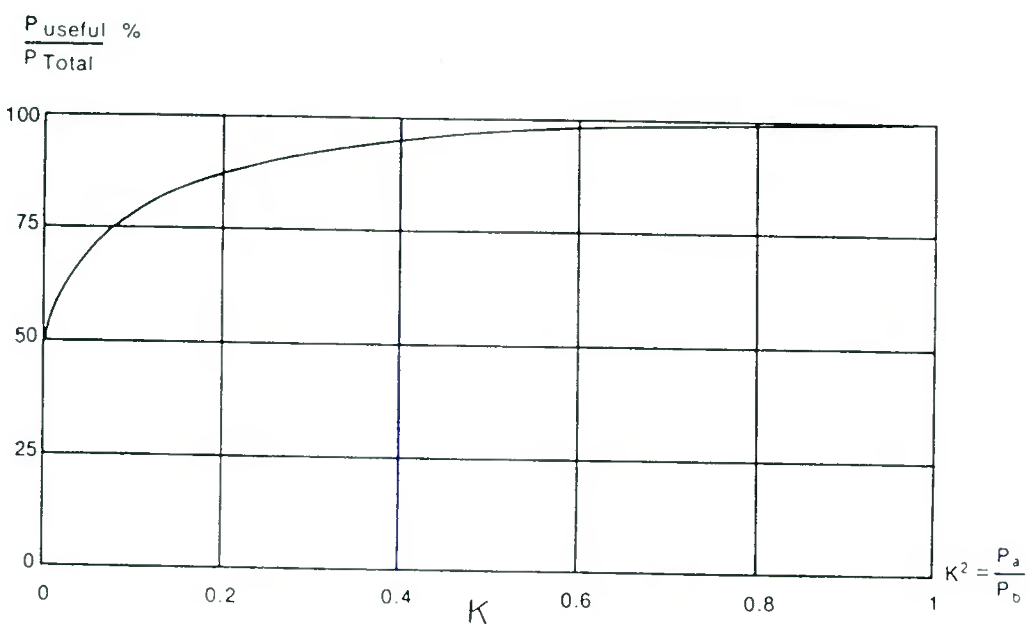


Figure 92. Power imbalance in hybrid couplers.

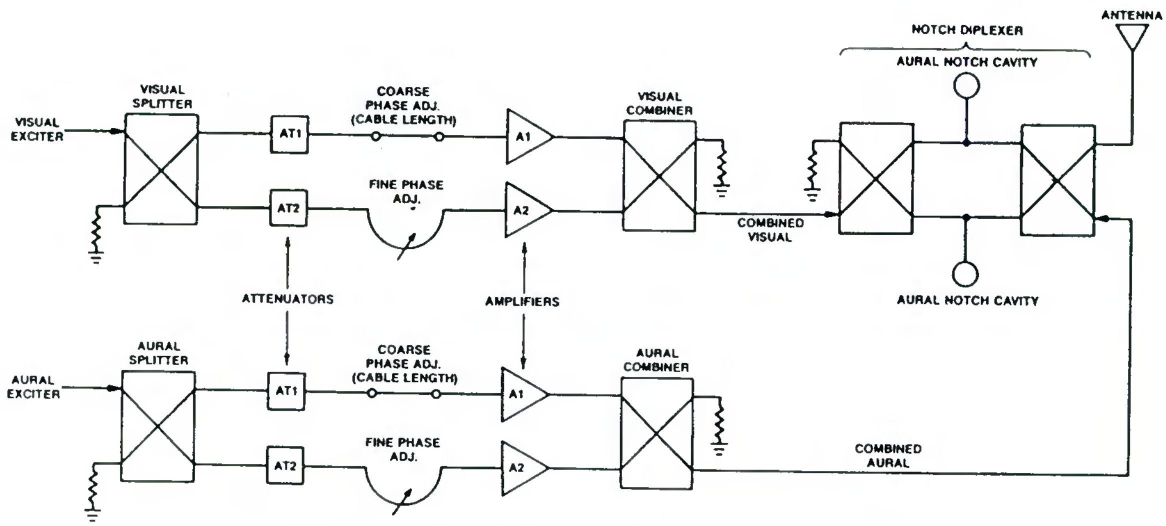


Figure 93. Typical dual transmitter with notch diplexer.

The switchless combiner concept has significantly increased the on-air availability of VHF transmitters.

UHF Magic Tee RF Systems

Since a similar problem exists for UHF transmitters, a similar concept can be applied again. However, since components are smaller many of them can be integrated into one assembly.

UHF transmitters typically are not dual systems but often do employ multiple visual amplifiers as shown in Fig. 96. A 180° hybrid implemented in waveguide accompanied by phase shifters located in the waveguide can accomplish the same function as the switchless combiner does with dual transmitters. See Table 2 for a diagram of what the phase relationships are and which amplifier is routed to which device.

One example of a unitized RF system, shown in Fig. 97, consists of a waveguide magic tee combining/switching system, a waveguide notch diplexer, two waveguide transfer switches, a 50 kW coaxial reject/test load, and a 100 kW waveguide test load interconnected and mounted within an open frame. The "magic tee" system provides visual amplifier power combining and power routing of either visual amplifier to the

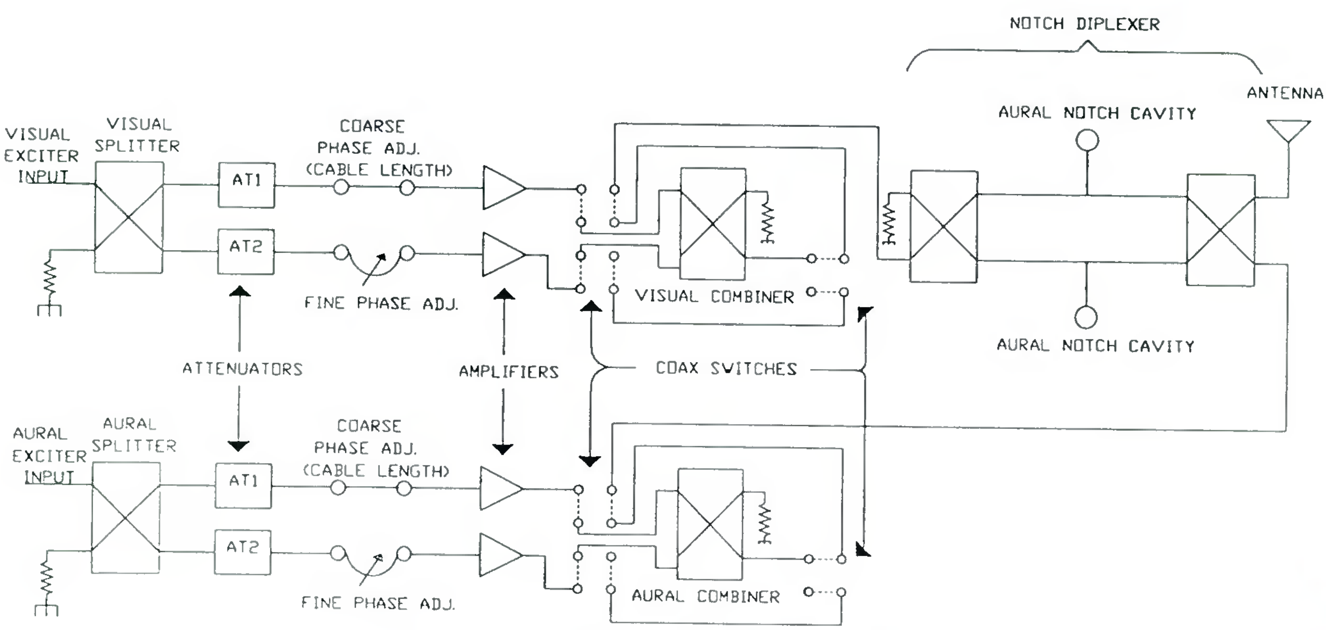


Figure 94. Typical dual transmitter with coaxial switchers.

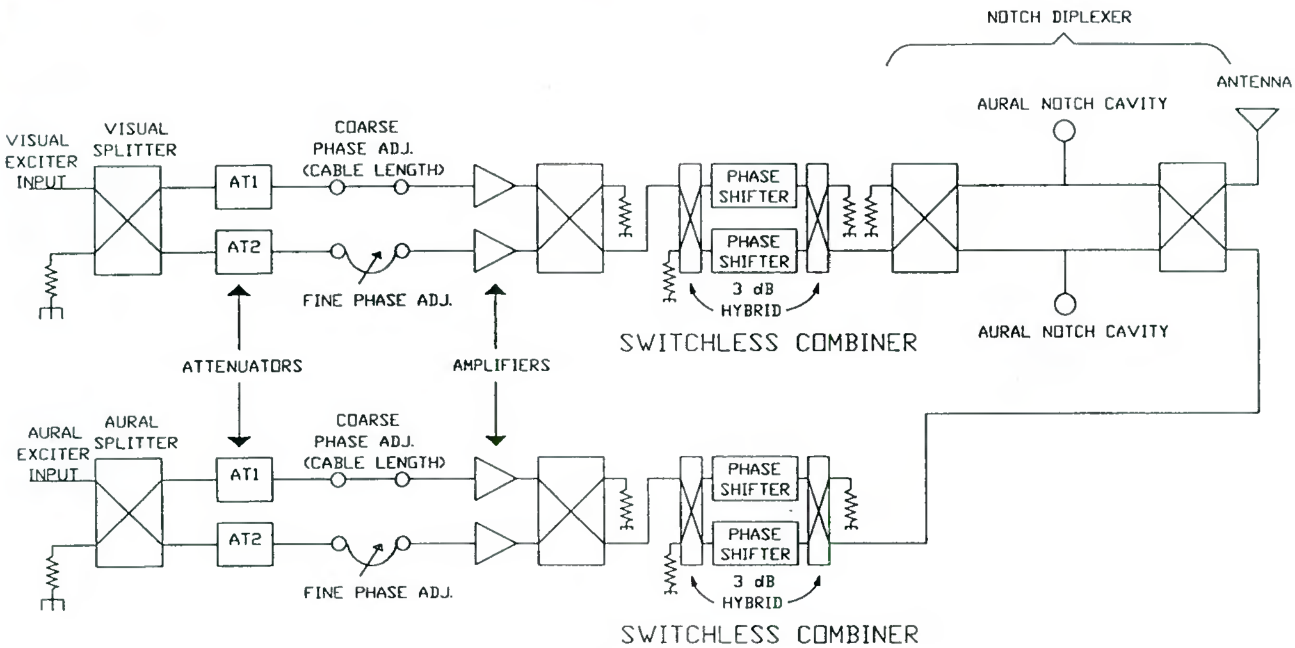


Figure 95. Typical dual transmitter with switchless combiner.

TABLE 1

A + B on AIR	180	270
A on AIR and B in TEST	180	180
B on AIR and A in TEST	90	270

TABLE 2

A + B	AIR	180	180
A AIR	B TEST	90	180
B AIR	A TEST	180	90

diplexer or 50 kW coaxial test load. The diplexer combines the visual and aural transmitter outputs. It is equipped with a lower color subcarrier notch filter and aural notch cavity de-tuning mechanisms to allow multiplexed visual/aural signals to be passed through the diplexer from visual input port to the antenna output port during emergency operation. One waveguide transfer switch routes the diplexer output either

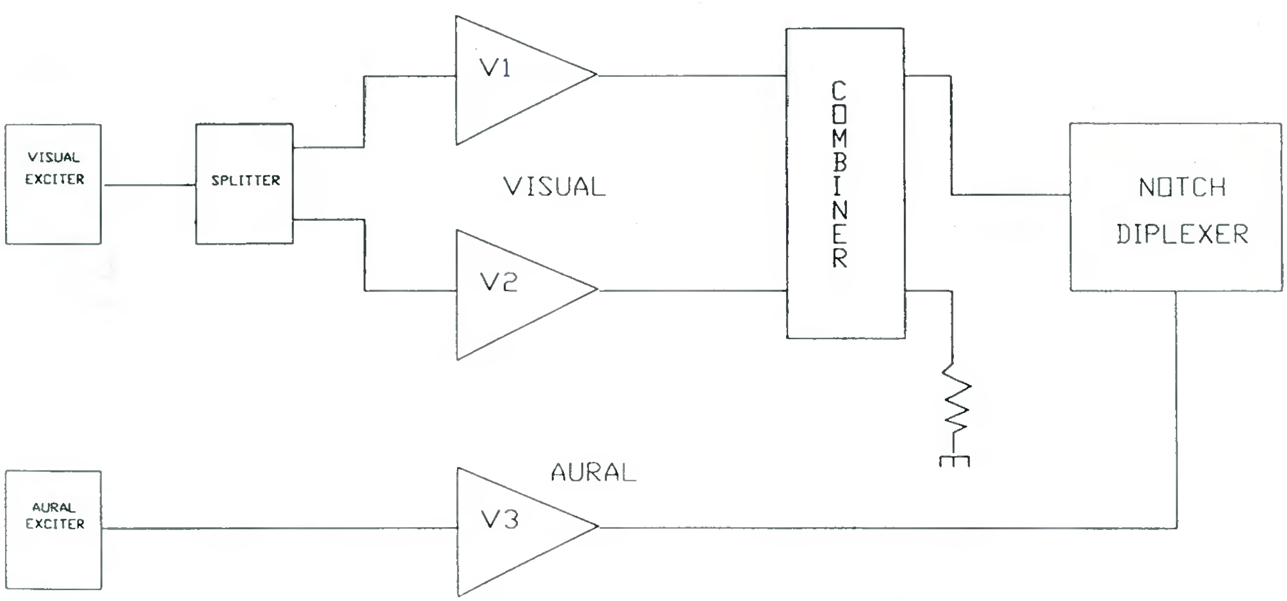


Figure 96. UHF transmitter with multiple visual amplifiers.

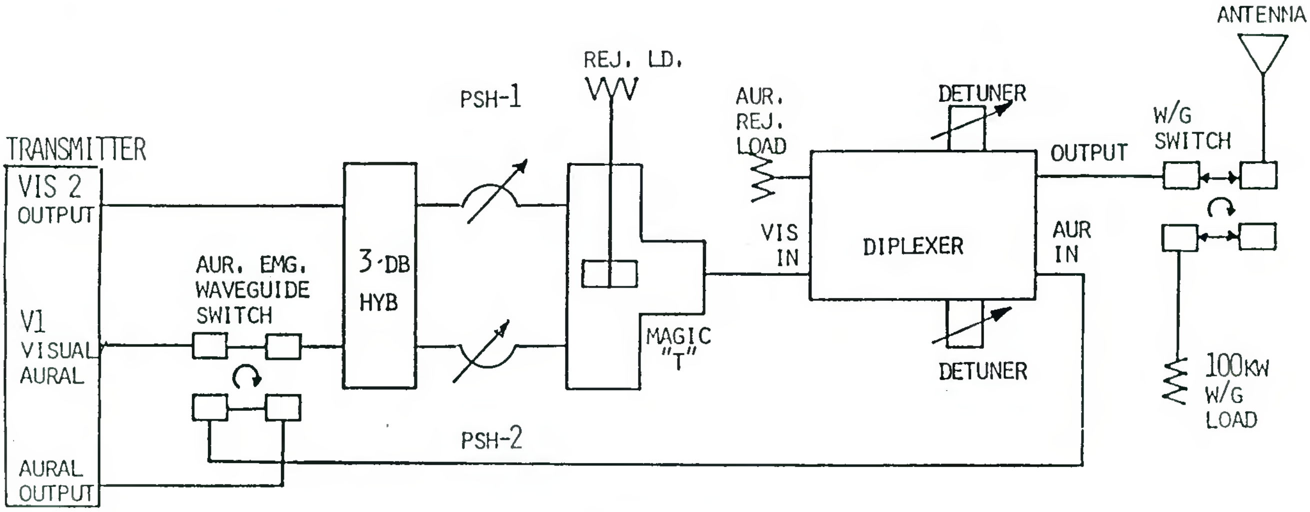


Figure 97. Unitized RF system.

to the antenna or to the 100 kW waveguide test load. The other waveguide transfer switch can route the output of a visual amplifier to the aural input of the diplexer. This feature allows a visual amplifier to become an aural amplifier in the event of failure of the normal aural amplifier. A control logic system is installed in the transmitter to coordinate transmitter and RF system operations. All functions of the RF system can be operated from the front of the transmitter or from a remote location.

PERFORMANCE MEASUREMENTS

The quality of the broadcast television transmitted signal is the resonsibility of the broadcaster. The introduction of IF modulation, solid state SAW vestigial filters and sophisticated precorrection circuits improved critical signal performance parameters two to one. Precision demodulators with synchronous detection and SAW filters are capable of near ideal detection. New test waveforms and digital signal synthesis now provide accurate test signal generation to complement the improved transmission facilities.

Home receivers employ high tech circuits including comb filters for luminance, stable SAW IF filters, high resolution display devices, and sophisticated LSI video processing circuits.

It is difficult to show the correlation between signal quality and audience ratings. The issue of viewer enjoyment is easier to demonstrate. Many scientific studies relating signal degradation to perceived quality have been made over the years.

The Electronic Industries Associates Standard RS-508 takes into account empirical quality factors and reflects a common denominator for new transmitter performance. This standard is a valuable reference document which describes performance parameters, standards and methods of measurements.

A thorough proof-of-performance at the time of installation is an invaluable record of normal operating performance. It also serves as verification of proper signal quality and emission standards. The number of detailed measurements required after the "proof" usually can be limited since several performance characteristics are a measure of the same impairment. Also some test waveforms are more useful for transmitter adjustment than others.

Transmitter measurements can be broken down into two primary categories as follows: frequency response and linearity.

Performance dependent on frequency response and group delay are sometimes referred to as linear distortions. That is, these distortions are not signal level dependent. In the time domain 2T, modulated sine-squared pulses, and multiburst waveforms are used in identifying and correcting distortions. In the frequency domain, swept amplitude and group delay measurements can be used but are normally reserved for out-of-service testing.

Linearity performance includes distortions which are signal level dependent and are a function of the instantaneous luminance level and on the average picture level over several lines. For example, level dependent chroma gain and phase are called differential gain and phase respectively. Luminance gain variation called low-frequency nonlinearity, and frequency response versus brightness which is the change in swept response with static luminance level are other level dependent distortions. Nonlinear responses can also change as a function of average picture level. To test for this, one line of the video waveform is alternated with four lines containing a static luminance level. Fig. 98 contains a modulated staircase with three average picture levels.

Modern transmitters are designed for unattended operation for extended periods of time. A transmitter operating with adequate cooling, regulated power lines and properly adjusted, may need only be checked in detail every three or four months or whenever a major component is replaced or repaired.

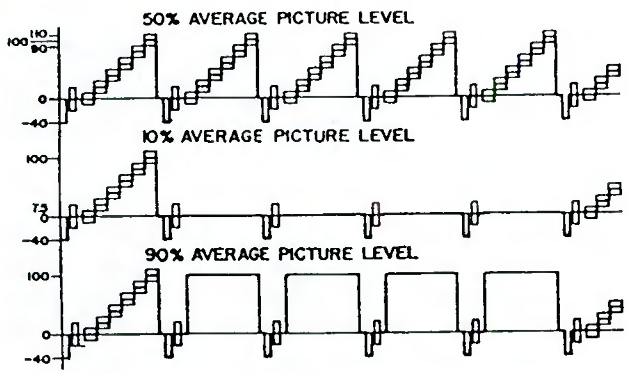


Figure 98. Modulated staircase waveform.

The following pretest checklist and sequence of tests are presented as a general guide for a properly adjusted transmitter.

Pretest Checks:

- Test equipment
- Input video
- Output transmission line, station load, and antenna
- AC mains input
- Record meter readings, adjustment settings, and key performance parameters

Transmitter Test Sequence

1. Exciter linear and nonlinear performance check.
2. Intermediate and driver linear amplifier performance check.
3. Swept frequency response.
4. Modulation depth (Include UHF pulser and sync ICPM adjustments).
5. Power output meter calibration.
6. Nonlinearity response checks. (Differential gain, ICPM, differential phase, etc.)
7. Linear response checks. (Group delay, pulse tests, composite waveform K factors, etc.)
8. Record meter readings, adjustment settings, and key performance parameters.

Many adjustments are interrelated and require returning to previous tests to verify proper performance. The sequence above is intended to minimize the number of adjustments.

At the time of a periodic measurement and adjustment sequence, it is a good practice to record meter readings and adjustment settings before and after the test. These recordings will indicate normal setting range and can aid in getting the transmitter back to near-normal in the event of mistuning errors during adjustments.

For off-air monitoring, a picture monitor is convenient for identifying gross video degradations quickly. Vertical interval test signals containing the composite waveform or other specialized test signals are useful for keeping track of depth of modulation, and several linear and nonlinear responses. These test signals are displayed on a waveform monitor and vectorscope. Records of transmitter meter readings and performance data are very useful in maintaining a transmitter at a high performance level. Maintenance logs should include date, time duration of outages, corrective action taken, and where possible, identify cause.

Monitoring TV Multichannel Sound

Maintaining a high quality aural signal requires high quality monitoring facilities. To be able to perform the proof-of-performance measurements is another important reason to have a monitor that demodulates the RF signal and separates the components in the composite MTS signal for analysis.

An ideal modulation monitor should consist of, but not be limited to, the following functions:

1. Demodulates the composite signal from the aural carrier.
2. Separates the components in the composite MTS signal for measurements.
3. Capable of off-air monitoring.
4. Covers all VHF and UHF channels.
5. Suitable for proof-of-performance use.
6. Contains a precision (BTSC) expander.

Equipment Needed

In order to make satisfactory measurements of video signals, a certain minimum of test equipment is required. Measuring the output of a television transmitter at baseband requires a precision television demodulator. For greatest accuracy, this demodulator should provide a zero carrier reference pulse for determining percentage of modulation, and ideally will have both a synchronous and envelope detection mode.

For most baseband measurements, a video waveform monitor is all that is needed, but to do a complete analysis of a system requires some additional equipment. The waveform monitor should include a provision for making measurements on vertical interval test signals (VITS) and provide filtering to allow separate examination of the luminance and chrominance components of color television signals.

In addition to the waveform monitor, a vectorscope must be used for making certain measurements on color signal components, particularly differential gain and phase measurements. The waveform monitor and vectorscope are all that are required to accomplish these measurements, but if a greater level of accuracy is desired in timing measurements, a conventional oscilloscope with an associated digital counter/timer would be a great asset.

PREVENTATIVE MAINTENANCE

A good preventative maintenance program should include periodic inspection and cleaning of the equipment. A vacuum cleaner is preferred to remove dust instead of compressed air which will simply blow the dirt into the air and let it fall back down on something else. A paint brush can be used to dislodge dust from delicate circuit boards. Avoid using a nylon bristled brush with a plastic handle as the static charge may damage CMOS or other static sensitive components. A natural bristle brush with wooden handle and metal binding is recommended.

High voltage wires and insulators must be cleaned with denatured alcohol, or other cleaner capable of removing the dirt without leaving any residue. Meter cases are cleaned with Glass Wax or other nonstatic cleaner.

Air filters should be replaced or cleaned as necessary to maintain adequate air flow to the equipment. A second set of washable filters can save time when using a single transmitter in critical service by quickly switching the clean filters and washing the dirty ones later. Blowers should be inspected to see if the curved fins are filled with any debris that would reduce air flow. Motor windings may collect a layer of dirt and interfere with the cooling of the motor itself. The fins of high-power tubes must be cleaned of any obstructions which may have passed through the air filters. Bearings should be lubricated and checked for excessive noise.

Color change in silver plated cavity parts can be a sign of over-heating and may require the disassembly of the cavity to check for obstructed air passages or loose connections. Set-screws in gear and chain drive tuning mechanisms should be checked for tightness. Black silver-oxide is a good conductor and need not be removed. Small parts should be dipped in Tarnix® for cleaning. Be sure to flush the parts after cleaning to remove any residue. Scotch-Brite® is a good nonmetallic cleaning pad for silver-plated parts. Remember that you do not want to remove the plating when cleaning these parts.

High current wires may move during turn-on surges and can suffer abrasions which may eventually cause an arc if not properly dressed away from sharp edges. Wiring on terminal boards may loosen through thermal cycling and vibration. All connections must be checked to be sure they remain tight. If wires need to be replaced, the correct gauge, voltage rating, and temperature rating must be considered when selecting the replacement.

Edge connectors on printed circuit boards should be cleaned with Cramolin® or other cleaner. A small amount of this cleaner is applied to the edge connector and then removed with a lint-free cloth. Do not use pencil erasers as this will remove gold or silver plating from the edge traces, and could degrade the connection or create an intermittent later as the sulfur in the eraser causes chemical reaction to the edge connector material.

Back-up systems or emergency modes of operation should be checked periodically. The worst time to discover trouble in the back-up equipment is when you need to put it on the air. Relays need some exercise to keep their contacts polished and in working order.

Transmitter site cleaning should also include a check of the building for such things as leaks in the roof which may cause damage to the transmitter, and presence of insects or small animals which may wander into unwanted areas.

Intake blowers with filters capable of creating positive pressure in the room can minimize the need for cleaning by keeping the dust out. A careful record must be maintained in order to establish a good history for future reference. Such a log will include a description of what was done, when it was done, and the name of the person performing the work. The question "How often must the transmitter be cleaned?", can be answered by "How often does it need cleaning?" Seasonal events such as harvesting in farm locations, severe weather, construction projects, in and around the transmitter building can require special action, but usually a pattern will emerge that allows the maintenance to be scheduled on a regular basis.

A complete set of meter readings taken when the equipment is working properly and updated weekly or monthly can greatly assist the engineer when trying to diagnose problems.

Create a maintenance program with weekly, monthly, quarterly, semi-annual, and annual tasks evenly spread throughout the year.

The following list of items can be used to develop a maintenance routine for any broadcast television transmitter facility:

Maintenance Items

Prefilter manometer readings
Inspect prefilters
Replace prefilters
Post-filter manometer readings
Cabinet input air manometer reading
Inspect transmitter air filters
Replace transmitter air filters
Vacuum cabinets
Clear tube fins or accumulated dust
Measure blower currents
Clean fan blades and motor windings
Connections checked for tightness
Inspection of MOVs

Recommended Data to be Recorded

Record all parameters on meters or user displays
Record DC input power and calculate dissipation (if applicable)
Record transmitter currents in black picture and at idle (no RF drive)

Visual Performance Checks

Ensure proper video level
Ensure proper sync level
Optimize differential gain
Optimize ICPM
Optimize differential phase
Optimize group delay using T pulses

Ensure proper power calibration
Ensure proper audio processor set-up
Optimize swept response

Aural Performance Checks
Ensure proper modulation levels
Ensure proper SCA input levels
Ensure proper power calibration
Ensure proper audio processor set-up
Optimize audio frequency response
Check and optimize distortion
Optimize stereo separation
Check and optimize crosstalk between MTS channels

Control System Checks
Verify proper operation of all interlock circuits
Verify proper operation of all overload circuits
Verify proper operation of all control processes (VSWR foldback, filament timing, coax switches, etc.)

RF amplitude response as a function of frequency is the variation in gain over the frequency range of the channel. The use of a television sideband analyzer or other frequency selective voltmeter provides a suitable means to measure a television transmitter amplitude response versus frequency.

At one time or another, just about every component part of a video signal must be measured or adjusted. It is good engineering practice that each part of a transmission plant including studio, distribution, transmitter, and monitor equipment provides minimal distortion to the video signal. Furthermore it is not recommended that one part of the system correct for another part except within a given system such as the precorrection circuits used in a transmitter to compensate for certain RF components. The signal going into the transmitter should be a standard NTSC video signal meeting its full specifications. The transmitter's correction circuits should not have to compensate for incoming video problems.

The overall amplitude of the signal and each of its component parts have strictly defined levels, and the relationship between the parts is also critical. Refer to RS-508 for a description of specific electrical performance parameters and standard test methods.

AIR SYSTEMS FOR TRANSMITTERS

Most electronic equipment which requires forced air cooling has the required blower or fan already designed into it. However, equipment such as transmitters, dissipate large amounts of power. Air which has already passed through the equipment and has picked up heat must be removed from the immediate vicinity of the equipment (exhaust air) in order to prevent the hot air from being recirculated. In addition to the air removal requirement, provisions must be made for sufficient make-up air (intake air) to replace that which has been circulated through the equipment and removed.

If the equipment user is to provide adequately for hot air exhaust and fresh air intake, the maximum and minimum environmental conditions which the equipment may operate in and the following information from the equipment manufacturer must be known:

Altitude: ______________________

Max. temp: ______________________

Min. temp: ______________________

Total air through the transmitter:

CFM ______________________

Pressure drop within the critical portion of the air circuit (across the transmitter tube for example. This is required for checking purposes after the system is built.)

Air pressure (Inches of water) ____________

Air temperature rise through the transmitter: ______

Air exhaust area: ______________________

Exhaust Air

Equipment layouts usually provide for heated air to exit from the top surface of the cabinet. The size and location of this exhaust area is usually shown on a manufacturer supplied outline drawing.

Most broadcast equipment internal air systems are designed to be operated into free space (back pressure of 0.0 inches of water) so any exhaust ducts must have minimum loss. A good practice usually is to design for no more than +0.1 inches of water pressure in the duct close to the exhaust area of the transmitter.

Any exhaust installation other than a large cross section duct (equal to the cross section of the transmitter exhaust port) with no bends, and with a long radius turn outside the transmitter building for weather protection, will need an exhaust blower or fan.

Keep in mind that the recommended system is sized only for cooling the transmitter and any additional cooling load in the building must be considered separately when selecting the air system components. The transmitter exhaust should not be the only exhaust in the room as heat from the peripheral equipment would be forced to go out through the transmitter.

The "sensible-heat" load is the sum of heat loads such as solar radiation, heat gains from equipment and lights and personnel in the area that is to be cooled.

The following hypothetical exhaust duct design illustrates the key cooling concepts:

Air flow (volume) through transmitter = 325 CFM (ft^3/min)

Air exhaust area = 3.4 square ft

$$\text{Air exhaust velocity} = \frac{325\ \text{CFM}}{3.4\ \text{ft}^2} = 94.5\ \text{ft/min}$$

The 94.5 feet per minute air velocity is relatively low which will allow a transition to a smaller diameter pipe if desired. Assume a transition down to a 10-inch diameter pipe, thus the following:

$$A = \frac{D^2}{4 \times 144} = \frac{100}{4 \times 144} = .545\ \text{square ft}$$

$$V = \text{CFM/A}$$

where: D = Duct diameter
A = Exhaust duct cross-sectional area
V = Exhaust duct air velocity

$$V = \frac{325\ \text{CFM}}{\text{dia. pipe } .545\ \text{ft}} = 596\ \text{ft/min air velocity in a } 10''$$

In a 10-inch diameter pipe the air friction chart in Fig. 99 gives 0.06 drop per 100 ft of pipe with a flow of 325 CFM. Assuming for this hypothetical design there is 20 ft of straight pipe to a roof, with two 90° elbows to turn the pipe down for weather protection, the total loss of the exhaust system may be estimated as follows:

Fig. 99 shows loss is 0.06"/100 ft. Since 20 ft is 0.2 of 100 ft,

0.2" × 0.06" = 0.012" of water pressure drop in 20 ft of pipe

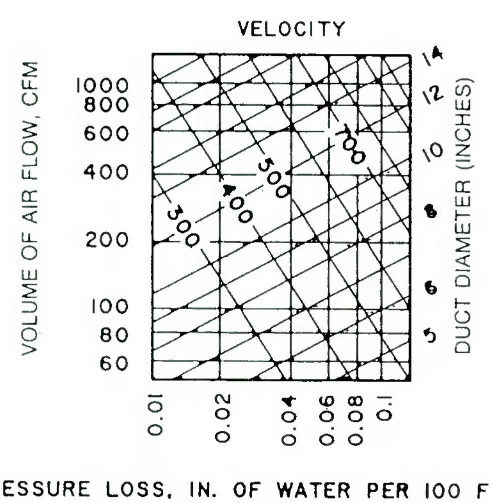


Figure 99. Friction loss in pipes.

Considering the static pressure drop of two 9[illegible] elbows, to give a 180° bend in the pipe, it is found from Fig. 100 that a 10 inch elbow at 900 ft/min gives less than 0.01 pressure drop, and there is just under 600 feet per minute in the system. Adding the two drops (one for each 90° elbow) and the 20 ft section of 10 inch pipe together, the result is:

0.012" + 0.01" + 0.01" = 0.032" of water total pressure drop

Therefore, no exhaust fan is necessary as the 0.032 inch of pressure is less than the 0.1 inch pressure level that requires a fan.

If the installer has a problem in exhausting the transmitter in this simple fashion (for example a roof exit is not available), and it is required to add two additional 90° elbows and a straight run of 10 inch pipe 100 ft long, the pressure drop in the hypothetical design exhaust system will be:

Friction loss in 20 ft of 10" dia. pipe	=0.012
4 elbows × .01" (water) each	=0.04"
Friction loss/100 ft of 10" dia. pipe	=0.06"
	=0.112" of water

The 0.112 inch of pressure now exceeds the level at which a fan is needed.

Because the transmitter manufacturer recommends no more than a 0.1 inch water pressure loss in an exhau[illegible] system, good engineering practice indicates that a[illegible] exhaust fan is now required in this configuration. The performance curve on the fan shown in Fig. 101, indicates that it will deliver 325 to 330 CFM into 0.1 inches of water back pressure. This is sufficient to handle all of the air coming from the transmitter and overcome all of the estimated duct losses in this configuration.

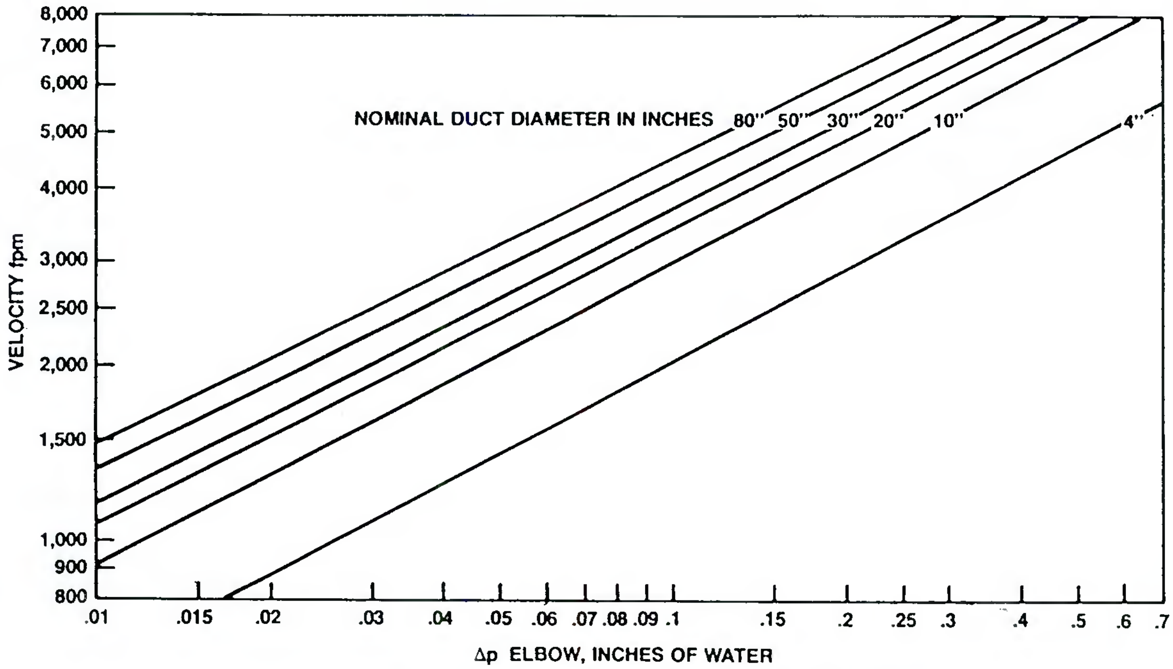
The outline drawing illustrated in Fig. 102 shows a typical exhaust duct and blower system. The recommended minimum ceiling height to properly handle exhaust air as shown is 12 ft The outline drawing also indicates a typical air intake and prefilter system.

Intake Air

An intake vent and blower should be sized to provide a desireable slight positive room pressure. Installing a manometer to sense pressure drop across the filters can help determine replacement interval of prefilters.

If existing space on site will not permit the construction of the transmitter manufacturer's recommended air system, care must be taken to modify the design to fit the available space and still properly cool the transmitter.

Referring again to Fig. 101, examination of the curve shows that at 325 CFM delivery, this fan will develop a static pressure of 0.1 inch water which is sufficient to overcome the less than 0.01 inch pressure drop of a 10-inch diameter 90° elbow installed on the outsid[illegible] wall of the building for a weather hood, as well as[illegible] filter and screen to clean the incoming air.



"ZERO LENGTH" LOSS, ADD TO DUCT FRICTION CALCULATED FROM INTERSECTION OF DUCT CENTERLINES.

Figure 100. Friction loss in 90° elbow.

20 ft of pipe	= .012″
2 elbows at .01″ each	= .02″
	.032″ total loss, maximum.

Additional flushing air is recommended for the removal of heat from any surrounding equipment that shares space with the transmitter. It is a recommended guideline to keep input air no greater than 5° C above ambient.

The above calculations are a very simple example of an air system. Most television transmitters, both UHF and VHF, have multiple enclosures with large volumes of air required. In these cases the equipment user is required to combine the heat exhaust air from several enclosures into a complex duct system. When these circumstances occur, the safest practice is for the user to contract for the services of a heating, ventilating, and air-conditioning (HVAC) consultant or company. The cost involved in having the best possible air exhaust and supply system will pay for itself in extended transmitter life and lower service costs.

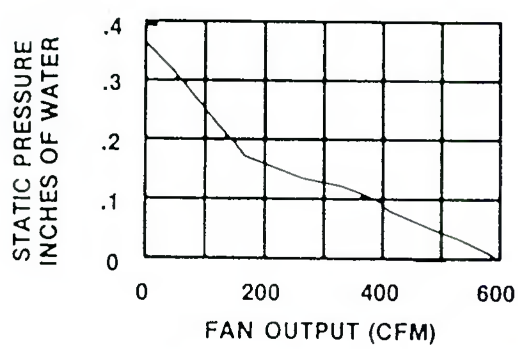


Figure 101. Vaneaxial Caravel fan performance curve.

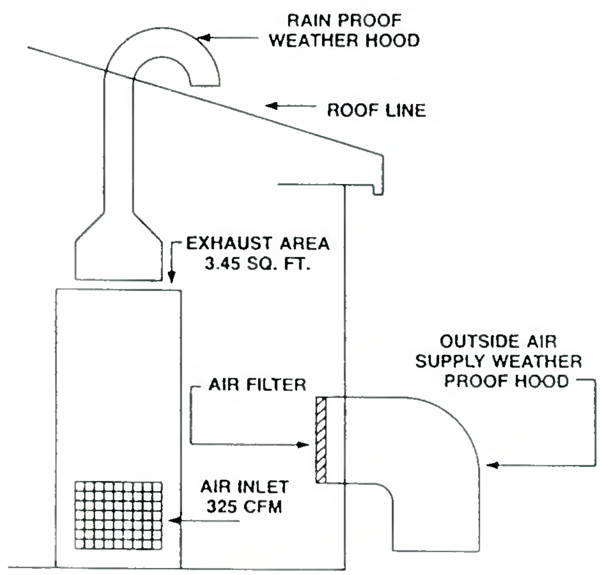


Figure 102. Suggested intake and exhaust duct installations.

Air Conditioning

It is a common practice to set the transmitter into a sealed wall to produce a plenum chamber and supply it with outside air, while providing separate air conditioning for the front side to cool personnel and source equipment.

In areas with severely polluted air, it may be necessary to run the transmitter on air-conditioned air to avoid bringing in corrosive salts or gaseous contaminants.

The amount of air conditioning will depend on several factors. It is strongly recommended that the air conditioning be shared by a number of units rather than one large central system so that operation can continue in the event of the failure of one unit. Air conditioning units are usually listed by "tons" of cooling capacity with "one ton" equal to 12,000 BTU per hour.

Again, consult experienced professionals in the area of HVAC design to achieve the desired result and to prevent problems from developing in the future.

ENDNOTE

1. Based on the stereo and auxiliary sound system recommended by the Broadcast Television Systems Committee (BTSC) of the Electronics Industries Association. See Chapter 3.6, "Multichannel Sound." of this Handbook.

BIBLIOGRAPHY

Weirather and Hershberger, "Amplitude Bandwidth, Incidental AM, and Saturation Characteristics of Power Tube Amplifiers for FM", IEEE Transactions on Broadcasting, March 1983, Vol BC-29, No. 1, p. 14.

W.L. Behrend and R.L. Rocamora, "Television Transmitter System Consideration for Multichannel Sound", EIA report on Multichannel Television Sound, Vol. 2A, October 31, 1983, Appendix D.

Electronics Industries Association. RS-508 "Electrical Performance Standards For Television Broadcast Transmitters".

Hans Schmid. "The Measurement Of Short-Time Waveform Distortion In NTSC TV Facilities." IEEE Trans. Broadcast, Vol 18, No. 3, Sept., 1971.

Charles Rhodes. "The 12.5T Modulated Sine-Squared Pulse For NTSC." IEEE Trans. Broadcast, Vol 18, Mar., 1971.

Pieter Fockens, Carl Eilers. "Inter-Carrier Buzz Phenomena Analysis And Cures." Transmitter Consumer Electric Vol 27, No. 3, August, 1981.

Bailey Neal. "Frequency And Amplitude Dependent Phase Effects In Television Broadcast Systems." IEEE Trans. On Consumer Electronics, Vol 23, No. 3, August, 1977.

T. M. Gluyas. "Influence Of RF Output Systems On TV Transmitter Picture Quality." NAB Convention Record, April, 1971.

L. E. Weaver. "The Multipulse Waveform." Application Note No. 30, Tektronix Corporation.

W. A. Resch. "Visual Performance Characteristics Which Affect Multichannel Sound." NAB Convention Record 1982.

W.L. Behrend. "Effects Of Incidental Phase Modulation Of TV Transmitters Or Other Circuits, On TV Signals." IEEE Trans On Broadcasting, Vol 19, No. 3, September, 1973.

Varian Associates, Inc. "Integral Cavity Klystrons for UHF TV Transmitters."

Varian Associates, Inc. "External Cavity Klystrons for UHF TV Transmitters."

Gregory L. Best. "Transmitter Conversion for TV Multichannel Sound." NAB Convention Conference Proceedings, 1984.

United Sheet Metal Bulletin 100-5-178 "Engineering Design Manual for Air Handling Systems".

Rotron Incorporated. "The Caravel Fan" Catalog Sheet E-3050C.

McGraw-Hill. Marks's Mechanical Engineers Handbook.

Robert Plonka. "Group Delay Corrector For Improved TV Stereo Performance." NAB Convention Conference Proceedings, 1989.

N. Ostroff, A. Whiteside, L.F. Howard. "An Integrated Exciter/Pulser System for Ultra High-Efficiency Klystron Operation." NAB Convention Conference Proceedings, 1985.

N. Ostroff, A. Whiteside. "Using Klystrode® Technology to Create a New Generation Of High Efficiency UHF-TV Transmitters." NAB Convention Conference Proceedings, 1986.

N. Ostroff, A. Whiteside, A. See, R.C. Kiesel. "A 120 KW Klystrode® Transmitter For Full Broadcast Service—The "SK" Series." NAB Convention Conference Proceedings, 1988.

N. Ostroff, A. Whiteside, A. See, R.C. Kiesel. "Klystrode® Equipped UHF TV Transmitters: Report On The Initial Full Service Station Installations." NAB Convention Conference Proceedings, 1989.

N. Ostroff. "Recent Advances In Klystrode® Equipped Transmitters." NAB Convention Conference Proceedings, 1990.

P. C. Turner. "Recent Developments In Solid-State TV Transmitters." NAB Convention Conference Proceedings, 1989.

G. Badger. "The Klystrode®, A New High-Efficiency UHF TV Power Amplifier." NAB Convention Conference Proceedings, 1985.

Donald Priest, Merrald Shrader. "The Klystrode®—An Unusual Transmitting Tube With Potential For UHF-TV." Proceedings of the IEEE, Vol. 70, No. 11, Nov. 1982.

George Badger. "Klystrode® Technical Performance For Modern High-Efficiency UHF-TV Transmitters." International Broadcasting Convention Proceedings, 1986.

Merrald Shrader, Donald Priest. "Klystrode® Update." IED Meeting, San Francisco, CA, December 1984.

Michael Chase. "The 15 kW Klystrode®: An Efficient Air Cooled Device for Television Broadcast Service." Engineering Conference, November 10-12 '87.

Robert Weirather. "A Distributed Architecture For A Reliable Solid-State VHF Television Transmitter Series." NAB Convention Conference Proceedings, 1989.

Martyn Horspool. "Designing An All Solid-State VHF TV Transmitter For High On-Air Reliability." International Broadcast Engineer, May 1989.

G. L. Best, F. A. Svet. "Architecture For A High Reliability VHF Transmitter." IEEE Broadcast Symposium, Sept. 1988.

Ulrich Gysel. "A New N-Way Power Divider/Combiner Suitable For High Power Applications." IEEE-MTTS-5, International Symposium Digest, 1975.

Gerald Collins. "Combiner For Platinum Series TV Transmitter." Harris Corporation Technical Paper, 1989.

R. S. Symons. " Depressed Collector Klystrons For High-Efficiency UHF-TV." Varian Technical Report, 1986.

Earl McCune. " MSDC Klystron Progress Report." SBE National Convention Proceedings, 1989.

James Pickard. "Field Performance of a Multiple Stage Depressed Collector Klystron." NAB Convention Conference Proceedings, 1990.

Greg Best, James Pickard, David Danielsons. "Subjective Tests on A Switchless Combiner TV Transmitter System." NAB Convention Conference Proceedings, 1987.

Electronic Industries Association. "BTSC System Recommended Practices." EIA Systems Bulletin No. 5, July 1985.

ANSI/IEEE C62.41 Standard (Sometimes Referred to as the IEEE-587 Standard).

George M. Badger, The Klystrode®, A New High-Efficiency UHF TV Power Amplifier. National Association of Broadcasters, Las Vegas, Nevada, 1985.