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Television Transmitters

Preface

Chapter 19 is comprised of two parts—the original Chapter 19 and an addendum which, together, constitute a comprehensive discussion of current domestic television transmitters.

Television transmitters in use in the United States today can be classified, broadly, into two basic types. One type, described in the original chapter, employes modulation at an IF frequency followed by a low-level IF vestigial sideband filter, an upconverter to final frequency at very low power, and a chain of linear amplifiers on the final frequency to reach the desired power level. A different type described in the addendum, employes modulation at some moderate RF level followed by linear power amplifiers and a high-level vestigial sideband filter at the transmitter output.

The unique circuits and special characteristics of each type is discussed in approximately equal depth in the two portions of Chapter 19. The original chapter includes much general information on television transmitters that applies to either type, and this information is not repeated in the addendum.

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Television Transmitters[1]

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The following chapter is written to provide the operating engineer with a better understanding of the functioning of his equipment. It is not the intent to provide sufficient information to design the equipment.

Emphasis and more detail is offered in areas thought to be new and not readily found in easily obtainable references. Areas which are long established, widely understood, with readily available references, are treated in less detail.

A television transmitter will accept a video signal at its input terminals and produce an amplitude modulated carrier of proper power in which the video information is contained. It also accepts an aural signal at a second input terminal and produces a frequency modulated carrier of proper level. It will generally include elements to combine visual and aural output signals into one transmission line.

Performance standards, power levels, and frequency assignments for TV transmitters are regulated by FCC Rules and Regulations (Volume III, pt. 73) which in many areas conform with recommendations made by the CCIR (International Radio Consultative Committee).

TV broadcast services operate on the following frequency bands:

54 to 88 MHz (VHF-Low Band) Channels 2-6
174 to 216 MHz (VHF-High Band) Channels 7-13
470 to 806 MHz (UHF) Channels 14-69

The channel width is 6 MHz and all VHF channels are exclusive assignments, i.e., they are not shared with other services. Some UHF channels are shared with land mobile services.

The system in the USA (and some other countries) is designated by the CCIR as "System M."

The limited number of available channels requires assignment of the same channel to broadcasting stations in different areas. In March 1974 almost 600 VHF stations were licensed, which would indicate that on the average each VHF channel is assigned 50 times.

To reduce interference between stations, the FCC limits the effective radiated power (ERP) of every station on Channels 2-6 to 100 kw, on Channels 7-13 to 316 kw and Channels 14-69 to 5000 kw. The ERP equals transmitter output power multiplied by the antenna gain (neglecting transmission line losses for the moment) and leads to the following predominant power ranges of TV Transmitters (Visual Transmitter):

VHF, Low Band:	100 watts to 35 kw
VHF, High Band:	100 watts to 50 kw
UHF,	1 kw to 220 kw

The majority of VHF stations operate at a transmitter output of approximately 75% of the high limit shown above with lower power operations predominantly serving small, isolated communities. At UHF very few stations operate near the high limit shown, with the majority operating from 30 to 60 kw transmitter output power.

The RF output power of the aural transmitter ranges from 10% to 20% of the visual transmitter power. The output power of the visual transmitter is the peak power transmitted during the synchronizing pulse. The average output power of the visual transmitter varies with picture content from approximately 20% for a completely white picture to 60% for a totally black picture.

Transmitters for VHF service employ gridded tubes in the power stages while most UHF transmitters use klystrons in the power stages. Some UHF transmitters with output powers of 10 kw and less use gridded tubes in the power stages.

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[1]Superscript numbers refer to references at the end of this chapter.

Acknowledgment. The author wishes to thank Amperex, Eimac, and RCA for furnishing the information contained in Table 1. He also wishes to give credit to the authors of the paper cited in Reference 15 for allowing the use of several figures from their paper. Mr. Fred W. Haushalter provided valuable assistance in writing about mechanical and cooling aspects of a TV Transmitter and credit is due Mrs. Madelyn S. Wade and Mr. Francis W. Wentura for their help and patience in the preparation of this manuscript.

PERFORMANCE REQUIREMENTS

Visual Transmitter

The visual transmitter is offered a video signal from a 75 ohm coaxial cable at 1 volt peak-to-peak. This signal contains all synchronizing and picture information and can be expected to show significant spectral components from dc to about 6 MHz.

It is the transmitter's task to amplitude modulate this signal to a carrier of constant frequency and to shape the amplitude response of this carrier signal to fit within the available channel. Otherwise the transmitter should be a linear device as far as voltage and phase are concerned (except for the prescribed receiver delay predistortion).

The frequency of the visual carrier must, under FCC Rules, be within ±1 kHz of the assigned value. If a group of stations operate under precision offset conditions to reduce co-channel interference, only ±2 Hz departure from the correct carrier frequency is allowable to maintain this advantage.

In order to fit all video information into the available channel space only a vestige of the lower sideband is transmitted. It is designed to essentially pass all information up to 750 kHz in a double sideband fashion and attenuate all lower sideband components over 1.25 MHz by at least 20 dB, except the lower color subcarrier sideband which must be attenuated at least 42 dB.

The upper sideband should be essentially flat to 4.18 MHz, but reject components over 4.75 MHz by 20 dB or more. The requirements of the FCC Rules and Regulations for amplitude and sideband response present only a small challenge to the present-day transmitter and a more typical response tolerance pattern is shown in Fig. 1.

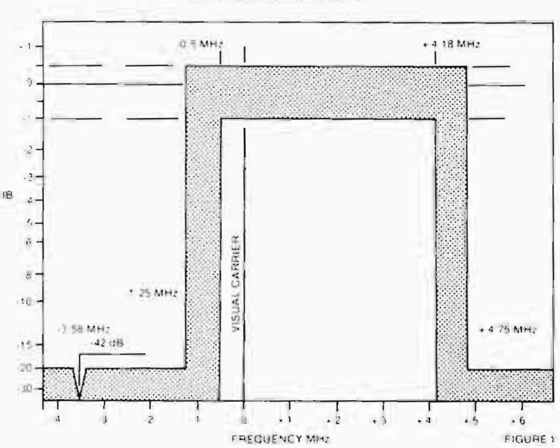


Fig. 1. Frequency response of modern TV transmitter.

Group delay through the passband from −0.75 MHz to +3 MHz of carrier should be flat. Most modern equipment exceeds the ±100 ns tolerance limit (.02 to 2.1 MHz) of the FCC Rules by a factor of 2. From 3.00 to 4.18 MHz a chroma advance is required of the transmitter with a straight line increase to place +170 ns at 3.58 MHz. The tolerance narrows from ±100 ns at 2.1 MHz to ±50 ns at 3.58 MHz and from there opens up again to reach ±100 ns at 4.18 MHz. It is highly desirable to provide a bypass switch for the circuit producing this FCC receiver predistortion in order to operate the transmitter in a "flat delay" mode for measurement purposes since no demodulators are available at this time which show a group delay behavior exactly complimentary to the one imposed upon the transmitter.

One will find, in a high power transmitter, that group delay varies with the picture level just as amplitude response varies with picture level.

If delay (or phase) of the carrier varies relative to the color sideband, we speak of differential phase. If, by chance, both vary by the same amount and direction, no differential phase could be detected even though the transmitter shows deficient performance.

Such shift, however, would readily be detectable in the aural output of an intercarrier receiver (limited of course to the rate of change which can pass the aural detector).

It is imperative, therefore, to specify allowable phase shift of the visual carrier caused by video modulation in order to preserve a minimum signal-to-noise ratio in the aural channel. No FCC limits are presently established, however, one may use the method and limits recommended by EIA or CCIR.

No further specifications, speaking in a fundamental fashion, in the frequency or time domain are required. However, it has become customary to make certain dynamic measurements which pinpoint weaknesses in the frequency or time domain under varying picture levels. Differential phase measurements using a stairstep or color bar to define phase changes of the color subcarrier over one line, and relative to burst, fall into this category. The FCC Rules allow ±10° of phase departure from the ideal (relative to the color burst) when 75% saturated primary colors are transmitted.

Since the amplitude response very close to the RF carrier cannot be properly determined using standard sweep techniques, observations in the time domain make up for this shortcoming. Improper transmitter operation in this area manifests itself by disturbances during the vertical blanking period (for very low frequency problems) and disturbances of the bar signal become visible for somewhat higher frequency imperfections.

From observations made in the time domain, changes of frequency determining elements can be made to correct for the above deficiencies. Line and/or horizontal rate squarewaves are suitable test signals by themselves or in conjunction with other test waveforms sharing a frame or line.

Other test signals, e.g., the 2t pulses and the modulated 12.5t and 20t pulses must also be viewed as measuring convenience signals since any limitations placed by them on the transmitter specifications are redundant once amplitude response, group delay, and transfer linearity are properly defined.

They also serve an educational purpose to emphasize the need to assess performance in the time domain rather than the frequency domain.

Therefore, if one is given a choice, resulting from certain system imperfections, to be only able to meet frequency response or acceptable wave shape of 2t and 20t pulses, one should choose the latter. There is no conflict in physical laws in such a possibility. The transmitter could, e.g., show a pronounced dip in response of say 2 dB at 2.8 MHz, reaching −.5 dB at 3.0 MHz and at 2.6 MHz, bringing it outside of the response shown in Fig. 1, yet allowing nearly perfect 2t and 20t signals to pass through the transmitter. An equipment designer would, of course, never rest at this point without finding the cause for the response disturbance, but an operator may be forced to place equipment with the above defect on the air and postpone the search for the cause. In such a case, pulse or transient behavior should overrule behavior in the frequency domain. Judgment is required on a case by case basis, which only a well-trained operator can provide.

It should be stressed that modern equipment generally meets the above limits irrespective of picture level (black, gray, or white picture), and over an output power range of at least 80% to 100% of rated power without the need for transmitter readjustment.

Specifications of the voltage transfer characteristic define the permissible departure from an ideal straight line for various signal conditions. Two dynamic measurements generally provide information of transfer behavior: a ramp signal for low frequency components and differential gain for one or more higher frequencies. Again, if the time or frequency domain characteristic of the transmitter were totally independent of picture level, the above requirement would be redundant. But as a practical matter, the determination is a useful one.

The FCC Rules do not define low frequency linearity, only differential gain. A tolerance of 20% from ideal is permissible. Modern equipment far exceeds this requirement.

There are several specifications controlling changes in the transfer characteristic based on changes of picture level; e.g., variation of peak power from black to white picture. The FCC Rules summarily lump many elements relating to changes in the transfer function into an overall limit of 5%, while expressing hope for a better definition in the future.

As in any communications channel, detrimental noise is generated in the system. Various methods exist to assess the detrimental effects of such noise. Some methods provide for noise weighing to take the degree of visibility into account. No such recommendations exist at this time for FCC Standards. It has, therefore, become the practice to view noise without weighing and to measure either RMS or peak voltages.

The only domestically manufactured noise measuring device measures noise by optical comparison on the waveform monitor, and, therefore, is defined as an average reading. It assumes a reference of 0 dB at 0.7 volts. In the past, transmitter noise specifications have been referred to a sine wave voltage which would modulate a transmitter from zero carrier to peak-of-sync (100% AM). Using this reference, most modern TV transmitters will meet a 50 dB requirement (RMS Measurement).

It appears that the least ambiguous approach to measure signal-to-noise in a TV transmitter would be to measure peak-to-peak noise over a defined frequency range through a suitable weighing network and refer it to the black to white transition. The same standard would be applicable to other elements of the system; for instance, demodulators, cameras, tape machines, etc.

The slight "disadvantage" of this latter method is that it produces smaller numbers than the presently employed methods, with the inherent danger of making better equipments look poorer when compared to old equipments characterized with the older method of measurement. A complete set of manufacturer's specifications are shown with typically commercially produced equipment towards the end of this chapter.

Mention should be made of the excellent tool for transmitter and system observation which is now widely available in the form of VIT and VIR signals, allowing continuous monitoring of most significant characteristics.

Only FCC Rules and Regulations are mentioned so far since the operator, by law, is only bound by them.

Yet a very useful additional standard exists: EIA, RS-240. This standard is more encompassing than FCC Rules and provides additional guidance by defining the method of measurement. Efforts are underway to revise and update both standards to bring them in line with present practices.

Aural Transmitter

The aural transmitter receives an input signal of +10 dBm from a symmetrical (balanced) 600-ohm line. It contains spectral components from 30 to 15,000 Hz. This information in turn is frequency modulated onto a carrier causing ±25 kHz carrier deviation when 400 Hz is present at +10 dBm. Seventy-five microsecond pre-emphasis is applied to the audio signal to improve system noise performance, a standard practice with FM broadcast systems.

The frequency of the aural carrier is specified to be 4.5 MHz above the visual carrier with a tolerance of ±1 kHz. Power of the aural transmitter may be between 10% and 20% of the peak-of-sync power of the visual transmitter.

The requirement for amplitude response under the FCC Rules and Regulations can be satisfied by a signal whose response can be fitted into a window 3 dB wide from 100 to 7,000 Hz with the signal allowed to drop to 4 dB at 50 Hz and 50 −5 dB at 15,000 Hz. The upper edge of this window is defined by the 75 microsecond pre-emphasis curve.

The voltage transfer characteristic is defined in terms of harmonic distortion of the input sine wave, allowing one-half of the system distortion for the transmitter proper. This will then establish limits of: 1.75% from 50 to 100 Hz, 1.25% from 50 to 7,500 Hz, and 1.5% for frequencies from 7,500 to 15,000 Hz. De-emphasis may be applied during distortion measurements, and all harmonics up to 30 kHz must be included in the measurements.

The FM signal-to-noise ratio including all spectral components from 50 to 15,000 Hz measured after de-emphasis should exceed 56 dB relative to ±25 kHz deviation by 400 Hz.

Since the FM carrier passes through selective circuits, a degree of FM to AM conversion will take place. No limits are established by FCC Rules and Regulations, but other administrations prescribe limits of −40 dBm relative to 100% AM for 25 kHz deviation and term this deficiency "synchronous AM."

If, in an FM system, the modulation transfer characteristic contains even order distortion, the mean frequency may experience one-sided shift with modulation. It is implied in the FCC Rules and Regulations that shift must not bring the mean (center) frequency outside the ±1 kHz tolerance, yet some administrations separate center frequency shift with modulation from carrier long-term drift.

The different treatment is largely brought about by traditionally different methods of controlling the center frequency of aural transmitters. In the U.S.A. where direct crystal control (e.g., serrasoid modulator or FMXO), or phaselocked loop systems are predominantly employed, center frequency shift with modulation is practically nonexistent as opposed to AFC type controls predominantly used in European countries which can show substantial center frequency shift with modulation.

It is questionable whether transfer linearity measurements presently defined in terms of harmonic distortion should be replaced by intermodulation measurements (acoustically a more meaningful quantity) as it would require every user to change his instrumentation. Once the new test equipment is available however, only a single measurement will be required to determine intermodulation distortion as opposed to a series of measurements for different frequencies with harmonic distortion. Furthermore, intermodulation measurements disregard frequency components falling outside the audible range. These components can be neglected.

Visual and Aural Transmitters

Every transmitter will produce spectral components appearing at its output terminals, or as radiation from its cabinet which are undesirable. These components become increasingly obnoxious as the density and utilization of the spectrum increases, with enormous differences in operating power of systems assigned to adjacent frequency bands.

The unwanted radiation consists largely of components harmonically related to the carrier, but sometimes may contain components not harmonically related to the carrier.

No modern transmitter should radiate signals created through parasitic oscillation of active devices. For this reason, every spurious component detected during measurements must be clearly identified for its frequency relationship (harmonic or intermixing) to the visual and aural carriers and any other primary frequency (e.g., the color subcarrier frequency).

An extremely useful practice in facilitating this identification is to modulate the primary signals (the visual and aural carrier) with identifying waveforms using extremely low modulation percentages to avoid creating additional sidebands. The detecting device must be able to produce a demodulated output from amplitude or frequency modulated signals or a combination thereof.

Industry standards generally call for unwanted signals in the output transmission line of 80 dB below carrier at peak of sync power. The FCC Rules and Regulation presently require spurious components to be attenuated at least 60 dB. Equipment can be produced without great difficulty in which spurious components are attenuated from 90-100 dB below carrier. One should remain suspect of receiving and monitoring devices since large input signals appearing in these devices may regenerate unwanted components in magnitudes considerably greater than the magnitudes appearing in the output of the transmitter.

Among emissions generated by a television transmitter one must also take into account acoustical emissions and x-rays. As a guidance, acoustical noise should be approximately 80 dB or less (remembering that in acoustics the dB is an absolute measure—not a relative one). Proof of compliance to this specification is quite difficult as measurements should be made in an anechoic chamber. Measurements made with the equipment sitting in a factory environment can only be considered for guidance purposes, as acoustical standing waves will materially change sound levels. Furthermore, a transmitter is a highly nonuniform and very directional sound source making levels highly dependent on the location of the sensing devices.

Similar reasoning applies to x-rays but in this area one must take the approach that anywhere outside the cabinet any detectable level of radiation must be comfortably within safe values.

The term "radiation" includes electromagnetic radiation as well as accelerated particles. High power transmitters, because of high anode voltage potentials, may generate gamma rays (x-rays). The presently accepted maximum level is .5 milliroentgens per hour anywhere along the outer surface of the cabinet, but it appears that further recommendations will place this level at .25 mR/hr.

It is not difficult in properly designed equipment to meet this requirement, especially in VHF transmitting equipment due to its lower anode potential as opposed to UHF klystron amplifiers which employ collector potentials up to 25 kv.

DESIGN PHILOSOPHY

The Engineering profession should accept a fundamental challenge in its design of any machine:

1. It should accomplish its assigned task with the greatest fidelity necessary.
2. It should do it with the best efficiency possible within an economic framework.

A measure of fidelity has been established under the heading "Performance Requirements" which state how well the process has to be performed without being carelessly tolerant or excessively stringent. Efficiency, in our commercial environment, is best measured by adding expenses and cost and replacement of capital.

Operating expenses include cost for operators, replacement parts, power consumption, repairs, etc. Capital cost include depreciation and interest on capital outstanding.

Engineering decisions should generally be made based on concrete information and numbers, but not on emotions and speculations. This, however, does not mean the engineer needs to be an emotionless or colorless person. Sometimes emotions will play a role in his decision, but he should be able to recognize when this is so.

One measure of efficiency, and in turn low maintenance and repair cost, is the statistic MTBF, (mean time between failure). The MTBF of complex equipment is the inverse of the sum of the failure rate of all components. Every effort must be made to find circuits requiring the least number of components consistent with stability and proper performance. One can expect that MTBF's for 50-kw TV transmitters may be around 500 hours. Therefore, the probability to get through an 18-hour broadcast day without any component failure is

$$R = e^{-t/T}$$

where R = probability of no failure (or reliability)
t = operating period
T = MTBF

$$R = e^{\frac{-18}{500}} = e^{-0.036} = 96.46\%$$

It should be remembered, however, that failure of a component does not necessarily cause a service interruption (e.g., an indicator light burning out is as much a failure, mathematically, as a short in the high voltage rectifier).

Consumption of electric power should be kept as low as possible. To aid in this direction air-cooled tubes with low-pressure anode radiators are preferable, all other things being equal over tubes showing high pressure drop across the anode radiator. It is more costly to produce high pressure air than low pressure air.

There are additional guidelines, e.g., adjustment for proper operation should, as much as possible, be done with built-in meters, components should be readily available; catalog items, component exchange should not cause undue equipment disassembly, equipment should be safe to operate, it should be self-protecting, etc.

TV TRANSMITTER FOR VHF SERVICE

Visual Exciter Modulator

The visual Exciter Modulator will generate a carrier at final frequency, modulated by video information and properly processed for sideband and delay shaping including moderate amounts of correction for linearity, differential phase and gain to make up for later stages in the transmitter. It's output power may be in the area of several Watts.

Fig. 2 shows a block diagram of a typical exciter-modulator.

In this type of exciter-modulator all signal processing is performed at video or intermediate frequencies including the modulation itself. After all processing is complete, the signal is raised to the final carrier frequency by an up converter and to the final power through suitable linear amplification.

The signal levels indicated on the block diagram are intended to show orders of magnitude only. It can be seen that most of the processing takes place around a level of 1 milliwatt, which is consistent with levels employed in most other video equipment. Only after the final frequency is reached is there gain to increase power levels.

The first task in the exciter-modulator using an IF modulation approach is the suitable selection of the intermediate frequency. The following criteria apply:

1. The frequency should be low enough to be advantageous for modulation and for processing.
2. It must not reach into the video range nor be so low as to cause image frequency or local oscillator separation problems in the required up converter.
3. The conversion process must not produce mixing products of intolerable magnitude within the final passband or mixing products outside the passband which may be unduly difficult to attenuate.

Intermediate frequencies have already been standardized for TV receivers: 45.75 MHz (in the U.S.A.), and 38.9 MHz (in Europe) for the visual carrier. The difference between the two standards is lodged in the different channel assignment for TV in the area from 50 to 90 MHz.

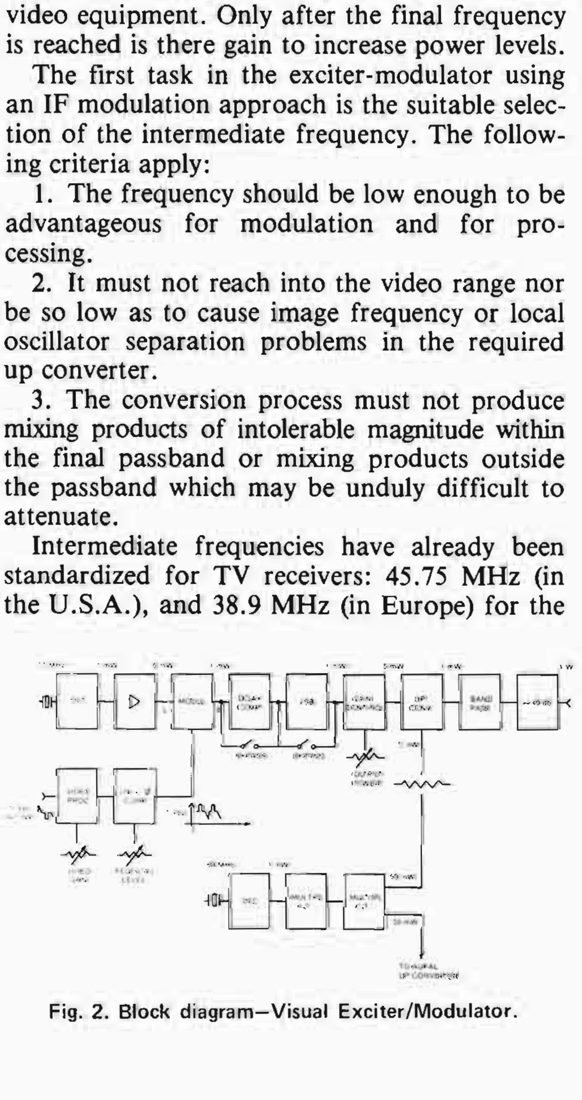


Fig. 2. Block diagram—Visual Exciter/Modulator.

For television transmitting equipment manufactured by the Broadcast Products Division of Harris Corporation an IF frequency of 37 MHz was chosen. Selection of this frequency relating to Items 1 and 2 above is largely a matter of judgment but careful analysis should prove the chosen frequency to satisfy Condition 3.

Similar conditions were encountered by designers of SSB communications equipment which uses IF Modulation processes exclusively.[1]

The mixing products investigation must first be conducted on the assumption that both input signals to the balanced mixer are inherently pure. This initial analysis quickly establishes that no in-band components exist for all UHF channels, that one in-band component exists for Channel 7 $\left(\frac{4 \times f_p + f_{IF}}{5}\right)$, a fifth order component with an expected magnitude of -100 dB relative to the wanted signal: $(f_p - f_{IF})$ but several in-band components exist in Channel 4 and one each for Channels 2, 3, and 5. The magnitudes of these in-band components are expected to be more than 100 dB below the carrier level of the wanted signal, except for one component in Channel 4 $\left(\frac{f_p + f_{IF}}{2}\right)$ at -80 dB below carrier producing a 3.405 MHz beat. Actual measurements confirmed the magnitude of this signal to be better than above indicated.

It is imperative to use a mixer circuit whose long-term stability is invariant and whose performance is insensitive to normally expected variations in the amplitude of the pump source.

Reference 1 shows samples of a graphical analysis for the determination of possible in-band mixing products.

The investigation must not stop at this point but be repeated for possible unwanted harmonic or subharmonic components which may be contained in the pump and IF inputs.

These secondary checks must be performed for all bands: VHF as well as UHF. The task is considerably simplified when the pump chain is designed with the highest possible crystal oscillator frequency, e.g., 60 MHz, which will of course reduce the number of possible harmonic and subharmonic components in the multiplier chain.

Again, a graphical analysis is feasible for subharmonic and harmonic components and it becomes obvious at once that low voltage levels of those spurious components are highly desirable. Fig. 3 shows the schematic of a multiplier

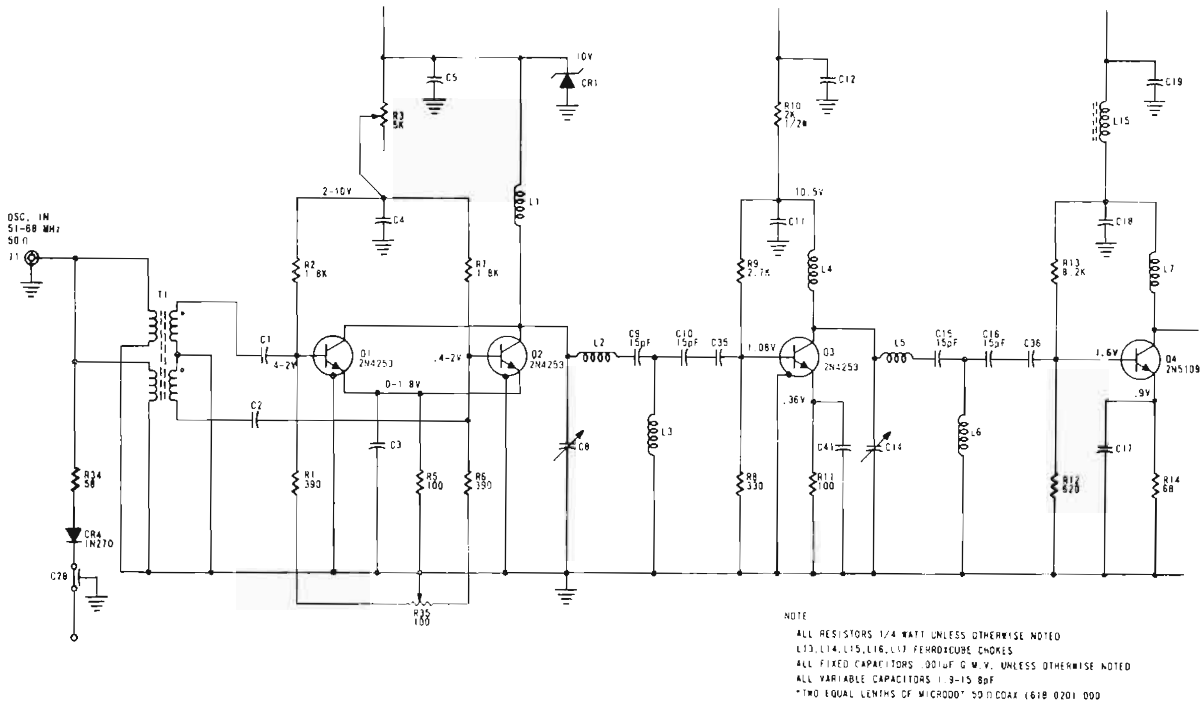


Fig. 3. Multiplier in pump chain.

chain specifically designed to maintain the level of unwanted subharmonics and harmonics at −50 dB relative to the wanted signal.

The selection of 37 MHz as IF for systems operating under FCC Rules and Regulations has one added advantage of being very close to the accepted European IF of 38.9 MHz, therefore, only one model of a visual sideband filter is required and can universally be used for any standard, any channel, and any CCIR system by allowing for a small amount of tuneability in the filter design. Fig. 4 shows a picture of this filter including a simplified schematic.

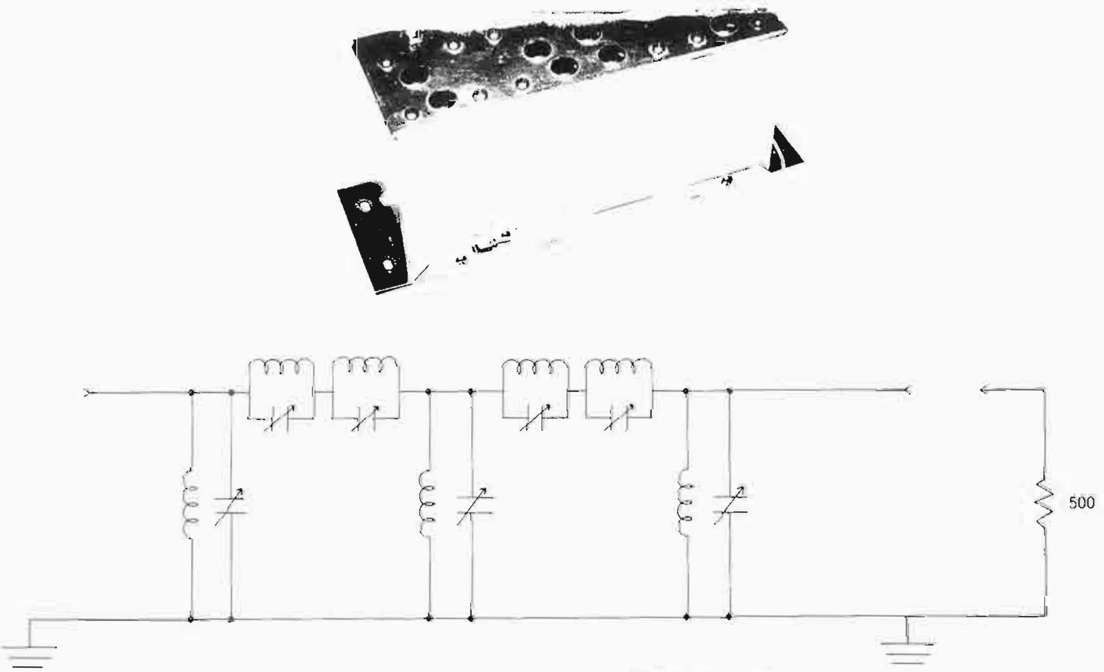


Fig. 4. VSB filter including simplified schematic.

After the choice of the frequency plan is confirmed to be without conflict, the selection of practical power level for processing must come next. Fundamentally, one should aim to work at as low a level as possible. This allows the design engineer to produce a more perfect system without paying penalties in performance by stressing active devices. The efficiency of the processing circuits, including the entire Exciter-Modulator, is of little consequence when compared to the whole transmitter. Therefore, emphasis should be placed towards attaining the most perfect signal possible with a minimum of circuit adjustments. Since the noise floor is predominantly established by active devices and cannot be altered, this level establishes the reference point upon which the choice of the signal level is based.

It was found that an operating level of 1 milliwatt (which is approximately 1 volt peak-to-peak in a 50-75 ohm system) would provide a generous margin in the signal-to-noise performance of the final processed signal and at the same time not unduly tax the requirement for voltage handling capability, linearity, etc., of transistors and integrated circuits using moderate supply voltages.

(This choice of level, it can be added, conforms to the level widely chosen for numerous types of video equipment, like recording devices, cameras, distribution amplifiers and the like.) Therefore, as can be seen from the block diagram Fig. 1, the power levels throughout the Exciter-Modulator are closely maintained around a level of 1 milliwatt.

RF Generator

The generation of the unmodulated RF frequencies (the IF carrier and the pump source) follows established principles and need not be dealt with in great detail. In both oscillators the crystals are operated at low power levels to avoid mechanical stresses which could possibly lead to destruction of the crystal. This precludes the need for built-in spares. In the past spare oscillators were often required when the crystals were operated at high power levels with a high probability of destruction.

Both oscillators are placed in ovens which are proportionately controlled for temperature and remain activated continuously. With this practice, frequency drift can be reduced to allow a UHF transmitter at the highest possible carrier frequency (the worst case) to remain well within tolerances prescribed in the FCC Rules and Regulations for at least 60 days. This means, of course, that the carrier frequency stability on Channel 2 is 10 times better than it actually needs to be.

Both unmodulated RF frequencies are available at proper levels and harmonic and sub-harmonic content is at least 50 dB below the level of the wanted frequency. To ascertain this level we again refer to Fig. 3.

Proper RF level is set through diode CR-4, followed by a balanced (Push-push doubler consisting of transistors Q1 and Q2 and a high pass filter to reject the fundamental frequency.) (See C9, C10, L3 and C15, C16, L6.)

Video Processing

Before going into actual circuits a policy should first be stated which may be subjected to debate but is defined as follows:

The transmitter should correct for its own problems but otherwise be transparent to the input signal. Nonstandard signals should not cause any damage to the equipment.

This policy does remove the burden of judging the correctness of the input signal from the transmitter to a processing amplifier if such judgment is required by other deficiencies in the system prior to the transmitter. Since the defects will vary with every system, correction outside the transmitter can better be tailored to individual requirements without placing the burden for the worst case on all equipments.

Following this policy, only four PC boards are required in the video portion of the Exciter-Modulator which will: (1) accept the input signal either single ended or with common mode rejection; (2) compensate moderately for the transmitter's own nonlinearity; and (3) will restore dc level of signal. A FCC receiver delay predistortion unit is provided in its own tray.

The sync negative input voltage is received by a differential amplifier offering 40 dB of common mode rejection. The main target is 60 Hz hum possibly appearing in phase on the inner and outer conductors of the coax line.

The FCC receiver delay predistortion is performed by active circuits possessing all-pass

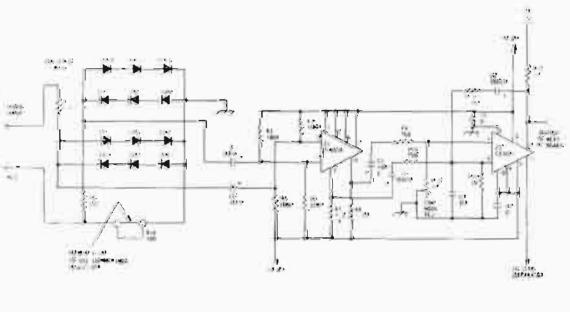


Fig. 5. Common mode rejecting and input protection.

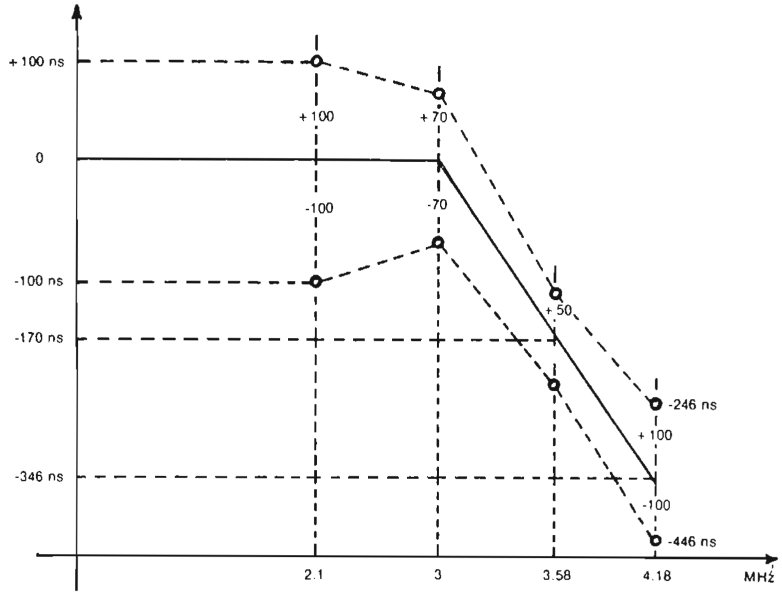


Fig. 6. Transmitter envelope delay.

characteristics. Four such all-pass networks are connected in series and will produce a variety of receiver predistortion curves under various transmission standards. The operating principle of an active all-pass will be described in more detail in the section dealing with "IF delay compensation." The common mode rejection capability is achieved by utilizing a differential amplifier in the input stage.

For protection against overvoltages of both inputs, the cable center conductor as well as the outside braid are clamped against the positive and negative supply voltages which should avoid damage to the input transistor. See Fig. 5. The allowable tolerance of overall transmitter envelope delay is shown in Fig. 6 which reflects paragraph 73.687(a)(5) of the FCC Rules and Regulations. The typical receiver should have an envelope delay characteristic complimentary to Fig. 6. However, no legal restraints have at this time been placed on receiver manufacturers due to a lack of jurisdiction of the FCC.

In the past, many delay compensators used passive devices connected as all-pass sections. Circuit configuration of the fundamental all-pass is shown in Fig. 7. The pole and zero placement of this all-pass is shown in Fig. 8 and typical delay of such an all-pass is shown in Fig. 9. The circuit shown in Fig. 7 is awkward to realize, in addition, it has no common ground connection. Under certain conditions, largely with restrictions placed on the values of com-

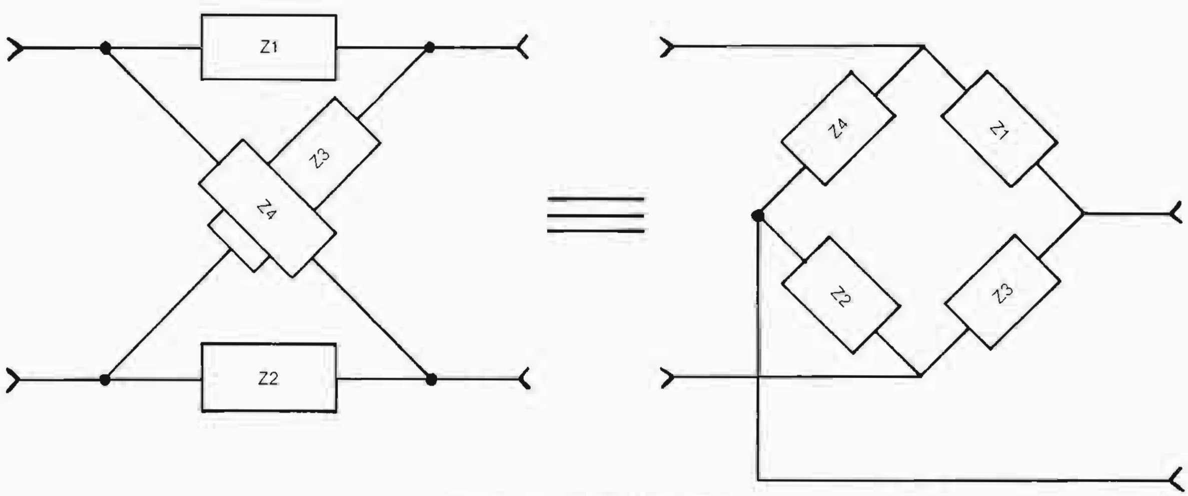
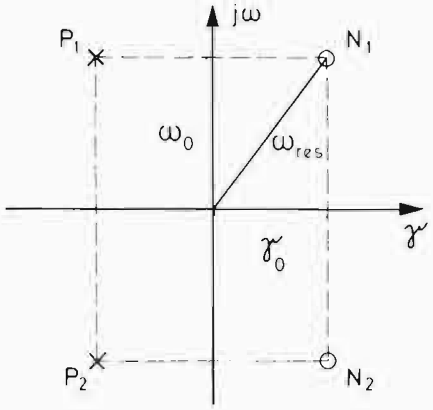


Fig. 7. Fundamental all-pass.



$$\tau(\omega) = 2\gamma_0 \left[\frac{1}{\gamma_0^2 + (\omega - \omega_0)^2} + \frac{1}{\gamma_0^2 + (\omega + \omega_0)^2} \right]$$

$$\omega(\tau_{max}) = \omega_0 \sqrt{2\sqrt{1 + \left(\frac{\gamma_0}{\omega_0}\right)^2} - 1 - \left(\frac{\gamma_0}{\omega_0}\right)^2}$$

$$\tau_{max} = \frac{2}{\gamma_0} \cdot \frac{1}{2\frac{\omega_0}{\gamma_0}\left[\sqrt{1 + \left(\frac{\omega_0}{\gamma_0}\right)^2} - \frac{\omega_0}{\gamma_0}\right]}$$

Fig. 8. Poles and zeros for all-pass network.

ponents or by introducing mutual coupling between coils, simpler all-passes are feasible are shown in Fig. 10.

No method of direct synthesis of component values from an established primary delay behavior is available and present methods of synthesis rely on analogy or cut and try approaches. Lacking a method of direct synthesis, quite elaborate methods of dc analogies have been proposed in the past.[2]

Once a passive all-pass network is adjusted for proper delay and return loss, it can not easily be readjusted without affecting its performance. Therefore, the standard method employing passive all-passes is to attain complex delay curves by suitable combination of many sections.[4] Frequently, amplitude compensation is required also since the finite Q of the all-pass components will not allow its amplitude response to be uniform. Delay compensation by video predistortion can not be provided for the double sideband region in a rigorous way.[3] However, a rigorous solution is possible in the single sideband region of the television signal, in other words, for frequencies from 1.25 MHz up if the transmitter is capable of handling leading overshoots which are placed on the signal through the predistortion process.

Returning to the video portion of the Exciter-Modulator, a second board contains a sync separator. The most significant feature of this circuit is its reliance on a critically damped tuned circuit (C4 and L1) to produce trigger pulses with the rise and fall of all sync signals. The negative trigger only is used to switch the keyed clamp in the linearity corrector and the modulator driver.

The reasons for choosing this circuit are largely twofold: (1) it is very insensitive to hum, noise, and other short-time disturbances riding on the sync signal and (2) its timing function is independent of supply voltages. The output from Z2 are one positive and one negative going pulse which will turn both clamps on during the back porch, see Fig. 11.

Modern power tubes for TV transmitter linear amplifiers require only very moderate amounts of correction for lack of linearity.

The third PC board of the video portion provides for differential gain correction of all causes which are part of the transmitter proper. Five

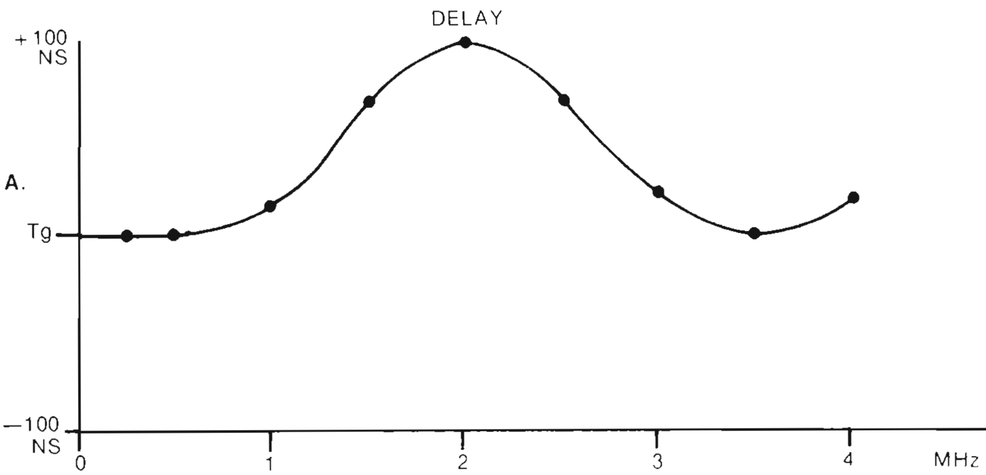
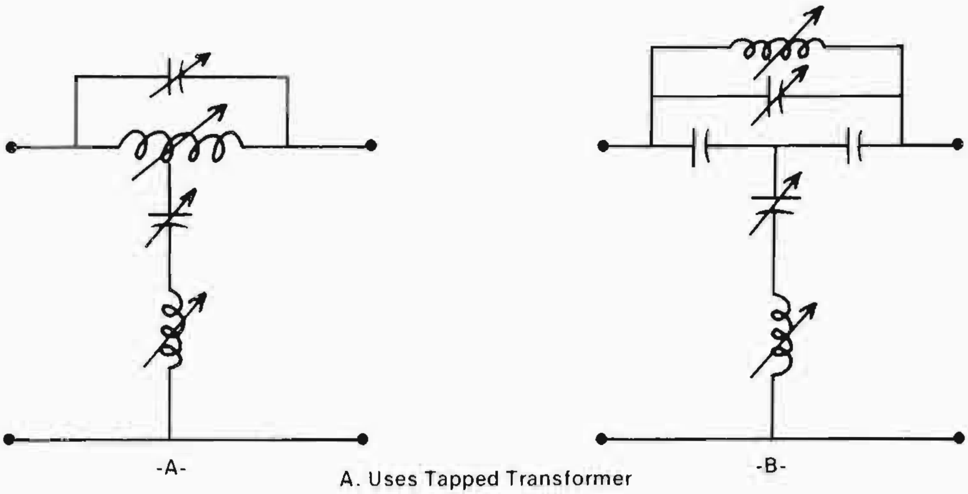


Fig. 9. Typical Delay.



A. Uses Tapped Transformer
B. Uses Split Capacitors

Fig. 10. Simplified all-pass.

steps are provided, including sync stretch. All but the sync stretch controls have potentiometers for adjustment of the cut-in level as well as the rate of correction, which will allow very precise compensation. Correction for differential phase is governed by the degree of correction applied for correcting differential gain.

The clamp circuitry, a necessary part of the linearity corrector, will not affect burst amplitude and will have only a minimal effect on the burst phase. Fig. 12 shows a simplified version of the linearity corrector, including the clamp circuit, to demonstrate the circuit's operation. The action of the linearity corrector can be defeated through a bypass switch.

The last board of the video portion of the Exciter-Modulator constitutes the modulation driver which is also clamped at the backporch level. Also included on this board is a monitor amplifier which offers a sync negative signal which is an exact replica of the signal applied to the modulator. Double sideband modulation is produced through a ring-modulator connected as a current controlled attenuator.

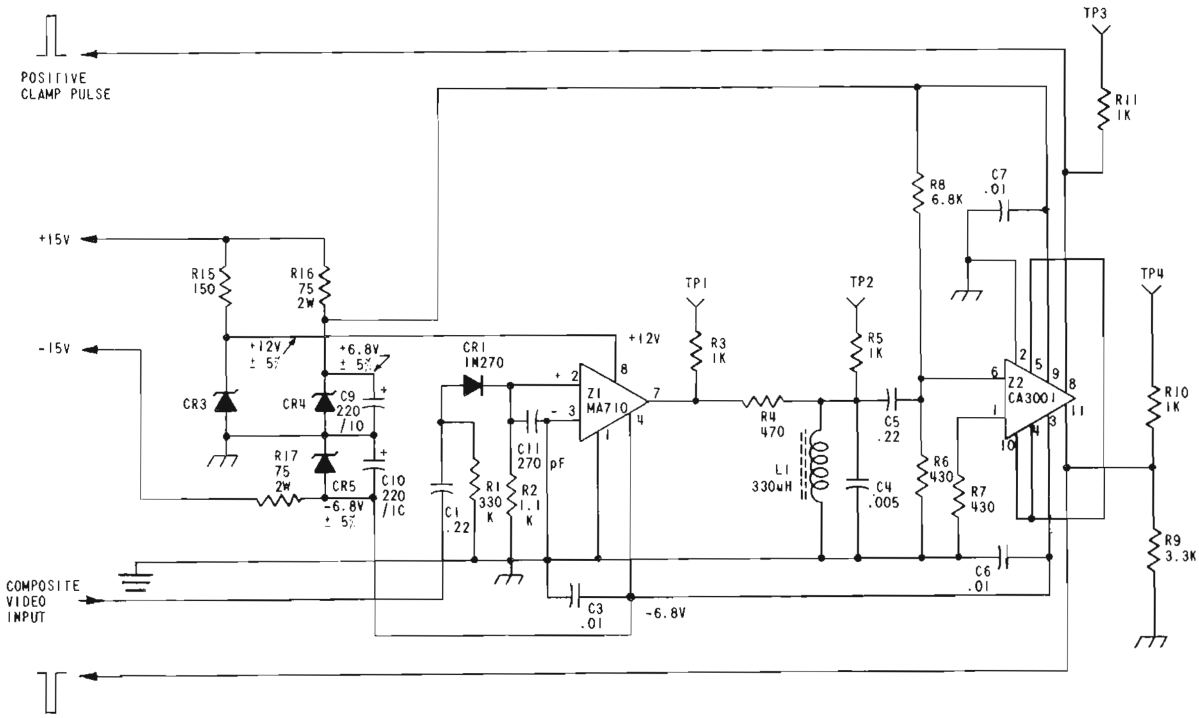


Fig. 11. Sync separator and clamp drive.

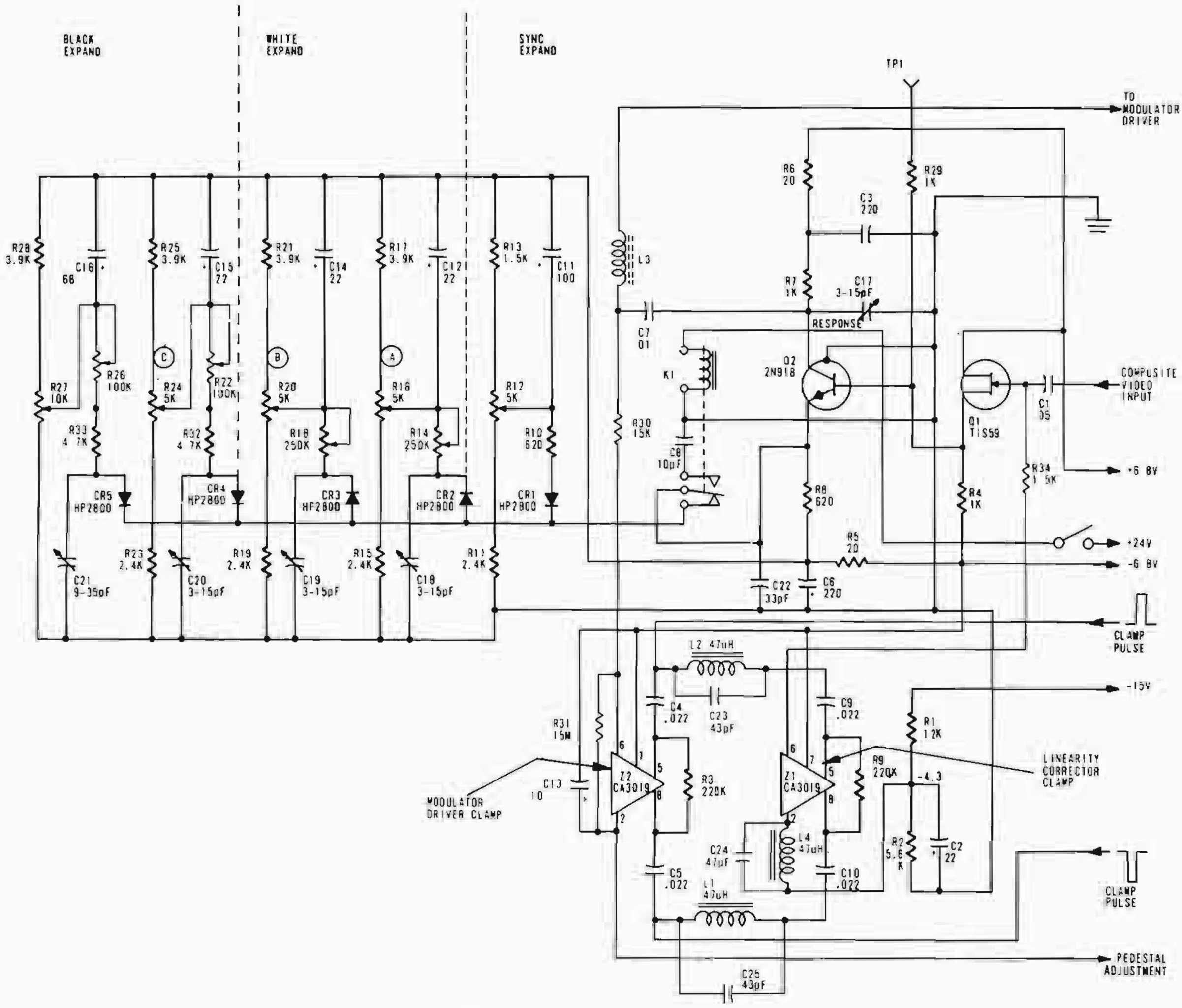


Fig. 12. Linearity corrector.

Modulator

Fig. 13 shows the transfer characteristics of a Hewlett-Packard HP10514 mixer which is typical of any ring-modulator. In the modulation circuit, peak of sync is chosen to be approximately −12 dB attenuation caused by a current of 1 ma. This is the point of maximum carrier. From this point attenuation is available through the reduction of current into the dc coupled port approximately 38 dB below the 1-ma reference, thus, allowing modulation percentages down to 2% if the peak of sync reference is chosen to be 100%. An excursion of current from 1 ma to 10 microamp will accomplish this range of modulation with good linearity.

Care must be taken to remove harmonics of the carrier frequency which consists mainly of the third, with a peak level occurring a maximum carrier approximately 15 dB below the fundamental. Proper harmonic attenuation is achieved with a 3-pole Chebycheff low-pass filter which reduces all harmonics to at least 45 dB below the fundamental, immediately following the modulator. Further harmonic reduction is afforded in all amplifier stages following as well as in the band pass VSB (vestigial sideband) filter.

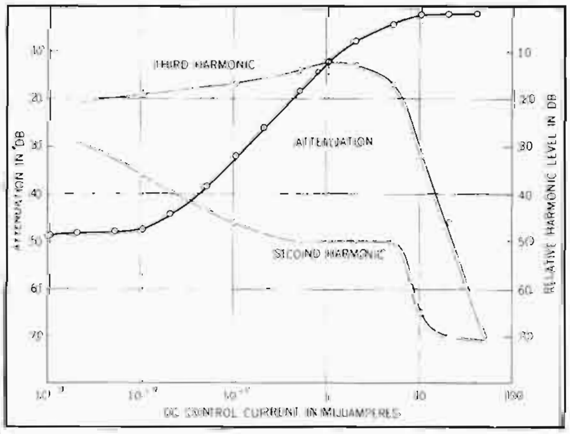


Fig. 13. Ring modulator transfer characteristic.

The modulated double sideband signal with a 37 MHz carrier appears at the R port of the ring modulator and has a level of approximately 1 mw during peak of sync. Typical performance of such a modulator is modulation linearity better than 2% relative to ideal straight line; differential phase less than 1°, and carrier phase shift (100% to 10% modulation) less than 3°.

The modulator has a distinct advantage of requiring no tuning controls. The double sideband modulated signal after passing the low-pass filter and a single stage amplifier is available at 50 ohms and will next be applied to the IF delay compensator.

IF Delay Compensator

In a television system the departure of envelope delay from a uniform curve is caused almost exclusively by the limitation of the bandwidth which is imposed by regulation. The limitation in channel bandwidth was imposed in the interest of spectrum conservation.

In a linear system delay caused anywhere in that system can be compensated anywhere else, e.g., delay in the receiver, can, in fact, be corrected by predistortion in the transmitter.

In a bandpass filter, spectrum components towards the edges of the passband are delayed more than components in the center of the passband. A steeper attenuation at the edges will cause a greater delay. The delay of a typical VSB filter is shown in Fig. 14. If the delay in the center of the passband is arbitrarily established as zero reference, Fig. 14 shows that the relative delay at carrier is approximately 80 ns, the delay at the color subcarrier is approximately 100 ns, and delays at the edges of the passband are slightly in excess of 300 ns.

The delay of a filter is an inherent characteristic (found in all minimum phase-type networks) and is directly related to its amplitude response. The delay cannot be changed without affecting response due to a relationship anchored in fundamental physical laws.

Since no circuits are available that will advance a signal in time, the only way to compen-

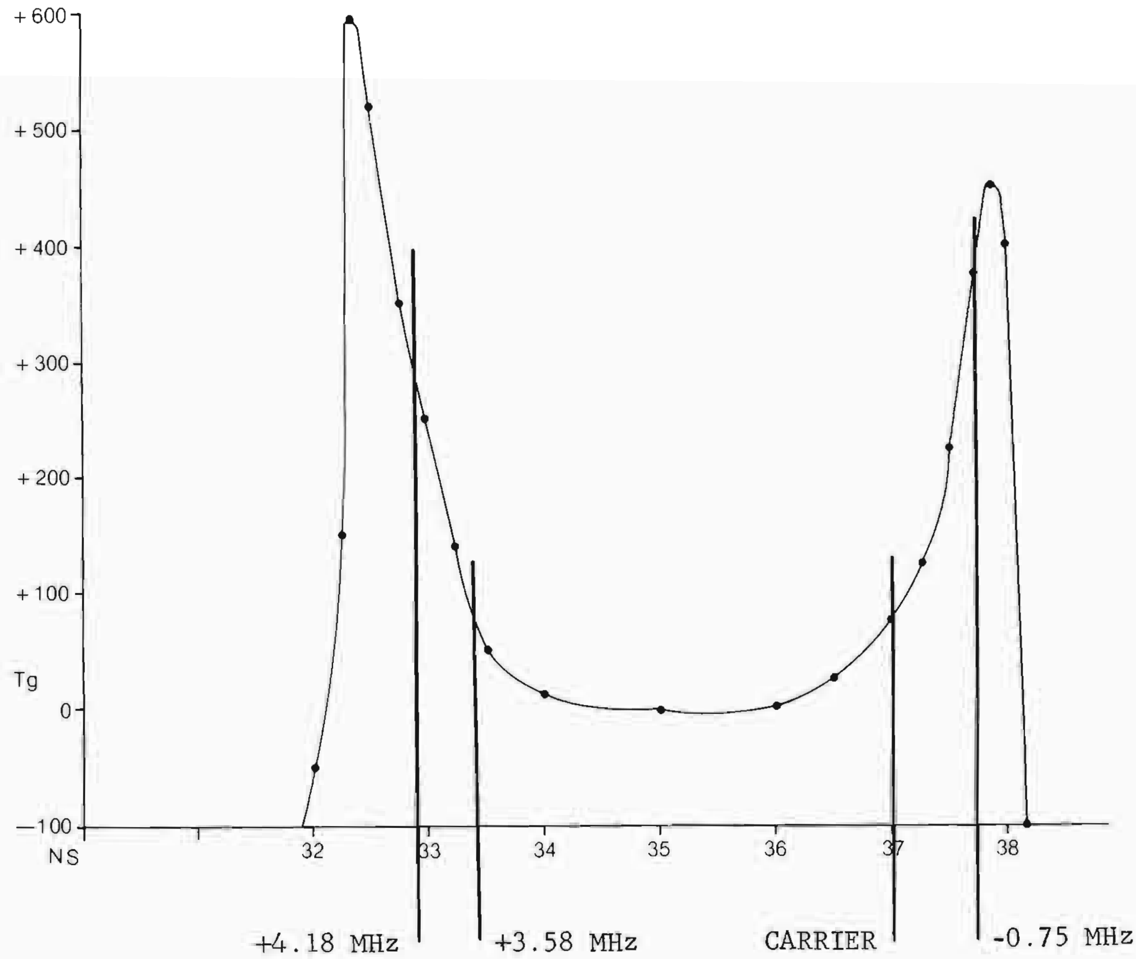


Fig. 14. Delay of typical VSB filter.

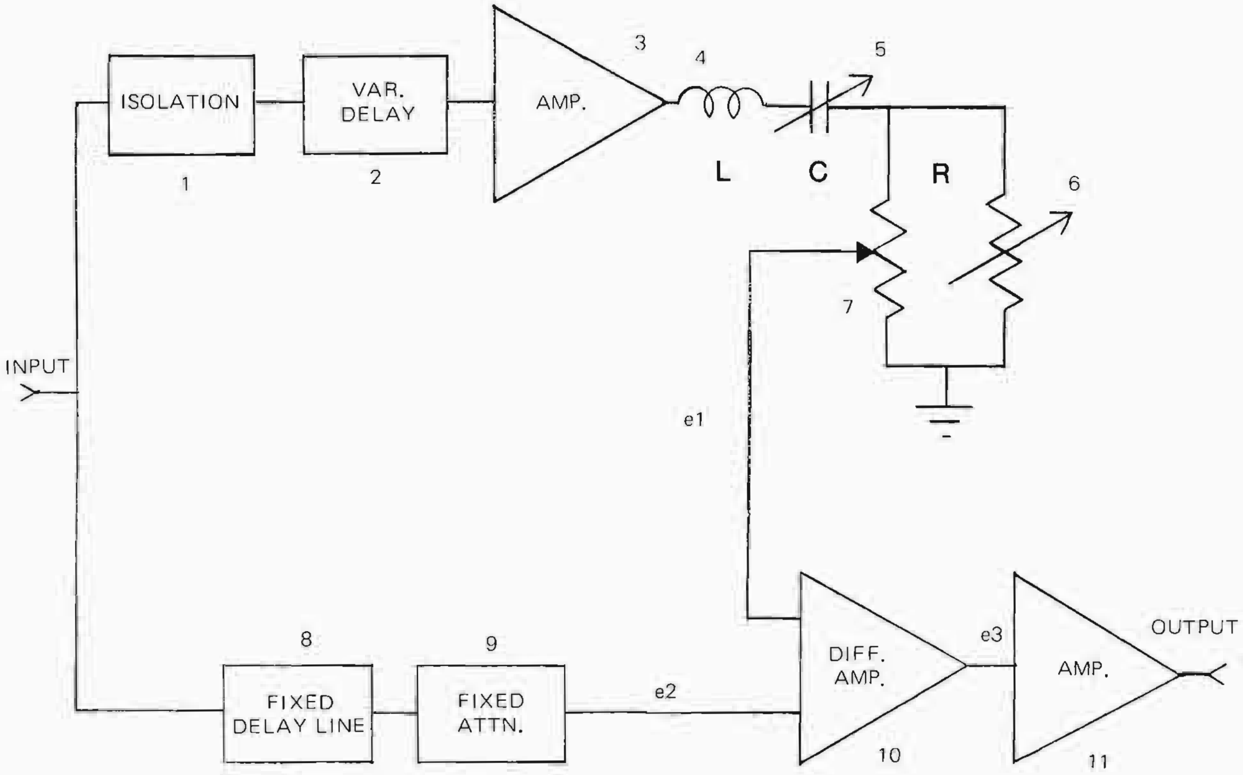


Fig. 15. Block diagram of active delay compensator.

sate nonuniform delay is by artificially delaying elements in the spectrum in order to equalize delay for all components.

This can be accomplished by using all-pass networks (which are not minimum phase type networks). An all-pass has an amplitude response which theoretically is independent of frequency and has a delay maximum at one frequency and zero or finite delay at all other frequencies. Delay compensation at IF frequencies has been accomplished using passive all-pass networks.[6] Compensation using active circuits is now feasible and practical, which provide greater flexibility and accuracy.

The concept of active all-passes is not new,[7,8] but the use at IF frequencies using solid state devices is a recent accomplishment. Fig. 16 depicts voltage relationships of an active all-pass. The phasor relationship through the upper path is shown by phasor e_1 which will vary with frequency and coincide with the real axis at resonance of the LC circuit. Its tip will travel on a complete circle from zero to infinite frequency. Phasor e_2 has a magnitude of one-half of e_1 at resonance and will be subtracted from e_1. It is, therefore, pointing to the left, parallel with the real axis, and projecting from the tip of e_2. The difference phasor of e_2 minus e_1 which is identified as e_3 has a magnitude which is independent of frequency and equal to e_2. Its phase is a function of frequency and will rotate from $-180°$ for zero frequency, to $+180°$ for infinite frequency. Envelope delay is defined as a rate of change of the phase angle of e_3 and this derivative typically is shown in Fig. 17. The point of maximum delay occurs at resonance of LC, with the magnitude controlled by R. If it were physically possible to reduce R to zero, delay would be infinite. If the area under the curve is integrated from zero to infinite frequency, one would find that it is independent of R.

Two quantities characterize the behavior of an active all-pass: ω_0 and γ_0 which are the coordinates of the poles and the zeros in the complex frequency plane as shown in Fig. 8. Envelope delay as a function of frequency is defined as:

$$\tau(\omega) = 2\gamma_0 \left[\frac{1}{\gamma_2^0 + (\omega - \omega_0)^2} + \frac{1}{\gamma_2^0 + (\omega + \omega_0)^2} \right]$$

Circuits intended for IF frequency applications are generally chosen such that γ_0 is considerably smaller than ω_0 which leads, with sufficient accuracy, to the following simplification of the frequency of maximum delay.

$$\omega(\tau_{max}) = \omega_0$$

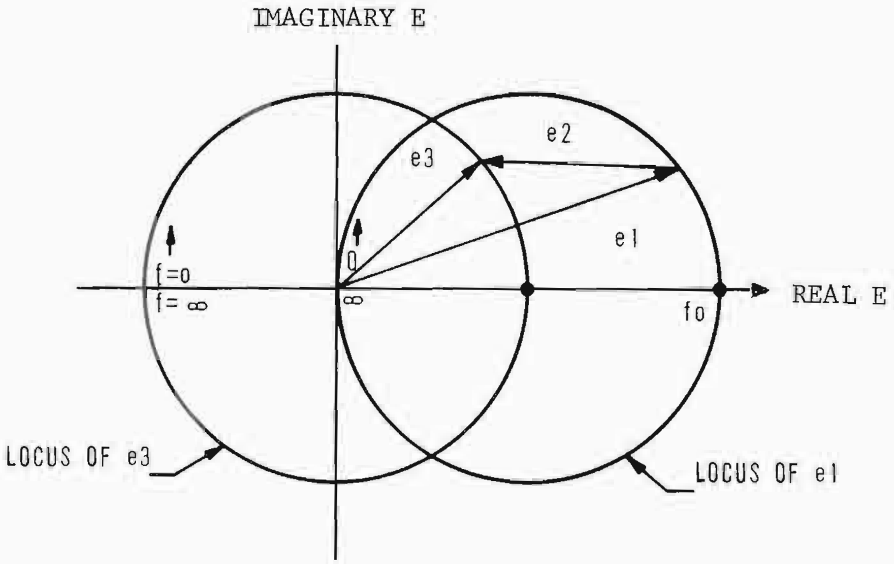


Fig. 16. Phasor relationship of active all-pass.

ω_o is the resonant frequency of LC circuit (components 4 and 5) and the equation simply states that the maximum delay occurs at the resonance frequency. (This is not true when ω_o and γ_o are nearly equal in magnitude which is analogous to the fact the resonant frequency of a tuned circuit with low Q does not coincide with the frequency of maximum impedance or equal magnitude for the reactance of L and C.)

The delay at this frequency is (still with the assumption the γ_o is substantially smaller than ω_o) $\tau_{max} = \frac{2}{\gamma_o}$.

A typical circuit with values of:

$L = 2.2$ uH
$C = 8.4$ pf
R = varies from 20-120 ohms.

Will produce

$$\gamma_o = \frac{R}{2L} = 4.545 \text{ to } 27.27 \times 10^6 \; [s^{-1}]$$

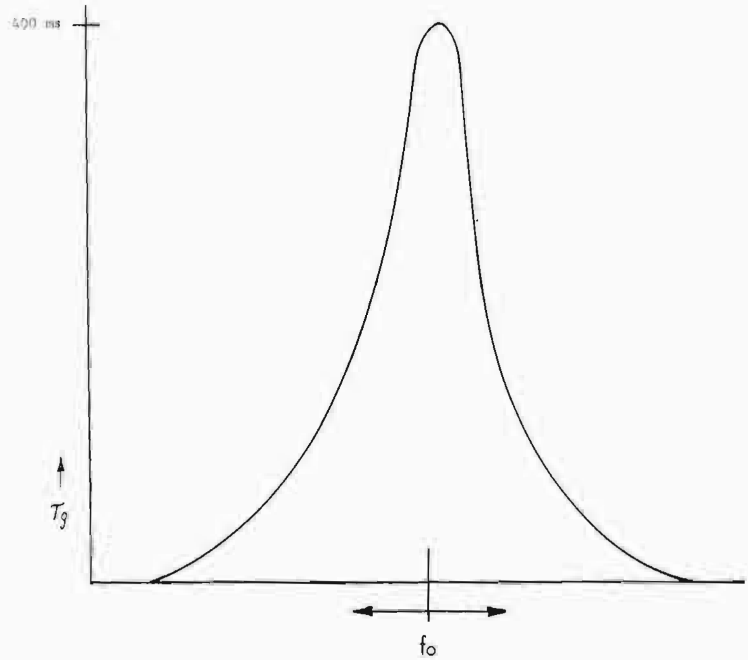


Fig. 17. Time-delay of typical active all-pass.

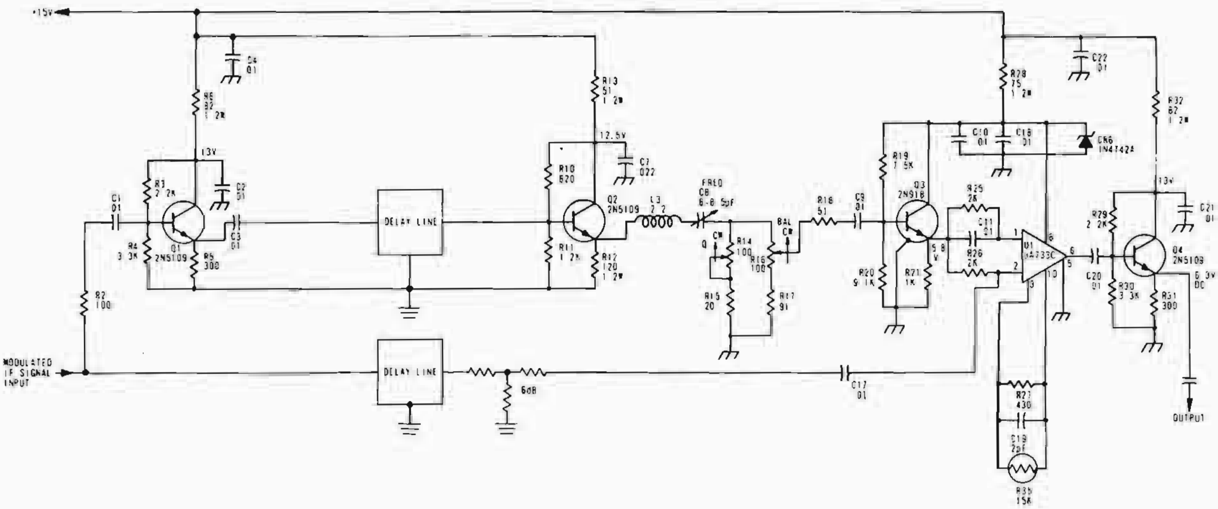


Fig. 18. Simplified schematic of active all-pass.

will have a maximum delay at 37 MHz with delay varying from 440 ns at an *R* of 20 ohms to 73 ns with an *R* of 120 ohms.

The frequency of maximum delay can be chosen anywhere within the passband without affecting the amount of maximum delay by simply tuning C. The amount of delay at any frequency can be varied by changing *R* without changing the frequency of maximum delay.

If γ_0 is chosen to produce approximately 350 ns of maximum delay, it is found that the three sections stagger-tuned over the passband of a television system satisfying the FCC Rules and Regulations will be sufficient to correct the delay of the VSB filter shown in Fig. 14, to ± 30 ns. Adding more sections will improve the correction. The small number of sections needed for this correction is a further advantage of correction of envelope delay at IF frequencies as opposed to correction at video frequencies where more than 10 times the number of all-pass sections are required to attain a comparable degree of correction. Fig. 18 shows a simplified schematic of an active all-pass section. The value of the variable capacitor is chosen to allow the circuit to be tuned from 32 to 40 MHz covering the entire IF bandpass.

Adjustable, frequency-independent delay is provided to assure proper timing of the signals through the frequency-dependent and through the frequency-independent path. Such a delay section is not required for all passes designed for use in video frequency ranges, as the time delay difference through the circuit is negligible at low frequencies.

One should keep in mind when trying to understand the functions of this circuit that the tip of voltage e_2 does not travel with constant speed along the circumference of the circle but moves very slowly at first from zero to say 80% of the resonant frequency, goes rapidly through resonance, and will slow down again continuing to infinite frequency. The speed-up is more pronounced with smaller values of γ_0.

The circuit has four adjustments. Two control only amplitude response and are generally set for a flat response at one time. They are R8 and R16. R8 affects the placement of voltage e_2,

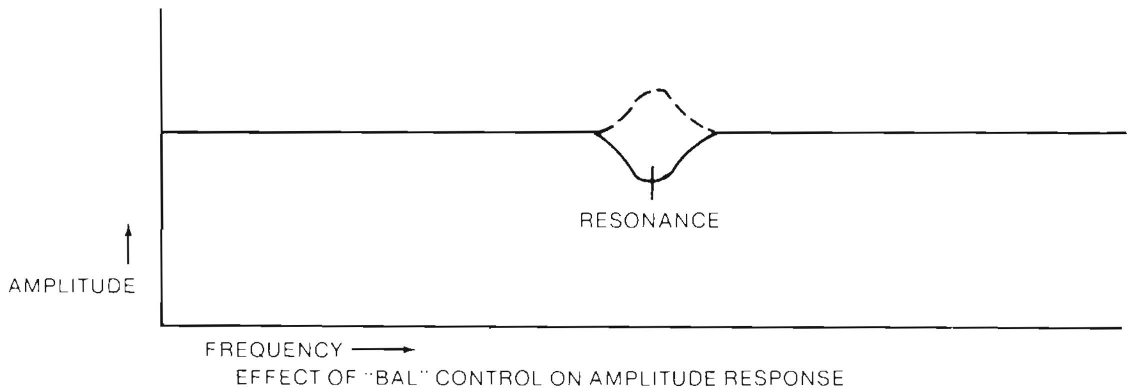


Fig. 19. Active all-pass, effective of mistiming.

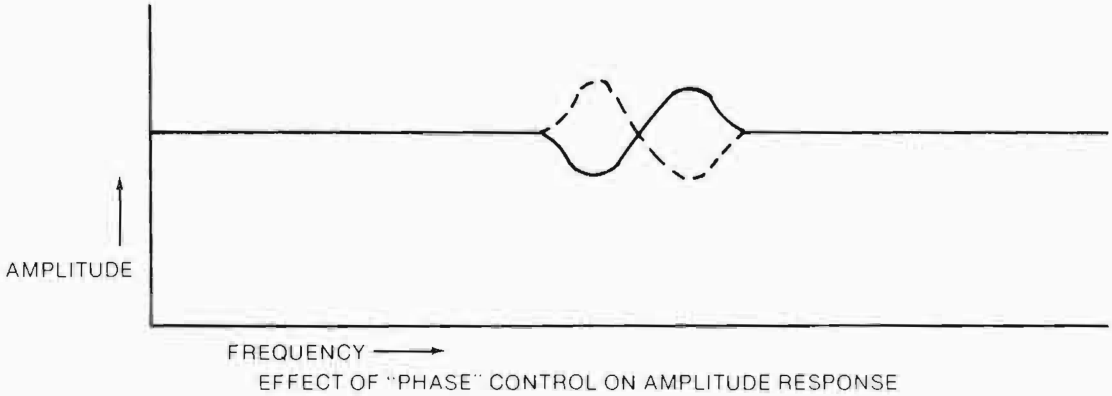


Fig. 20. Active all-pass, effect of variation in ratio of $\frac{E1}{E2}$.

which should be exactly parallel to the real axis. R16 controls the length of e_2 and, therefore, the ratio of e_2 to e_1, which should be exactly 0.5.

Misadjustment of R8 will cause amplitude variation as shown in Fig. 19. The effect of R16 is shown in Fig. 20, both adjustments only affect the amplitude response of the all-pass, they do not affect the delay. Since the disturbances shown in Figs. 19 and 20 occur at the frequency where L and C are resonant, slight adjustment away from balance of R16 serves as a ready means to find the resonant frequency of L and C.

To change the frequency of maximum delay only C needs to be adjusted. The effect of R14 which changes γ_o is shown in Fig. 21. This control is labeled "Q" control as it somewhat resembles the effect of a series resistance on the amplitude response of a resonant circuit.

Initial adjustment begins with setting R8 and R16 for each printed circuit board for a flat response while using the above mentioned R16 to set the resonant frequency correctly for each board. To facilitate each adjustment it is beneficial to bypass all but the printed circuit board which is being adjusted. This will enable the operator to view each board independent of all other boards.

Final adjustment for proper delay is made through small changes in C and R14 while sweeping the unit for envelope delay, or by using proper test signals in the time domain.

Delay adjustment near carrier (±50 kHz) can only be performed with test signals (window) in the time domain since no sweep equipment is available which allows observation sufficiently close to carrier, irrespective of whether amplitude or time delay is swept. The importance of performance within ±50 kHz of carrier is often overlooked.

Fig. 22 shows the schematic of one section of an active all-pass compensated for video fre-

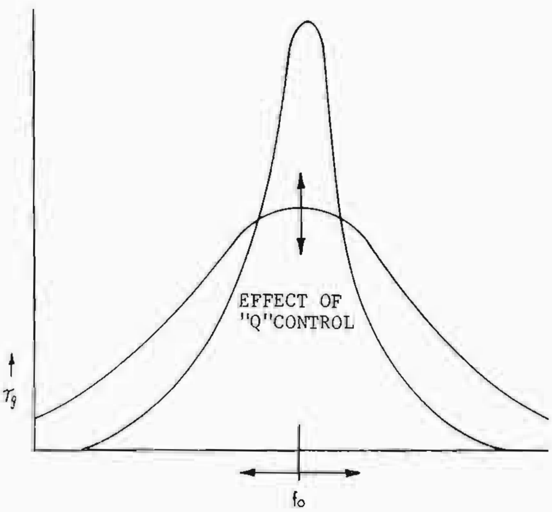


Fig. 21. Active all-pass, effect of R14 and γ_0.

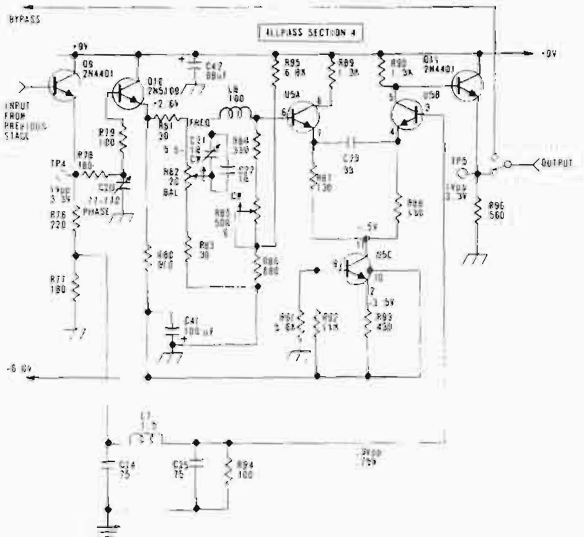


Fig. 22. Schematic of video active all-pass.

quencies. The shorter transmission delay relative to one period of the highest operating frequency of 4.2 MHz simplifies subsequent subtraction of both signals so that only one fixed time delay section is required.

Visual Sideband Filter

To conserve bandwidth, TV signals are generally required to be transmitted with part of the lower sideband removed.

Fig. 1 shows the passband response of typical equipment which will satisfy the requirements of the FCC Rules and Regulations.

If a transmitter exhibits a sufficiently linear characteristic, the shaping of the passband may be performed at the transmitter output at an intermediate stage or, in part, at video frequencies. Operating systems are in use employing all of these methods.

While there is no difference in principle between bandshaping at high or low powers, there are practical differences affecting the degree of presence of spurious frequencies, the ease of tuning, the ease of repair of failures, size, cost, and complexity of the equipment.

Passband shaping can be performed at intermediate frequencies and at low power. The design must assure that the process of frequency translation and the process of power amplification is linear in amplitude and phase over the operating range and frequency in question. A bandpass to meet the requirements of Fig. 1 can, without great difficulty, be realized at IF frequencies in the 30 to 40 MHz range.[11,12]

Such a filter is shown in Fig. 4 with a picture of its amplitude response shown in Fig. 23. The envelope delay of such a filter can be found in Fig. 14 in the previous paragraph. The designation of this type of filter is CO53357 as outlined in Reference 12.

After determination of the required number of poles from Fig. 1 in Reference 12 (five poles in this case), the design of a prototype low-pass filter can be formulated as shown in Fig. 24.

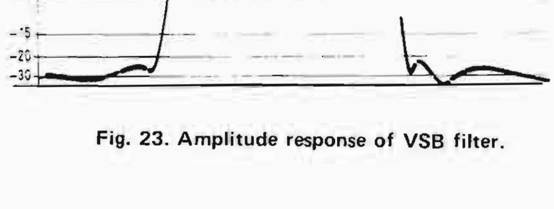


Fig. 23. Amplitude response of VSB filter.

This low-pass prototype is then transformed into its bandpass equivalent as determined by the required mean frequency. In doing this, some unrealizable circuit elements are found in the series portion of the filter which require transformation from the original configuration of the series elements as shown in Fig. 25A into a more suitable set of equivalents as shown in Fig. 25B.

The actual physical form of this filter must be chosen carefully to minimize stray capacitance and stray coupling and to assure that each reactance performs only the function for which it is provided. The success of the designer in constructing a useful filter in this particular application is not determined by his skill within the mathematical portion, but by his skill in the practical circuit realization.

This type of filter is known either as Cauer or elliptical and is generally chosen when a rapid transition from passband to stopband with a minimum number of circuit elements is desired, and a certain amount of ripple in the pass- as well as in the stopband can be tolerated.

The frequency of operation makes it possible to use air coils, air capacitors, and porcelain capacitors which produce a design of unusual stability, which is practically unaffected by changes of ambient temperature and the course of time.

The tuning range of the capacitors is chosen to allow tuning of this VSB filter over a frequency range sufficiently large to meet all the different standards of television systems around the world (with the exception of SYSTEM L, CCIR).

There is no question that any manufacturing economy achieved by using fewer components, easier assembly methods, simpler tests, larger number of similar components, or more universally useful circuits will quickly benefit the final user in lower capital costs, lower operating expenses, higher reliability, and better performance.

The output of this VSB filter is applied to an amplifier which will overcome the passband attenuation of approximately 8 dB.

The filter can be bypassed through a front panel switch without causing any change in gain at the carrier frequency and will then extend the passband of the Exciter-Modulator to ±10 MHz symmetrical to the carrier frequency.

The last stage of the IF assembly is a PIN diode attenuator with a range of approximately 13 dB. It controls total system gain and can change, therefore, the output power of the transmitter from 110% of rated power to 5% of rated power. The controlling quantity is a dc current through the PIN diode. This current is adjustable by a front panel control which is

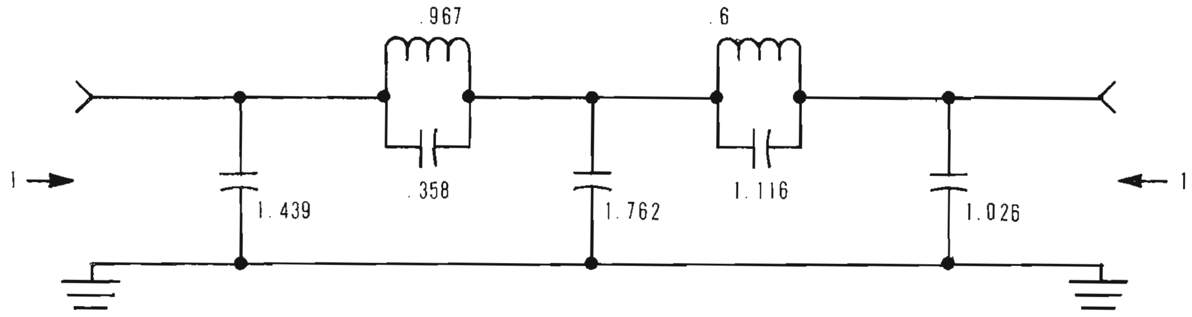


Fig. 24. Low pass prototype of VSB filter.

also motor driven for remote access. The PIN diode looks into the base of an emitter follower whose output will be connected through a suitable attenuator to the up-converter. The change in performance of the processed IF signal of the above stated range from 5 to 100% is almost unrecognizable.

The level of the IF voltage at the peak of sync is available at the front panel multimeter. This signal is applied to the up-converter to be translated to the final frequency of the station.

Up-Converter

A ring modulator, also known in this application as a double-balanced mixer performs the translation to the final operating frequency. This type of conversion has a very desirable output spectrum with both input signals as well as all even order mixing products suppressed.

The mixing products, other than the wanted frequency requiring the greatest attention are the sum frequency and the pump frequency.

The difference frequency between the pump and the IF is the wanted signal and should be available from the up-converter without any deterioration. The sum and pump frequencies are rejected by a bandpass filter. To ascertain preservation of the wanted signal a wide bandwidth of four times the actual passband is employed. A commercially available bandpass suits this application and provides 30 dB attenuation for the pump signal and 45 dB for the sum frequency. Coupled with the conversion loss and the transfer data of the mixer plus the frequency rolloff of the following amplifier and the Exciter proper, it will reduce the level of the sum frequency to a neglible value (in excess of 90 dB) and reduce the pump frequency from 30 to 35 dB below the wanted signal.

The selectivity of the following transmitter stages will finally reduce the pump frequency to below 90 dB. The characteristic of a double balanced mixer and its isolation between ports is shown in Fig. 26. Because of the inherent design, these values can be maintained with utmost certainty, and they are not influenced by time or temperature.

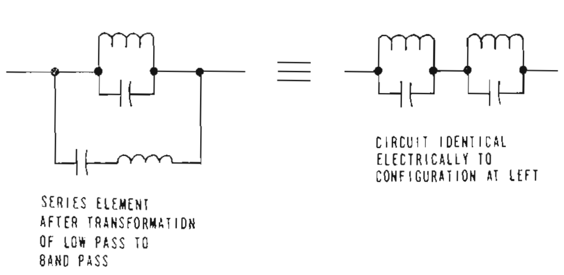


Fig. 25. Series element transformation a, b.

Power Amplifier (Exciter)

In specifying an amplifier for a television visual application the following parameters are of importance:

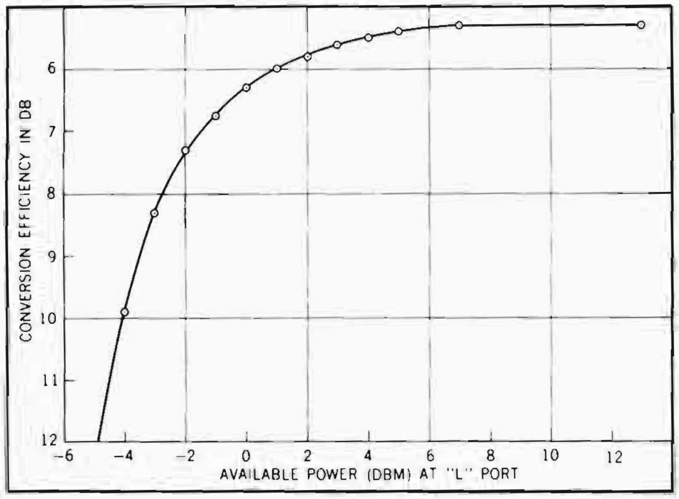
1. Linear frequency response at any power level over the frequency range in question,
2. Linear transfer characteristics, input versus output.
3. Linear phase versus frequency characteristics independent of power level.

The order stated is representative of the degree of difficulty. Item 3 is especially difficult in higher power solid state amplifiers, i.e., where more than 10 watts peak of sync power is demanded of a single device.

Fig. 27 shows the schematic of an amplifier meeting the above stringent requirements at an overall gain of 40 dB with a maximum peak of sync output of 1 watt.

Fig. 28 shows typically the frequency response of this amplifier.

The departure of the input to output phase at varying power levels can easily be measured with a Vector voltmeter and will reveal variation



Conversion Efficiency

Product	Level†	Product	Level†
$2f_L - f_R$	30 dB	$2f_R - f_L$	65 dB
$3f_L - 2f_R$	70 dB	$3f_R - 2f_L$	65 dB
$4f_L - 3f_R$	70 dB	$4f_R - 3f_L$	85 dB
$5f_L - 4f_R$	90 dB	$5f_R - 4f_L$	90 dB
$6f_L - 5f_R$	95 dB	$6f_R - 5f_L$	100 dB
$7f_L - 6f_R$	100 dB	$7f_R - 6f_L$	100 dB

† Referred to f_X level.

10514A

MIXER CONVERSION LOSS (Single Sideband):

Frequency Range		Conversion Loss (dB max.)
f_L and f_R (MHz)	f_X (MHz)	
0.5 - 50	dc - 50	7
0.2 - 500	dc - 500	9

NOISE PERFORMANCE:

Frequency Range		Noise Figure (dB max.)
f_L and f_R (MHz)	f_X (MHz)	
0.5 - 60	0.05 - 60	6.5
60 - 500	0.05 - 500	9

Less than 100 nV per root cycle max. at output for f_X at 10 Hz.

MIXER BALANCE:

Mixer Balance for	In Frequency Ranges (MHz)		Referenced to
	f_L, f_R: 0.5 - 50; f_X: dc - 50	f_L, f_R: 0.2 - 500; f_X: dc - 500	
f_L at R	40 dB	30 dB	f_L
f_L at X	40 dB	20 dB	f_L
f_R at L	45 dB	30 dB	f_R
f_R at X	25 dB	15 dB	f_R
f_X at L	35 dB	15 dB	f_X
f_X at R	25 dB	15 dB	f_X

Fig. 26. Double balanced mixer characteristic.

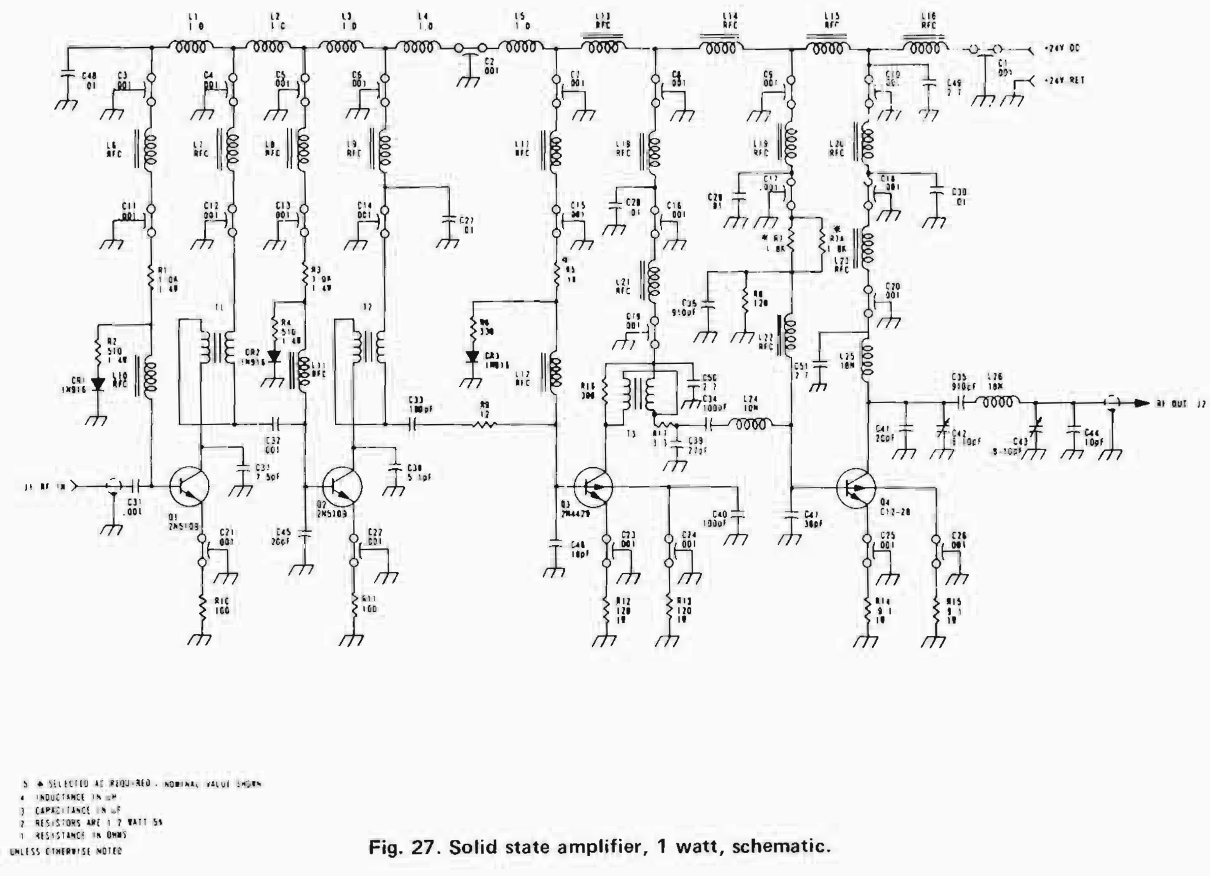


Fig. 27. Solid state amplifier, 1 watt, schematic.

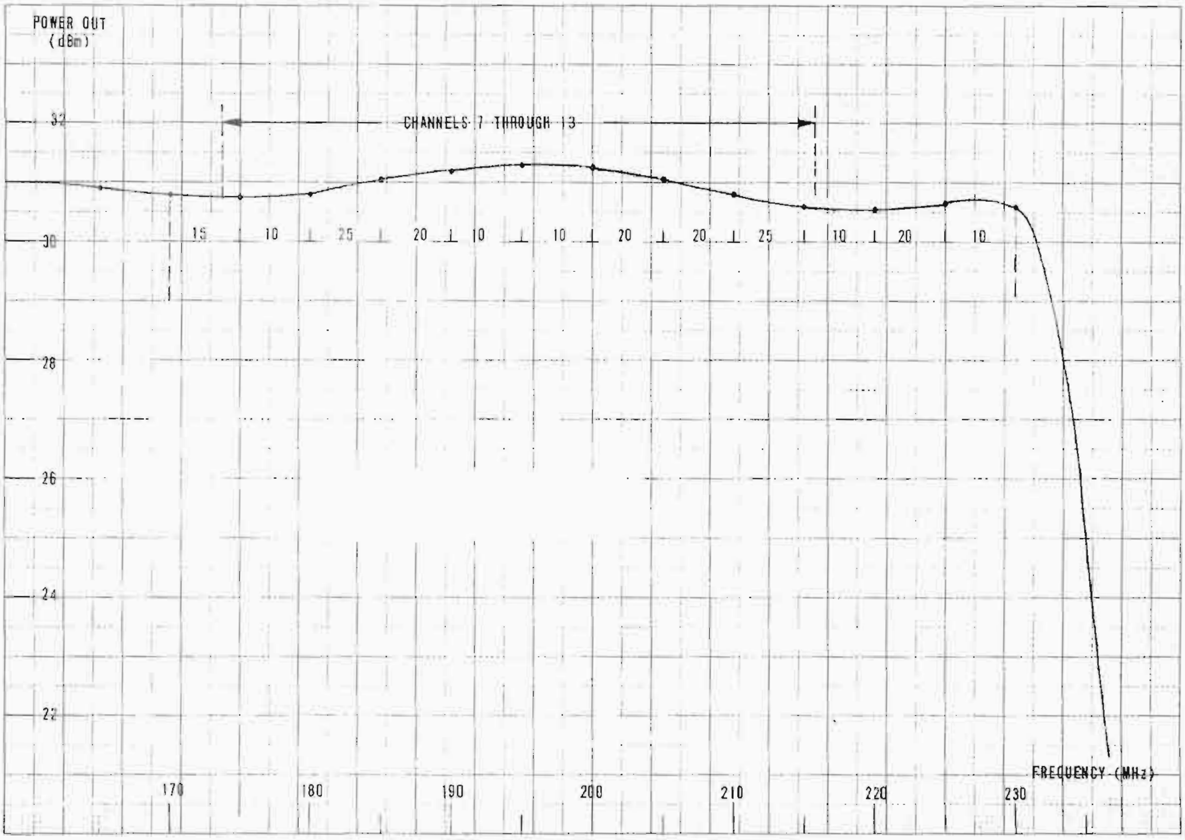


Fig. 28. Solid state amplifier, 1 watt, frequency response.

of less than 1° from 1 mw to 1-watt output power. Design of this amplifier is so arranged that it can safely be operated into any termination, including open or short circuits. Following this amplifier a directional coupler will allow sensing of the peak of sync power level leaving the Exciter as well as the power level of reflected power. Both are indicated on the front panel meter.

The circuitry described so far has produced a television visual signal at a power level of approximately 1 watt peak-of-sync, which carries the original video information with a very high degree of fidelity, proper modulation percentages, proper passband, and the proper envelope delay. This signal can, if desired, be transmitted. To obtain powers in excess of 1 watt all that is required now is amplification.

Intermediate Amplifier

There is little difference in principle between tube type amplifiers for 100 watts output or 10 kw output. The different approaches taken, however, are dictated by the physical size of components and devices and how their size relates to the frequency or wavelengths employed and, especially in higher power devices, by the need to remove large amounts of unwanted heat generated in the device.

Design of a linear amplifier using thermionic tubes for television service progresses as follows:

1. Determination of the highest load impedance which will provide the required bandwidth.
2. Establish maximum power available under conditions of 1 and determine stage gain.
3. Check that no ratings of the device are exceeded and establish safety factors.

Relative to Condition 1, Reference 13 provides a ready made design approach which comes close enough to theoretical limits to be of great practical help. See Fig. 29. C1 in Fig. 29 is the output capacity of the tube including all stray capacity. It will become quickly clear that the higher the tube's output capacity, the lower the load impedance will be for a given bandwidth. Therefore, smaller tubes will allow higher load impedances or greater bandwidth. Typical output capacities for tubes in the 10 to 100 watts power range are 1 to 10 pF.

A typical small tube, e.g., 8122, will show the following characteristics:

$$f_o = 54 \text{ MHz}$$
$$2f_1 = 6 \text{ MHz}$$

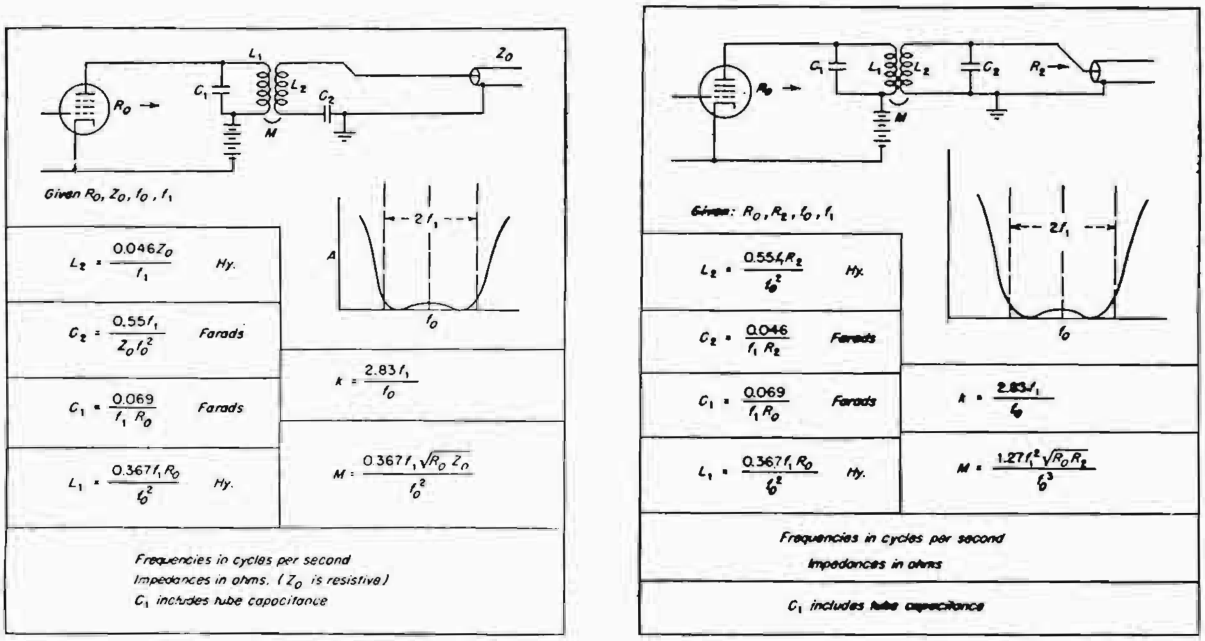


Fig. 29. Double tuned circuit.

$$
\begin{aligned}
C_1 &= 15\ \text{pF}\\
L_1 &= 578\ \text{nH (for R2 = 50 ohm)}\\
C_2 &= 11.3\ \text{pF}\\
L_2 &= 766\ \text{nH}\\
R_0 &= 1533\ \text{ohm.}
\end{aligned}
$$

For R_0 a load line can be constructed (assuming for the moment that R_0 is resistive over the entire bandwidth) to determine output power capability of the tube.

A graphical method established by Reference 14 is most convenient. Plastic overlays for this analysis are available from Eimac. To utilize Chaffee's method, constant current curves must be available which for most power tubes are part of the data sheet supplied by the tube manufacturer.

Fig. 30 shows a sample calculation using a 8122 tetrode which indicates power output, load impedance, and gain to be expected under the conditions as stated. It should be noted that the tube is driven only lightly into grid conduction.

The methods cited so far are intended to demonstrate to a television transmitter user the thinking on which the designer bases his decisions. The methods are chosen mainly because of their simplicity rather than accuracy, but it can be expected that the results obtained will be within 20% of actual performance. For example, the calculated RF anode current does not include the "negative" swing. However, if closer accuracy is desired, more elaborate versions of the tube calculator are available. Finally, checks must be made to ascertain that all circuit elements for power levels and voltages in question operate safely within their maximum ratings established by individual manufacturers, plus any safety factor the equipment designer wishes to apply to gain a greater degree of reliability. From the example chosen above, it is evident that one is in a transitional area where lumped circuit components become increasingly ill-defined and impractical, but distributed elements are still often unwieldy. The designer's choice rests largely on his personal judgment and his mechanical skills of circuit implementation. Often a softened approach is taken by only partially employing distributed devices and also, through the use of "Striplines" whose physical embodiment do not create pure distributed elements. Yet, they are extremely useful in this transitional area.

Fig. 31 is a photograph showing a typical approach in this area. The foregoing comments apply largely to VHF low band equipment. For use on high-band channels from 174 MHz up, lumped elements are practical only to levels of 10 to perhaps 100 watts, but from thereon coaxial construction is generally required. Table 1 lists a series of tubes manufactured by several companies which find application in television service. These are arranged in ascending power dissipation capability.

Power Amplifiers

Most TV transmitters over 5 kw require RF linear amplification to reach power levels up to 25 to 30 kw per tube.

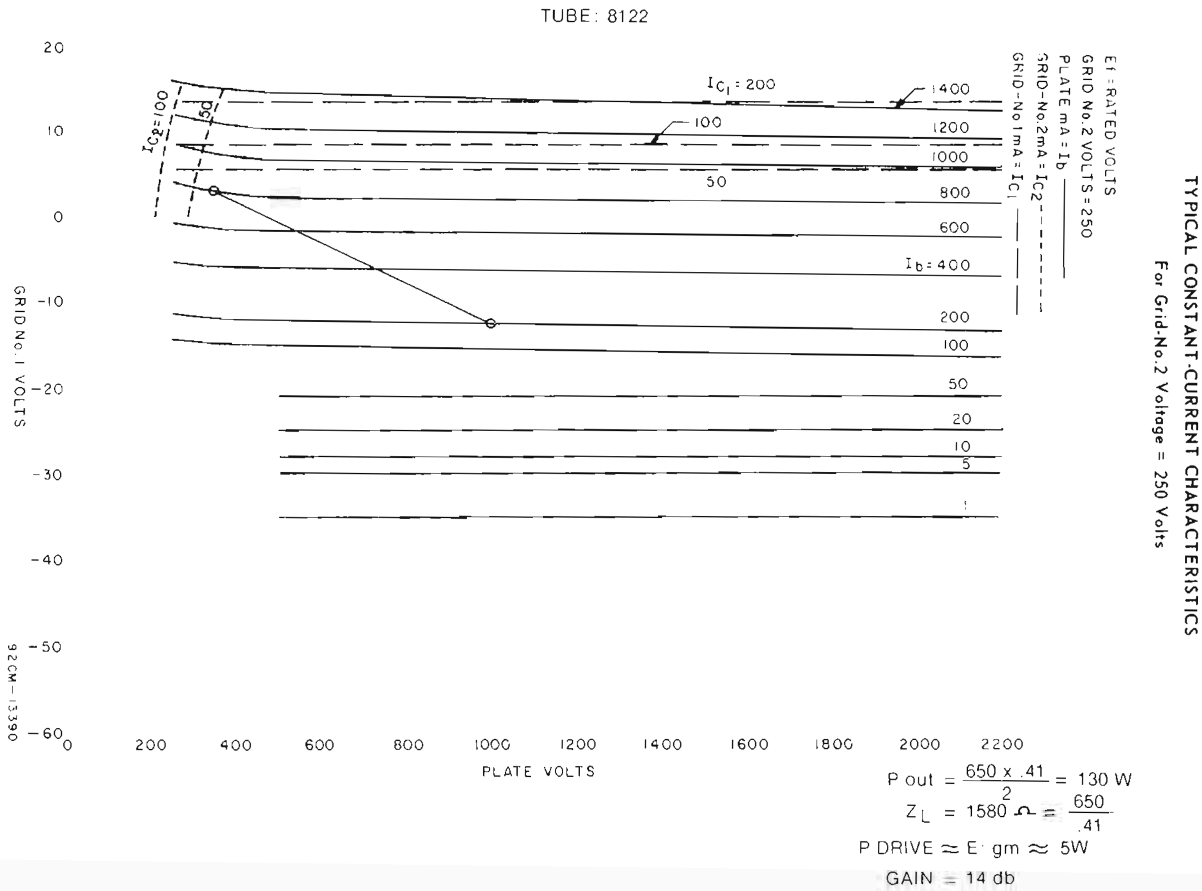


Fig. 30. Power output of 8122 tube.

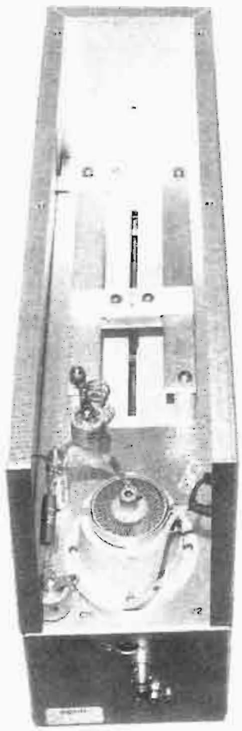


Fig. 31. Photograph of IPA anode tuning circuit.

Table 1 shows the number of different tubes available for this type of service. At first the list appears to provide a fairly wide choice of tube types. But additional requirements, e.g., cost, cooling, cathode construction, size, etc., often narrow the choice, sometimes to such a degree, that compromises are required to find even one tube type for a particular use.

The principal approach to design a high power stage is not different from the outline given in the previous section. It becomes obvious, however, that physical size works against the designer in a multifaceted way. While pseudo-distributed circuits, like striplines (in a sufficiently small enclosure) are still feasible for Low Band VHF transmitters, coaxial cavities are the only useful and practical approach for Band service. Operation at UHF frequencies using gridded tubes will not be considered here even though interesting gridded tube designs do exist. From principles mentioned in the previous Section, it can be stated qualitatively that if a tube, or other device, has zero output capacity an amplifier with infinite bandwidth could be

TABLE 1

Power Tubes for TV Linear Amplifier Service (Maximum Ratings)

Number	Manufacturer	Type	Frequency (in MHz)	Power[a] (in watts)	Output capacity (in pF)	Gain[b] (in dB)	Anode voltage (in volts)	Cathode current (in mA)	Anode dissipation (in watts)	Remarks
YD1300	AMPEREX	Triode	1,000	35[c]	3.5	20	2,000	200	300	
YDi302	AMPEREX	Triode	1,000	55[c]	3.5	19	2,000	250	325	
8814	AMPEREX	Tetrode	260	1,550	9	14	4,000	1,000	1,500	
YD1334	AMPEREX	Triode	1,000	110	8	14	3,500	550	1,800	
YD1336	AMPEREX	Triode	1,000	220	8	15	3,500	550	1,800	
YD1335	AMPEREX	Triode	1,000	550	8	14	3,800	700	1,900	
YL1540	AMPEREX	Tetrode	260	1,150	8	20[d]	4,000	1,500	2,000	
8812	AMPEREX	Tetrode	250	8,600	16.4	14	6,500	2,250	6,000	
8813	AMPEREX	Tetrode	250	18,400	18	14	9,000	3,500	12,000	
8915	AMPEREX	Tetrode	260	27,500	23	14	9,000	7,000	18,000	
8873	EIMAC	Focused triode	500		6		2,200	350	200	Conduc. cooled
4CX250B	EIMAC	Tetrode	500		4.5		2,000	250	250	
8875	EIMAC	Focused triode	500		6		2,200	250	300	Low vel. air cool
8874	EIMAC	Focused triode	500		6		2,200	350	400	
8877	EIMAC	Focused triode	250		10		4,000	1	1,500	
8938	EIMAC	Focused triode	500		13		4,000	1	1,500	
4CX1500B	EIMAC	Tetrode	110/225[e]		12		3,000	900	1,500	
3CX3000A7	EIMAC	Triode	110		24		5,000	2.5	3,000	
4CX5000A	EIMAC	Tetrode	30/225[e]		14.5		7,500	4	5,000	
3CX10000A7		Triode	160	15,000	36		8,000	5	10,000	
4CX15000A	EIMAC	Tetrode	110/225[e]	16,500	25.5	10	6,500	5	15,000	
3CX20000A7	EIMAC	Triode	110/216[e]	25,000	36	13	8,000	6	20,000	
CX20000V7	EIMAC	Focused triode	250	25,000	20	13	10,000	5	20,000	
8122	RCA	Tetrode	500	130	7	14	3,000	300	400	
8791V1	RCA	Tetrode	400	500	5	13	3,000	750	1,000	
8792V1	RCA	Tetrode	400	1,350	16	12	3,500	1.25	1,500	
4680	RCA	Tetrode	900	1,500	16	11	5,000	2	1,500	
5762	RCA	Triode	220	6,350	18.5	8	4,500	2	4,000	
8890	RCA	Tetrode	400	5,000	12	16	8,000	4	5,000	
4682	RCA	Tetrode	400	7,500	10	16	13,000	4	8,000	
8501	RCA	Tetrode	900	5,500	13	9	7,000	4	10,000	
6166A	RCA	Tetrode	220	14,000	24	9	7,500	4	12,000	
8806	RCA	Tetrode	400	10,000	13	15	8,000	4	12,500	
8807	RCA	Tetrode	400	17,600	18.5	14	9,000	6	15,000	
4681	RCA	Tetrode	400	17,600	18.5	14	7,500	6	15,000	
4683	RCA	Tetrode	400	17,600	18.5	16	9,000	6	15,000	
8891	RCA	Tetrode	400	20,000	17	14	9,000	6	17,500	
8916	RCA	Tetrode	400	27,500	17	14	13,000	6	22,000	
6448	RCA	Tetrode	890	15,000	30	14	7,000	7	26,000	
6806	RCA	Tetrode	890	28,000	30	15	9,000	8.25	35,000	

[a]Peak of Sync Power in TV linear service, 6 MHz minimum bandwidth at -1 dB points.
[b]Gain in TV linear service as grounded grid amplifiers.
[c]Sync power output in TV translator service, combined sound and vision.
[d]Grid-driven amplifier.
[e]Max. frequency for TV visual service.

built using an ideal coil and resistor as loads. If, however, distributed reactive elements, even ideal ones, are used, this is no longer true, even if the characteristic impedance of the distributed line goes to infinity. With finite impedance, one has to contend with a bandwidth shrinkage factor which has its cause in the stored energy of the distributed line.

When the cavity impedance reaches a value of about one-half the reactance of the anode capacitive reactance of the tube, the bandwidth is cut in half. A typical 15-kw tube may have the total output capacity of 30 pF, giving a reactance of 20 ohms.

A line impedance of at least 20 to 30 ohms is desirable and a typical cavity design can be found in Reference 15. Fig. 32 from Reference 15 illustrates the evolutionary steps in the realization of a cavity design, starting with a lumped equivalent through a distributed model to the

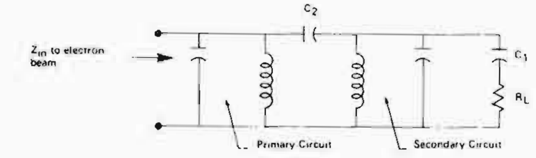


Fig. 32a. Cavity, double-tuned circuit model.

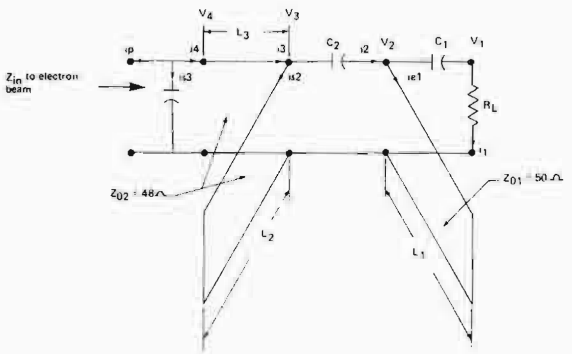
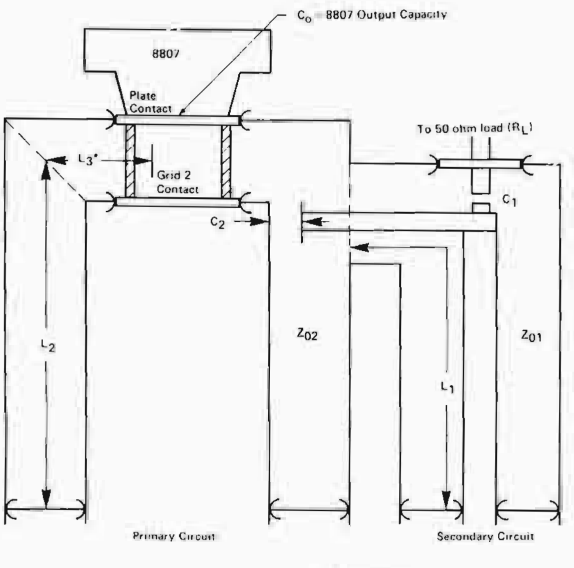


Fig. 32b. Cavity, double-tuned transmission line model.



* L_3 includes transmission line sections within tube structure

Fig. 32c. Cavity, double-tuned including secondary circuit.

mechanical model. Fig. 33 depicts the final manufactured device.

The use of double-tuned circuits will increase the usable bandwidth by a factor of 1.6 over that of a single tuned circuit. The resistive component of the load presented by the above circuit is shown in Fig. 34.

Another important criteria in linear amplifier design is its transfer characteristic. This will determine gain and distortion. In a SSB, or vestigial sideband system (VSB), nonlinear distortion will not only falsify the information

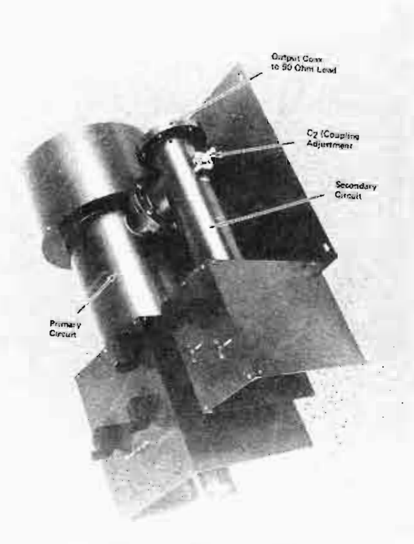


Fig. 33. 18 kw cavity for 8807 tube.

being carried, but also recreate the unwanted sideband or a portion of it.

TV systems worldwide need not meet very stringent requirements of unwanted sideband suppression as opposed to SSB or ISB (independent sideband systems). Government regulations of TV systems universally require 20 dB attenuation of the suppressed portion of the lower sideband. The remaining vestige of the lower sideband extends to 0.75 MHz below carrier. The 20 dB attenuation must be maintained from 1.25 MHz below carrier. The lower side-

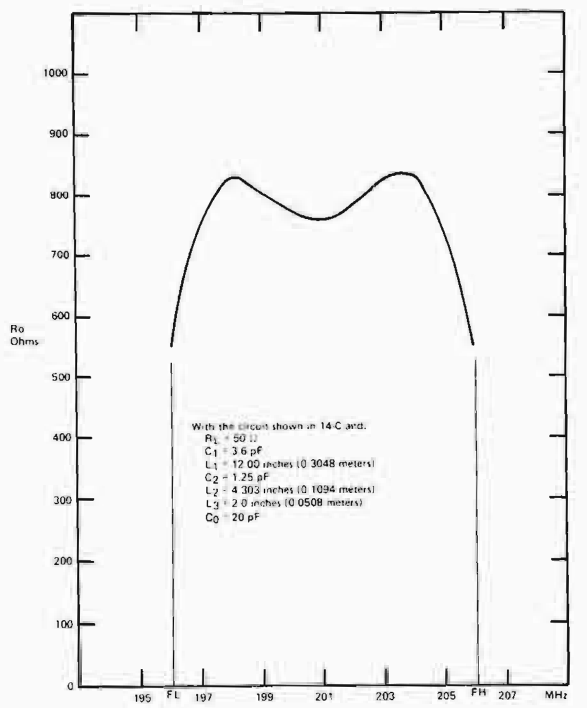


Fig. 34. Load impedance of double-tuned circuit.

band of the color subcarrier (−3.58 MHz) must be attenuated to −42 dB. Reference 15 provides much insight into the mechanism responsible for nonperfect transfer characteristics. Fig. 35 gives an indication of intermodulation performance of a typical tube.

Linearity can be measured using two-tone signals, especially in narrow band systems. Two RF carriers of equal magnitude, 1 kHz apart, are used to drive the amplifier. Third and fifth order products are measured and related to the magnitude of one of the two RF carriers as a means of linearity. If f_1 and f_2 are 1000 and 1001 kHz, the first third order product falls at 1002 kHz ($2f_1 - f_2$) and 999 kHz ($2f_2 - f_1$) while the fifth order product falls on 1003 kHz ($3f_2 - 2f_1$) and 998 kHz ($3f_1 - 2f_2$).

Tubes intended for TV translator service are rated employing a three-tone test which more closely duplicates actual operating condition in TV service. The three tones represent: the visual carrier, the aural carrier, and the upper color subcarrier sideband, with amplitude relationships of −8, −7, and −16 dB. The sum of the three signals is 0 dB and the reference normally chosen is the peak-of-sync operating power. Under the above conditions, the most visible interference is the difference frequency between the aural carrier and the color subcarrier sideband, a beat of approximately 920 kHz. Widely applied standards require this mixing product to be −52 dB or less, relative to the 0-dB reference signal.

There are a number of factors which will influence linearity of performance of an amplifier. Among those are the idling current and screen voltage (assuming the anode voltage is fixed). Fig. 35 shows that lower distortion results when higher idling currents are used. Decreased amplifier efficiency (higher input power) is the price one must pay for this improvement.

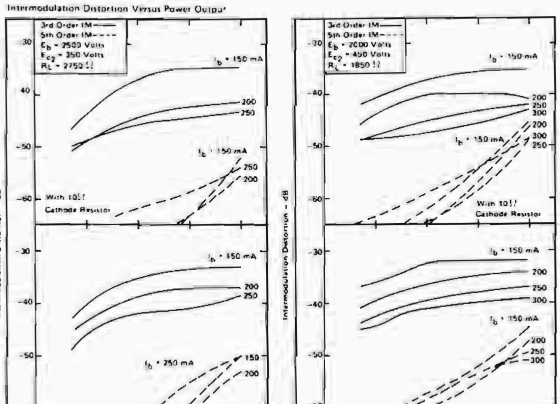


Fig. 35. Intermodulation distortion in linear amplifiers.

Fig. 36. Cermolox tubes (RCA).

Typical tubes for television service manufactured by RCA are shown in Fig. 36.

Typical 3-tone performance of power triodes suitable for translator service manufactured by Eimac are shown in Fig. 37.

To determine gain, power capability and dissipation, the same methods can be followed which were outlined in the previous paragraph dealing with intermediate power amplifiers.

Amplifiers over approximately 1kw must incorporate safety and protection circuitry not necessary for low power stages, since the destructiveness of failures is much more pronounced and costly in the higher power stages. Large power tubes are quite likely, especially during infancy to develop internal arcs (gas pings). The support circuitry (filament, bias, screen, and anode supplies) must be designed to be tolerant of such arcs and to arrest rather than maintain them.

To this end, the maximum current to be drawn from any supply should be limited to safe values. This value is different for all electrodes and varies with the internal construction of the tube. Maximum permissible values are supplied by the tube manufacturer. Often the user will find 1 to 10 ohm high wattage resistors

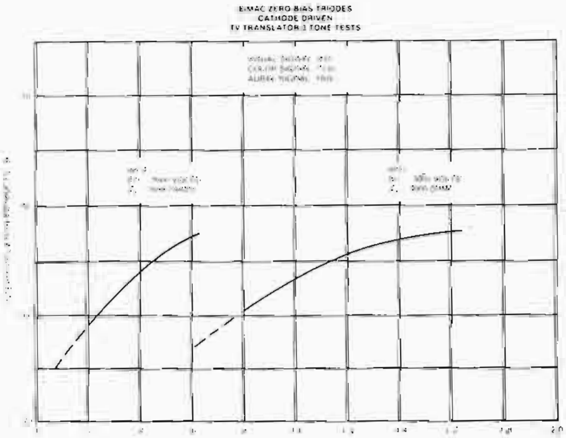


Fig. 37. Intermodulation distortion of 8877 and 8875 tubes.

(GLOBAR) in strategic places of the transmitter which perform this current, or energy limiting function. The application of supply voltages should not cause overshoots which often result from L-C type ringing in long supply leads. If the leads cannot be shortened, this L-C ringing must be damped by resistive means. Filament voltage should be applied slowly without exceeding maximum values defined by the tube manufacturer. The mechanical deformation of the tube filament between hot and cold states is a strong factor in determining the life of the cathode.

Tube life furthermore depends on operating temperature, and one should remember the rule of thumb (which incidentally, also applies to other incandescent devices operating in this general temperature range, for instance—light bulbs) that every 5% increase in supply voltage will cut life in half while each 5% decrease will double life, and so on.

Power tubes should be handled with kid-gloves (relatively speaking), especially when new and cold. The filament wires when cold are in their most brittle state. One should keep in mind that the tube elements are made from relatively soft copper and are not made of steel. (Even though they may look like steel.)

All power tubes listed on Table 1 are cooled by forced air. The final judgment of cooling effectiveness is anode core temperature. Therefore, if in doubt, one should measure anode core temperature in several places following the manufacturer's data sheet recommendation. For this purpose a temperature sensitive paint is most expedient and available from: TEMPIL, Hamilton Blvd., South Plainfield, NJ 07080. The suggested ranges are: 253°C, 226°C, 204°C, and 173°C.

Since the thermal output of a power tube under dc conditions can be measured quite accurately, a convenient method is available to measure airflow through the tube requiring only a thermometer. Assume a 8807 tube operated with 7kw anode voltage and idled at 1 amp (no RF drive applied) and the filament at 9.5 v, 145 amps. (The screen dissipation is negligible.) The tube is dissipating 8380 watts.

If the inlet temperature is 70°F, the outlet temperature (averaged over the width of the duct and measured approximately 4 to 6 feet away from the tube) is 115°F. The air flow in CFM is:

$$f_{CFM} = \frac{0.24\ w}{\rho\ \Delta T}$$

Where w = dissipated power in watts
ρ = air density (.075 lbs/f^3 at sea level and 68°F)
ΔT = Temperature in °F.

When values of the above example are inserted:

$$F = 600 \text{ CFM}.$$

Much useful information can be gathered from tube manufacturers application notes such as—AN-4869 available from RCA, Electronics Components, Harrison, NJ 07029.

AURAL TRANSMITTER

Aural Exciter/Modulator

The aural Exciter/Modulator will generate a 5-watt aural carrier, at the final frequency which carries the audio information, frequency modulated, onto the carrier. In addition a subcarrier can be frequency modulated on this carrier. This subcarrier constitutes a communications link from the transmitter back to the studio.

Fig. 38 is a block diagram of the aural Exciter.

The audio signal is available at a level of 10 dBm and will be terminated with the Exciters input impedance of 600 ohms.

The subcarrier amplitude is terminated in 2,000 ohms. A voltage of approximately 500 mv will cause ±2.5 kHz deviation of the aural carrier. The pump signal for the up-converter is provided by the visual Exciter and 5 mw are available into 50 ohms.

The aural Exciter/Modulator uses an IF approach. A modulated intermediate frequency carrier is produced at 32.5 MHz. The center frequency of this signal is controlled by a phase locked loop system.

The modulated oscillator which is placed in an oven with proportional temperature control operates directly at 32.5 MHz and employs a single varactor diode for frequency modulation. By maintaining a dc path through the diode an error voltage from the phase locked loop system will maintain the average (center frequency) within very narrow limits. The output of the modulated oscillator, after decoupling, is avail-

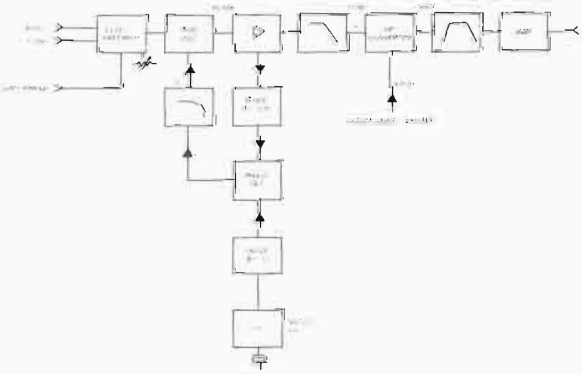


Fig. 38. Block diagram aural exciter/modulator.

able for AFC processing and up-conversion. To maintain the conversion process free of unwanted mixing products (as covered in detail in the Visual Exciter/Modulator section), a low pass filter is placed in the signal path before it is connected to the X port of the up-converter.

Again, the difference frequency is chosen and separated from the sum by a band-pass filter.[11] Amplification in a solid state amplifier is straight-forward and follows principles already explained in the visual Exciter/Modulator section. Biasing of the amplifier is chosen to operate all stages in a Class C mode as opposed to Class B for visual service.

The amplifier covers all high band VHF channels without further tuning and produces 46 dB of gain. A similar amplifier is used for all low band VHF channels. Class C operation makes it desirable to place a low-pass or band-pass filter at the output of the amplifier since higher levels of harmonic signals must be expected in a Class C amplifier. A 3-section (7-pole Chebycheff) bandpass with a bandwidth of approximately 15 MHz will assure sufficient harmonic attenuation.

Directional couplers and diodes allow to read forward and reflected power on a front panel meter.

The frequency response, FM signal-to-noise ratio and distortion are largely determined by circuit design of the modulated oscillator and can be maintained to very narrow limits. Stability of center frequency is determined by the control system chosen. Frequency control where the reference is a discriminator will always be several magnitudes poorer in stability than phase lock systems which only allow temporary departures from exact center frequency where the designer is able to choose the degree of timing of those departures freely.

The system employed in the Exciter/Modulator described here is a phase lock system. In such a system carrier deviation resulting from wanted modulation must be limited to the linear range of the phase detector or about $\pm 60°$.

The relationship between frequency and phase modulation is as follows:

$$\Delta f = \Delta \phi \times f_m$$

Where all fs are in Hz and ϕ is in radians.

If $\Delta\phi$ is limited to $\pm 60°$ or about $\pm$one radian, the Δf must be ± 50 Hz if 50 Hz is the lowest modulating frequency.

Consequently, the modulated oscillators frequency must be divided by 500 to reduce the normal ± 25 kHz deviation to ± 50 Hz or less. To allow for ± 50 kHz deviation used in other CCIR system and to provide for some margin of safety, a division ratio of 2,048 was chosen, easily obtainable with binary TTL dividers. This ratio will also place the comparison frequency just outside the aural passband (15.869 kHz). Broadcast systems require modulation capability at the low end of the audio range to 50 Hz, often to 30 Hz. The phase lock loop must leave these frequencies unimpaired and the in-loop, low-pass filter must be chosen accordingly. As a consequence, initial capture range of this system before lock up is severely restricted. Therefore, the in-loop low pass filter is automatically changed between search and lock up conditions, to allow a locking range of approximately ± 100 kHz. Once lock-up is achieved, the cutoff frequency of the loop low-pass filter is lowered so that low modulation frequencies can be handled without impairment.[16] Phase-locked loop systems have the impressive feature of very close frequency control and very small carrier shift with modulation. Accuracies of 100 Hz or less and shifts with modulation of less than 100 Hz are easily achieved in practical operation.

Phase lock systems used to suffer from difficulties with regenerative dividers. These difficulties have been totally eliminated by untuned dividers available in great variety as integrated circuits. The reference chain of the AFC system employs an oven controlled crystal oscillator whose frequency is so chosen to allow use of AT cut crystals (as opposed to NT cut crystals at lower frequencies), which exhibit much better temperature stability and ruggedness. The task of dividing this signal down to 15.869 kHz is easily accomplished in two ICs.

As the phase detector, a version was chosen which will produce zero output should one or both inputs disappear thus avoiding large center frequency shift in case of failures in one or both divider chains.

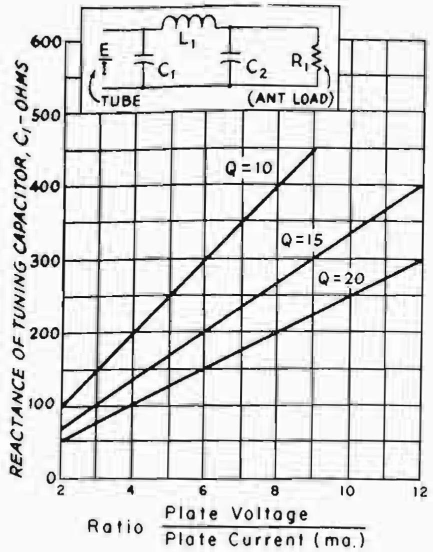
A single sensor will give a clear indication of proper lockup by comparing the phase detector input signals. Lockup is indicated by the front panel light.

Output power from the Exciter can be varied continuously by controlling the IF input voltage to the X-port of the up-converter through a motor driven attenuator.

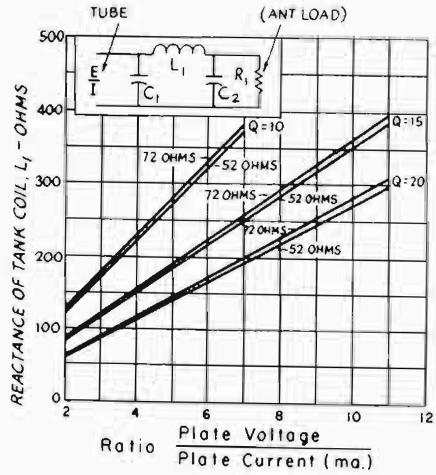
Aural Intermediate Power Amplifier

Amplification of frequency modulated, narrow band signals can follow widely accepted practices. For low power stages, e.g., up to several 100 watts, efficiency is not of primary concern and for this reason such amplifiers are often operated Class B rather than Class C since more gain is obtainable and fewer harmonics are produced.

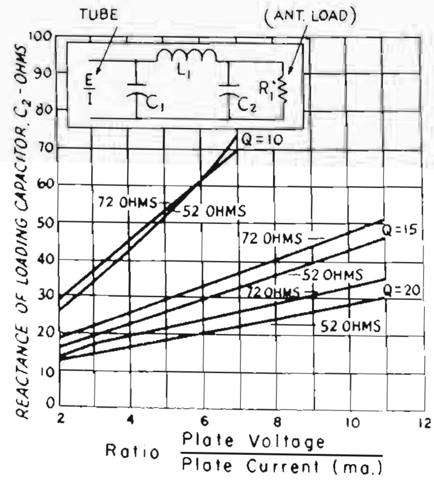
PI-NETWORK DESIGN CHARTS FOR FEEDING 52- OR 72-OHM COAXIAL TRANSMISSION LINES.



Reactance of input capacitor, C_1, as a function the ratio of plate voltage to plate current.



Reactance of tank coil, L_1, as a function of plate voltage and current, for pi networks.



Reactance of loading capacitor, C_2, as a function of plate voltage and current, for pi networks.

Fig. 39. Pi-section nomographs.

This reasoning is applied to a tube amplifier employing an 8122 tetrode to produce 250-watts output power with 5 watts drive, a gain of 17 dB. Operation as a cathode driven amplifier avoids the need for neutralization which simplifies the circuit and its operation. Input and output circuits are single tuned, using L and Pi sections.

The input capacity of the output Pi is produced by the tube. Tuning is accomplished by a parallel inductance which will tune-out a predetermined amount of the tubes output capacity.

One of the easiest to use nomographs for determining the circuit element values for Pi output sections can be found in the *ARRL Amateur Radio Handbook*. Fig. 39 is a reproduction of this nomograph.

The following example will illustrate the use of the nomograph for a tube type 4CX250B or 8122. With an anode voltage of 2 kv and a dc input of 500 watts, the anode will draw 250 ma of current leading to a voltage-to-current ratio of 8. At an assumed efficiency of 70% the output power produced will be 250 watts. The reactance values for the Pi network using Circuit Q of 15 will be as follows:

$$X_{C1} = 270 \text{ ohm}$$
$$X_{L1} = 275 \text{ ohm}$$
$$X_{C2} = 36 \text{ ohm}$$

which will lead at a frequency of 200 MHz or approximately Channel 11 to the following values.

$$C1 = 3.0 \text{ pf}$$
$$L1 = 210 \text{ nH}$$
$$C2 = 22 \text{ pf}$$

Attention must be paid to adequate bypassing, especially of the screen grids while properly considering the higher circulating currents encountered with higher average powers for aural service and possibly higher Circuit Q as opposed to visual service which will produce an average picture power of approximately 25% of the peak-of-sync power at which the transmitter is rated.

If the Q of the tuned circuits in the aural amplifier is only 15 or 20, no noticeable degradation of the modulation content of the signal will occur, especially with the small amount of deviation employed in television service.

The only noticeable effect of a power amplifier on an FM signal is simultaneous AM caused by the frequency modulated signals (FM to AM conversion). To minimize this effect for which no FCC limits are in existence, one should tune the amplifier to maximize the second harmonic of the AM envelope. This is easily done by detecting the aural carrier in an AM detector, e.g., a diode, or a diode noise meter (used to measure AM noise), feeding the output into a distortion analyzer which is tuned to remove the fundamental frequency of the tone causing FM modulation. This leaves the second (and higher) harmonics which can now be read on the meter of the distortion analyzer.

The anode tuning of the stage being checked should now be adjusted for maximum second harmonic output which will coincide with a minimum output at the fundamental frequency. At the same time the peak-to-peak amplitude modulated envelope of the frequency modulated signal will go through a minimum.

For amplifiers up to several hundred watts output power, grid bias may conveniently be obtained through a cathode resistor or a zener diode (which should be properly protected), or a combination of both. In this way a separate bias supply can be eliminated. Higher power amplifiers make this approach, however, wasteful. All tube stages in the television transmitters produced by Harris Broadcast Products Division operate grounded grid and grounded screen, thus, avoiding any need for neutralization. Less power gain is utilized this way, but in a television transmitter the tradeoffs are favoring the approach that we choose.

Input tuning of all aural amplifiers can be single-tuned using an L section to match the 50-ohm transmission lines to the 30-40-ohm input impedance of the cathode driven amplifiers.

Aural Power Amplifiers

TV transmitters employ aural power amplifiers up to about 8 kw in a single tube configuration. From several hundred watts up the use of lumped elements becomes quickly unattractive, striplines are employed for frequencies up to 100 MHz. Higher frequencies require anode circuits in the form of coaxial cavities, some designs in circular and some in square cross-sectional configuration.

Care must be taken with the very high unloaded Q values available (up to several thousand) and the comparative ease in which cavity loading can be changed not to operate the tube in a high-loaded-Q-output circuit, e.g., a Q of 100. This would lead to unacceptably high circulating currents which could do damage to the screen contacts and possibly the tube seal. A loaded Q of 15 should be selected. This can be easily checked (since sweep equipment is generally available in a TV station) by driving the aural power amplifier from the visual exciter, with the vestigial sideband filter and delay compensator disabled. Under this condition, loading (and therefore, loaded Q) for a bandwidth of f_C/Q at the 3-dB points can be adjusted, e.g., for Channel 11 one should seek a 3-dB bandwidth of 13 MHz. This will mean, in essence, that the response of the aural power amplifier at the visual carrier is not quite -3 dB. For this test care should be exercised not to operate the amplifier in a saturated condition. The best way to assure this is to operate it at 20% to 50% of normal power. If sweep equipment is unavailable, one can first drive the aural power amplifier from the aural exciter, driving it to 50% of normal power and measuring the drive power necessary. Then the aural exciter is replaced with the visual exciter (with visual gain control counterclockwise), driving the aural power amplifier with the same power originally obtained from the aural exciter. At this point the power output from the aural power amplifier should not fall below half the power obtained when it was driven by the aural exciter. This would indicate that the response at the visual frequency is 3 dB down from the response at the aural carrier.

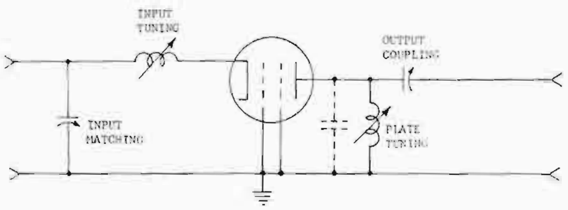


Fig. 40. Circuit equivalent, aural power amplifier.

Assessment of proper operation as far as power output, efficiency, and load impedance are concerned should follow the reasoning expounded in the previous paragraph dealing with low power stages.

The main difference in a high power stage is in the greater effort required for circuit protection against arcs and other mishaps in order to limit the destructiveness of such malfunctions. A larger resistor for limitation of short-circuit currents can be used in the aural amplifier since the constant load of this amplifier better tolerates a high impedance anode supply. Most modern TV transmitters have extensive provisions for remote control and remote reading of typical parameters. For this reason, the anode current of the aural PA is transformed to ground potential through the use of a magnetic amplifier. The equivalent circuit of an aural power amplifier is shown in Fig. 40.

Control Circuits

The control circuits of the television transmitter perform the following function:

1. Upon receipt of suitable command turn transmitter on through proper starting sequence.
2. Upon receipt of command, turn transmitter off.
3. Restart transmitter in proper sequence following short or long power line interruptions.
4. Provide status indication for various transmitter conditions.
5. Provide corrective action and indication derived from sensing an overload condition within the transmitter.
6. Cause partial or complete system shutdown in case of failure.
7. Allow system expansion, such as dual transmitter configuration.
8. Provide interface, in case of remote or unattended operation, to other functional elements of the overall system.

The traditional way to design control circuits was by using mostly telephone type relays to build up logic circuitry in accordance with established requirements. These relays provided two logic states: 0 and 1. To implement elaborate control circuits often required the use of dozens of such relays.

More elegant solutions are available now in the form of Solid State Logic building blocks, avoiding the use of electro-mechanical elements for logic functions entirely. Power switching, however, will still be performed by conventional contactors. Each cubicle of a particular transmitter contains its own control circuit and can be independently activated from switches on that cabinet. This feature was incorporated to allow easier system maintenance and tune-up. The entire transmitter may also be operated from the first cabinet.

Some of the features of each control circuit include filament warm-up time-delay, air pressure interlocks, overload cycling, automatic reactivation with ac failures and automatic high voltage removal in any power amplifier which has had three overloads.

The four basic digital logic circuit elements employed are the AND gate, the OR gate, the BINARY and the MONOSTABLE MULTIVIBRATOR. The equivalents of these circuits can be seen in Fig. 41. The two types of gates replace the usual series-wired relay contacts, or the parallel wired contacts while the binary replaces the usual latching relay. To illustrate the basic operation of these circuits, three opera-

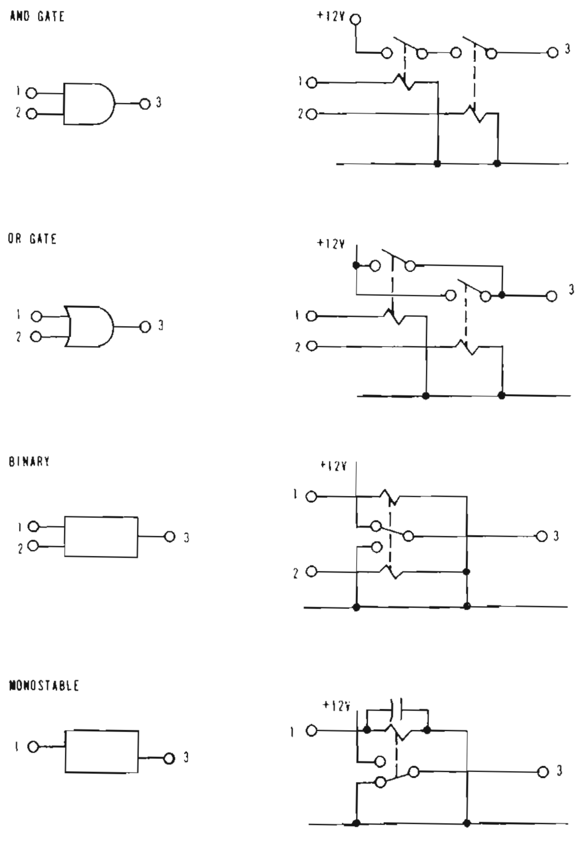


Fig. 41. Logic circuit equivalents.

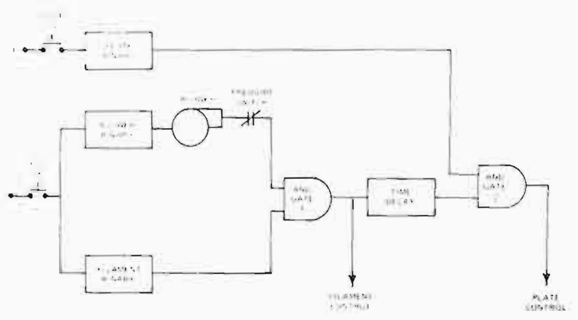


Fig. 42. Turn-on block diagram.

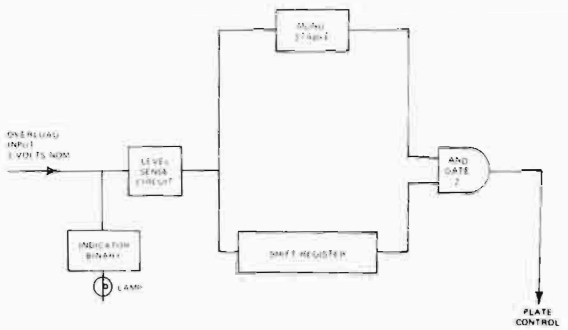


Fig. 44. Overload block diagram.

tional sequences are described in Figs. 42, 43, and 44.

Turn-on: Fig. 42 shows basic turn-on circuitry. A momentary input pulse, provided by the front panel filament ON pushbutton changes the state of the filament and blower binaries. This action starts the blower and applies the positive logic level at one input of AND Gate 1. When air pressure is adequate, a positive logic level is applied to the other input of the gate. With both inputs positive, the output of the gate goes positive, turning on the filaments and starting a time delay circuit. At anytime, the plate ON pushbutton can be depressed. The state of the plate binary changes immediately to the positive logic level. But if the time delay has not elapsed, anode voltage cannot be applied. If the plate ON pushbutton has been pushed and the time delay has elapsed, both inputs of AND Gate 2 will be in the positive state and anode voltage will be applied.

Transmitter Turn-Off: In Fig. 43 the basic turn-off circuits are illustrated. Plate control may be removed by momentarily depressing either the plate or the filament OFF pushbuttons. Again, only a momentary voltage at the plate or filament binary OFF inputs will change the state of the binary, and by reducing the input of AND Gate 2 to zero will remove plate control.

If the filament OFF button is depressed, filament control as well as plate control will be removed. In addition, a time delay circuit is activated which, after one minute, changes the state of the blower binary and removes blower control.

Transmitter Overload Features: Fig. 44 illustrates circuitry used to provide overload protection for the system. Overload inputs to the control circuits consist of dc voltages related to the system status. When an overload circuit develops a nominal 3-volt signal, the logic circuits respond. This nominal 3-volt signal is used to activate an indicator binary and a level-sensing circuit. The indicator binary changes status, driving a lamp which identifies the source of the overload. The level-sensing circuit delivers a negative-going output pulse which triggers both a monostable multivibrator and a three-stage shift register.

The monostable, when triggered, holds the input of AND Gate 2 in the zero state for one second and, thus, removes plate control for that period of time. The shift register counts the number of overload pulses and after three pulses are applied, holds the other input of AND Gate 2 at zero, and removes plate control.

The transmitter is, therefore, deactivated for a short time when an overload occurs and is disabled after three repetitive overloads. A flashing lamp on the meter panel indicates when three overloads have occurred.

Because all ON and OFF functions are achieved with the application of momentary voltage pulses, remote control is simplified. The current pulse required is small enough to permit easy interfacing with computer control devices. In addition, the voltage levels used to indicate system status can be easily sampled and computer analyzed. The use of solid-state circuits for controlling television transmitters yield numerous benefits to the user. The first is increased

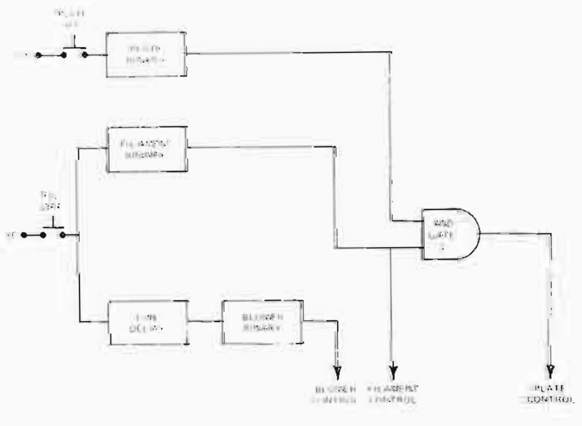


Fig. 43. Turn-off block disgram.

reliability; no relays are used to perform logic functions. Second, reduction in physical size of control circuits helps to decrease the total size of the system. Third, the simplification of intercabinet wiring makes initial system installation and, later, system maintenance much easier. Fourth, flexibility is enhanced. Fifth, remote control is easily interfaced.

Power Supply Circuits

A television transmitter requires a number of supply voltages as shown below:

1. Ac or dc filament voltages typically 5-10 volts, 1-200 Amps.
2. Bias supplies 10-250 volts, 0-300 milliamps.
3. Screen supplies 200-2,000 volts, 10-150 milliamps.
4. Anode supplies 1,000 to 8,000 volts, 100 milliamps to 15 Amps.
5. Ac or dc auxiliary supplies, e.g., for solid-state circuits and amplifiers, for logic circuits, indicator lights and power contactors, 12-28 volts and 1-5 Amps.

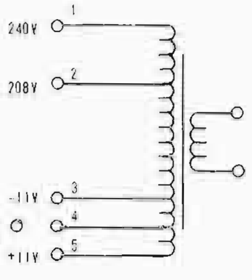
Filament Supplies

Filaments requiring ac voltage offer the simplest condition. Sometimes special conditions, e.g., limit in maximum current, or high voltage insulation, may lead to nonstandard designs. The ac supplied filaments are generally preferred due to their ruggedness and simplicity. If ac voltages cause undesirable effect, like 60-Hz hum modulation, dc supplies with varying degrees of ripple and regulation can be used. Care should be exercised not to overspecify the supply requirement as it leads to poor economy and reliability.

The degree of allowable ac ripple and the ripple frequency should be the determining factors for choosing half-wave or full-wave single-phase or multiple-phase circuits, since filtering by means of a low pass filter is quite impractical due to the very low load impedance encountered.

Most filament transformers used by the Harris Broadcast Products Division have input taps to allow adjustment for different input voltages from 197 volts to 251 volts in 5% steps. A typical arrangement is shown in Fig. 45. If the transformer is designed for 50-Hz operation, it will in tap position +11 and 208, accommodate the most prevalent voltage of 50 Hz systems: 220 volts.

The dc supplies should always use current limiting circuitry to avoid damage to wiring harnesses in case of a short at or near the load. Typical filament requirements for power tubes are: 10 volts at 150 amps.



VOLTAGE	TERMINALS
197V	2-3
208V	2-4
219V	2-5
229V	1-3
240V	1-4
251V	1-5

Fig. 45. Schematic of typical filament transformer.

Bias Supplies

The most important requirements for bias supplies are constant voltage with widely varying positive and negative loads. A negative load is, in essence, another supply in parallel with the existing supply whose current must be sunk into the existing supply. In a television transmitter, load variations occur at a very slow rate (camera pan or change in scene lighting), also at a frame rate (60 frames per second) and at a line rate (15,750 lines per second, or 525 lines per frame), making it necessary for the bias supply to be able to respond over a frequency range from dc to 4 MHz. Over this range the internal impedance, if not zero, should be preferably constant or change slowly and show no resonances.. Modern power tubes, fortunately, ease the requirement on the power supply by drawing only moderate amounts of grid current.

The most widely used, if somewhat wasteful, approach to accommodate grid current is to draw a steady current from the supply, about 30% higher than the maximum grid current expected, which will then "outbalance" the grid current. The bias supply will see a gradually decreasing load as grid current is increased. Grid current in excess of the maximum bleeder current will increase the bias voltage to more negative values which will tend to create a self-balancing situation if the bias supply components can stand the additional stress. However, serious waveform distortion is caused by this behavior, largely on the synchronizing pulses.

A number of triodes are available which can be operated with zero bias between grid and cathode. The lack of a separate bias supply for these tubes is one of the major advantages. Typical tubes are listed in Table 1.

Bias supplies generally range from 10 to 250 volts at currents up to several hundred milliamps. A limited adjustability of the voltage is a requirement for most supplies.

Screen Supplies

The foremost requirement of a screen supply is a constant voltage characteristic up to a defined maximum current, at which it reverts into a near constant current supply as shown in Fig. 46. During the constant voltage portion the operating voltage should be stable to ±1%, and the requirements of internal impedances outlined under Bias Supplies also apply. Voltage should be adjustable approximately ±15% from a center value.

Great care must be exercised by the designer to make the screen supply impervious to high spikes and overvoltages caused by gas pings in new power tubes. This dictates the use of thermionic devices at the interfaces with the power tube as protection necessary for solid-state devices would make it uneconomical and unreliable to use solid-state devices directly interfacing with high power tubes. The increased component count, furthermore, would decrease overall reliability. A typical simplified circuit is shown in Fig. 47. This circuit has the additional advantage of slaving the screen voltage to the (well regulated) bias supply so that a failure of the bias supply will automatically reduce the screen voltage to near zero, this way fully protecting the tube.

The function of the supply can be easily understood. If the voltage of the anode of the shunt-regulated tube tends downward, the grid voltage of the tube will get more negative, counteracting the original downward move. Amplifier A provides sufficient gain to maintain high enough loop gain for ±1% regulation on the assumption that the reference supply remains at approximately ±.2%.

Arc Gap E1 protects the amplifier tube and its screen blocker against sustained overvoltage, especially in case of failure of the shunt regulator tube.

In executing the various requirements of the screen supply, care should be exercised to keep the total energy stored in the supply below safe values. The permissible amount of energy (in Joules or Watt seconds) is obtainable from the

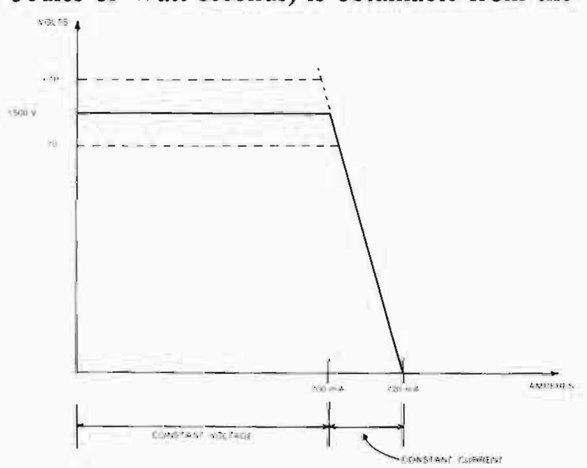


Fig. 46. Screen supply, load characteristic.

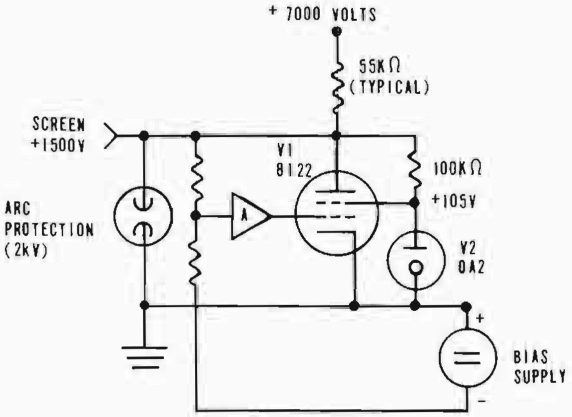


Fig. 47. Simplified schematic, screen regulator.

tube manufacturer. Any capacitors connected directly to the tube elements contains stored energy according to: $E = \frac{C \times N^2}{2}$ (Joules). In addition, energy will be supplied for a short time until the high voltage supply is disconnected from the power line (in a typical case, 50 milliseconds).

The total energy in a typical case may be:

$$E\text{total} = \frac{1.2 \times 10^{-5} \times 1.5^2 \times 10^6}{2} + .12 \times 5 \times 10^{-2} =$$

$$1.35 \times 10^1 + .6 \times 10^{-2} = 13.5\,\text{ws}$$

This is a very safe value for a power tube in the 15-kw plate dissipation category.

Anode Supplies

This supply provides the anode voltage to all power tubes. When thermionic devices were used as rectifiers, designers were forced to use choke input type supplies to limit the maximum instantaneous current to safe values for those devices. With the advent of solid-state rectifiers these limitations no longer applied as the maximum instantaneous current capabilities of silicon rectifiers were orders of magnitude higher than those of thermionic devices.[17]

Single phase capacitor input supplies require slightly larger power transformers (due to higher I^2R losses). Their main advantage for TV applications is excellent, frequency independent regulation, leading to almost total lack of amplitude disturbance during the vertical interval. Supplies of this type require very high filter capacitors, the major uncertainty relative to overall reliability. Data gathered over a five-year period on the operation are favorable, but not as yet fully conclusive.

All capacitor input supplies over approximately 1 kw should employ a starting mechanism

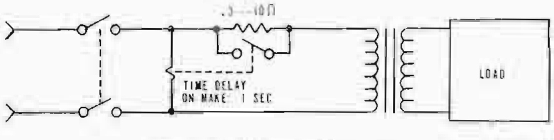


Fig. 48. Stop-start circuit.

to limit in-rush current. A scheme extensively employed is shown in Fig. 48.

The dropping resistor and time delay relay are chosen such that with a 1 second delay the output voltage under load will reach approximately 60% to 70% of the final value before the resistor is shorted out.

Supplies for power stages over approximately 1 to 2 kw are fed from three phase ac lines. For reasons stated above, capacitor input supplies are also considered. TV service with its widely and continuously varying load is demanding on power supply performance. Therefore, we choose three-phase, full-wave cascaded supplies providing 12 current "pulses" 30° apart during one cycle of the 60-Hz input period, and are therefore classified as 12-phase supplies.

From schematic Fig. 49, it becomes obvious how this is accomplished, namely, by using two full-wave three-phase supplies in a series with two secondary windings of extended delta configuration on a common transformer core.

For perfect input phase balance the ripple frequency is predominantly 720 Hz. (12 times 60 Hz). Input line phase imbalance will cause 120 Hz ripple to appear. Capacitive filtering, following the rectifier will suffice to reduce 720 Hz and its harmonics to very low levels, since without any filtering whatsoever, ripple amplitude is already at about −40 dB relative to the dc output voltage.

If serious phase imbalances of the primary feed are to be expected, line voltage regulators can be used to correct the imbalance. If this is not feasible or practical, the addition of a low pass filter to any 12-phase supply will reduce the 120-Hz component. Even with the additional filtering the 12-phase supplies still provide superior performance since very small chokes

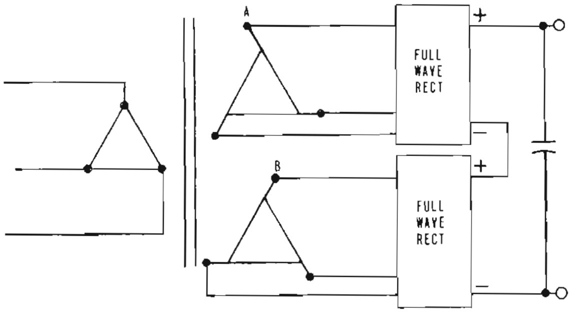


Fig. 49. Twelve-phase supply.

and capacitors are required (e.g., 1 Henry and 2 to 10 microfarads).

An extremely important element in high power anode supplies is protection of the supply itself and circuits connected to it.

The following protective elements can be found in typical supplies:

1. Circuit breaker of fast response in the input line;
2. Contactors of fast response which can turn the supply off through an external command, e.g., from within the transmitter resulting from a tube overload;
3. Arc Gaps in strategic locations to avoid overvoltages;
4. High energy resistors to limit peak currents especially of short duration to safe values; and
5. Capacitor-resistor combinations to suppress line produced and related spikes.

When properly designed, such a supply shall withstand repeated direct shorts across its output High Voltage terminals. It should also withstand line surges of at least two times normal line voltage. A typical supply is shown in Fig. 50.

Ancillary Supplies

Supplies for low voltage applications to feed, for instance, solid state RF amplifiers, discreet transistor circuits, and integrated circuits should incorporate two important parameters:

1. Fold back current limiting,
2. Overvoltage crow-bar.

Both requirements are designed to avoid destruction of delicate solid state devices. If an integrated circuit is subjected to small overvoltages, e.g., +30%, many such devices can be lost, necessitating, in large systems, very elaborate repairs.

Numerous designs for such supplies are available in the marketplace, and it makes it uneconomical, in most cases, not to use a vendor furnished power supply. Extensive design information is also available through application notes of many solid state manufacturers for IC Voltage regulators, and there is no need to elaborate on this further.

Combining Circuits

Two different types of combining circuits must be considered:

1. Combining of sources with the same frequency;
2. Combining of sources with different frequencies.

For Application 1 90° hybrid couplers have found almost universal acceptance. It is, for TV power applications, a relatively new component

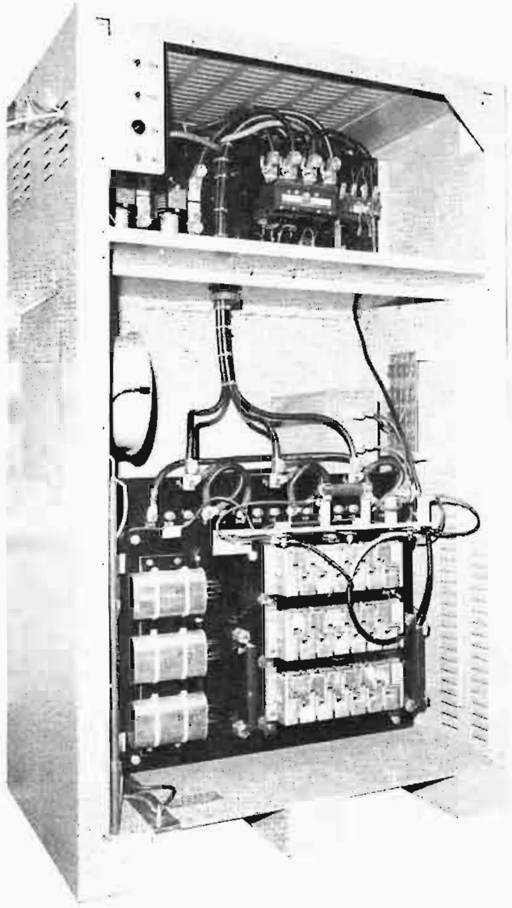


Fig. 50. Photograph power supply BT25H1.

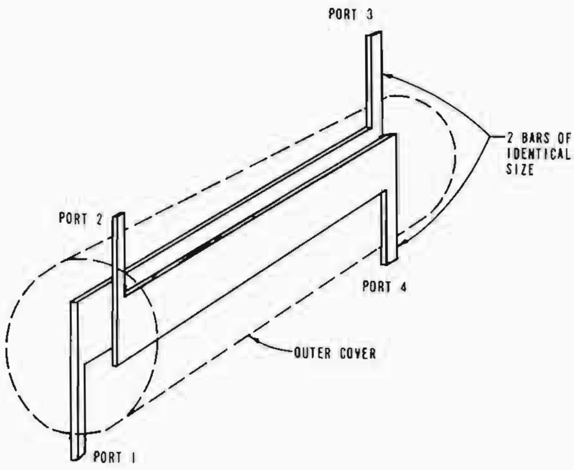


Fig. 51. Physical Model of 90° hybrid coupler.

and it's operation may not always be fully understood.

A detailed mathematical treatment with extensive other test and application data is found in Reference 18.

The following are highlights from Reference 18 which should help to visualize the elegant performance of this device. Fig. 51 shows an exploded view of a typical 90 ° hybrid coupler. The arrangement for Ports 3 and 4 is selected as shown for the ease of interconnecting the device. Both ports could be reversed in direction if this were more advantageous.

A 3-dB coupler consists of two identical parallel transmission lines coupled over a length "d" equal or approximately equal to $\lambda/4$ and mounted in a common outer conductor.

The construction is symmetrical, i.e., both inner conductors have the same physical dimensions and have the same capacitance per unit length with respect to the outer conductor.

The 3-dB coupler can be used as a power-splitter or as a power combiner. If the coupling attenuation at the center frequency f_0, corresponding to $4 = \lambda/4$, is exactly 3 dB, the voltage distribution as shown in Fig. 52 will be obtained. Normally 3-dB couplers are designed for a coupling attenuation slightly lower than 3 dB in order to minimize the voltage differences as measured at Ports 2 and 3 over a broad frequency range.

Compare Fig. 52 for a perfect 3-dB coupler with Figs. 53 and 54 for a coupler of 2.95- and 2.9-dB coupling, respectively.

PROPERTIES

1. The characteristic impedance Z_0 of the coupler is identical at all ports. This characteristic impedance is frequency independent and is solely determined by the geometry of the coupling region. Thus, with Z_0 connected to Ports 2, 3, and 4, the impedance as seen in Port 1 is also equal to Z_0. Furthermore, it is still equal to Z_0 at f_0—the center frequency—if the output Ports 2 and 3 are terminated in identical mismatches (phase and amplitude) as long as Port 4 remains properly terminated. Fig. 55 shows the mismatch as measured at Port 1 for the case where Ports 2 and 3 are terminated in identical mismatches of different magnitude. The worst case shown is when Ports 2 and 3 are open circuited or short circuited. It is apparent from this curve that the VSWR remains less than 1.05 for a frequency range of $\pm 14\%$ about the center frequency f_0.

2. The phase-shift between the two outputs is always 90° and is independent of frequency. If the coupler is used to combine two transmitters, then these two signals have to be fed to the 3-dB coupler in quadrature.

3. Irrespective of frequency no voltage will appear at Port 4, the isolated port, as long as

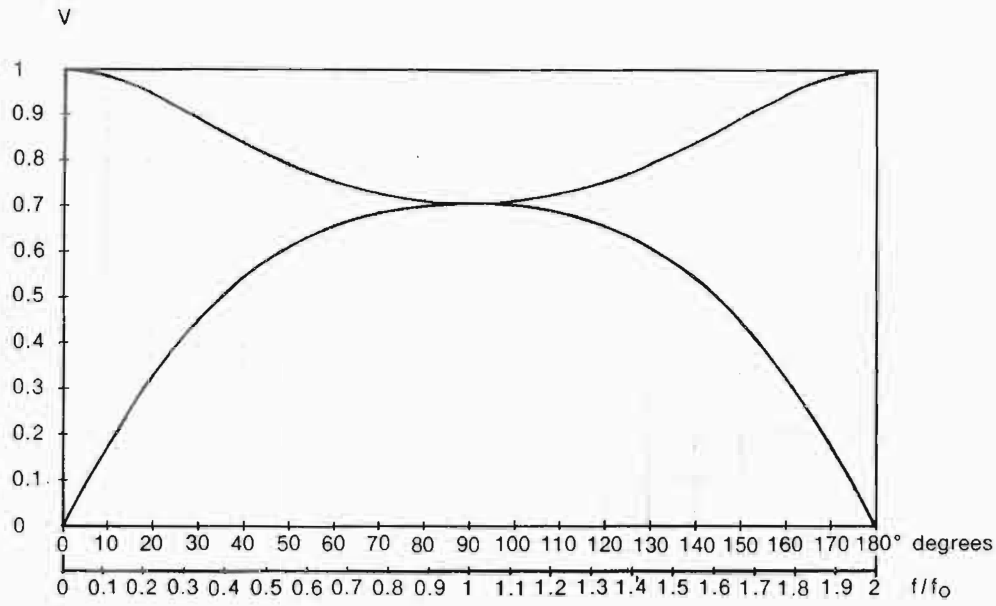


Fig. 52. Normalized voltage distribution of 3 dB hybrid coupler.

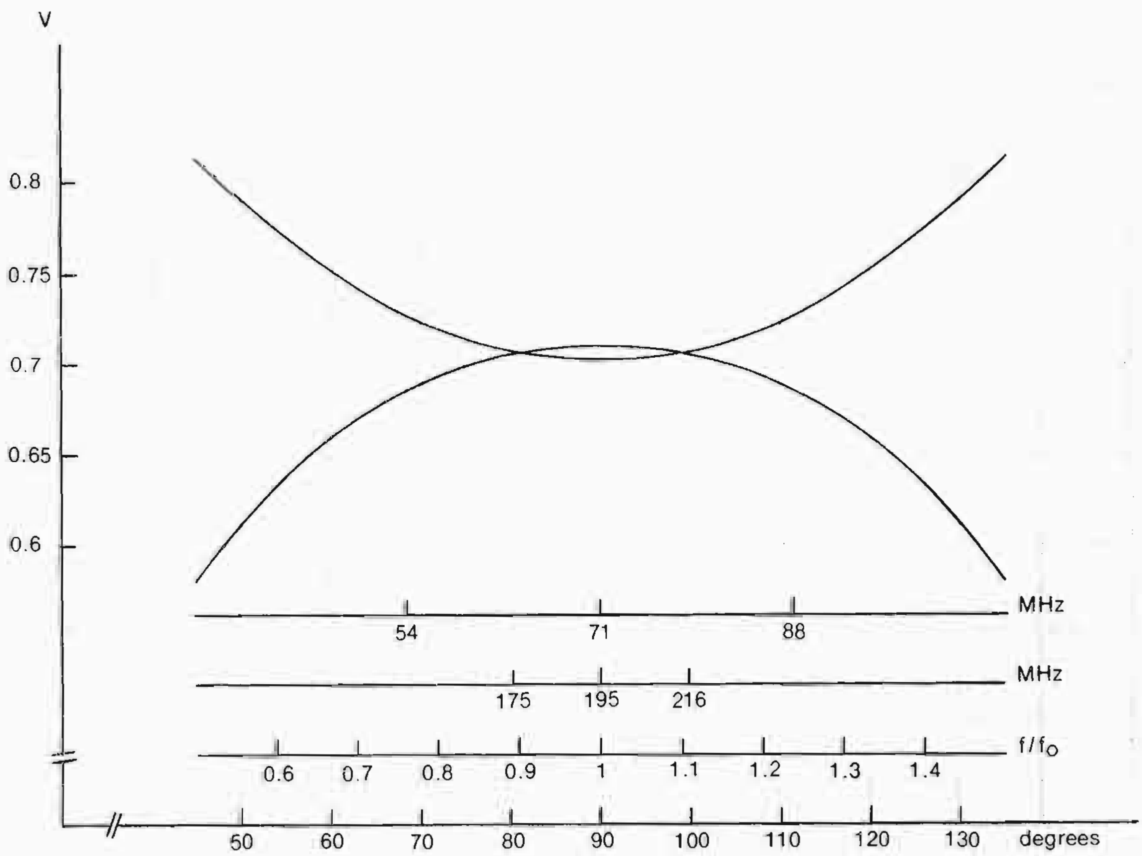


Fig. 53. 2.95 dB hybrid coupler.

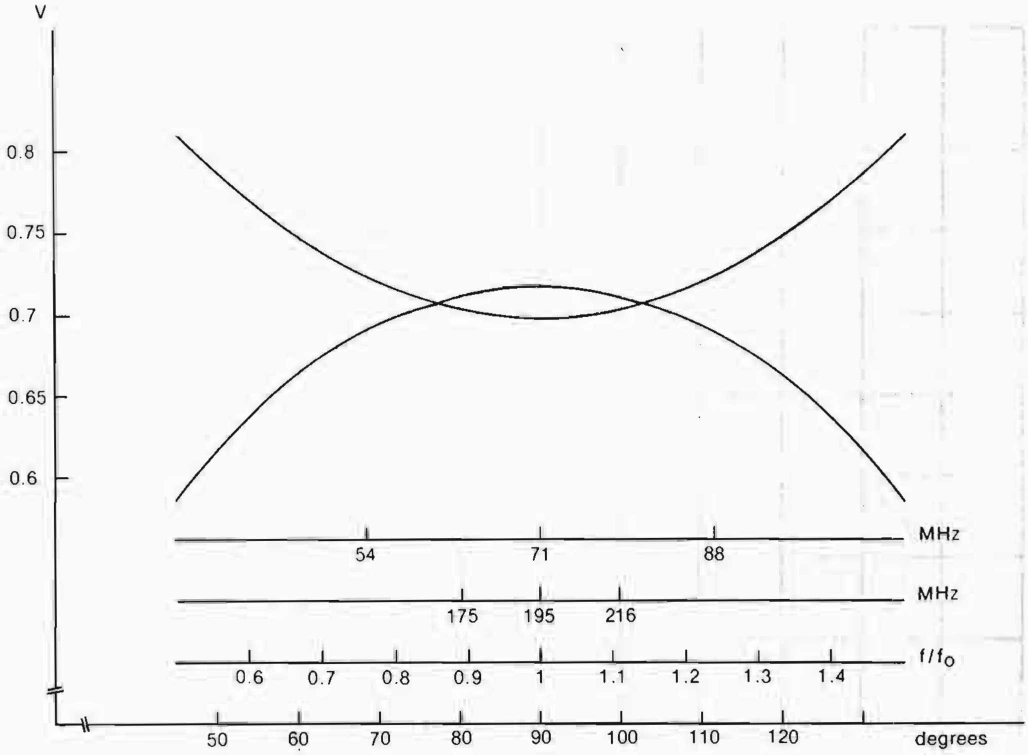


Fig. 54. 2.90 dB hybrid coupler.

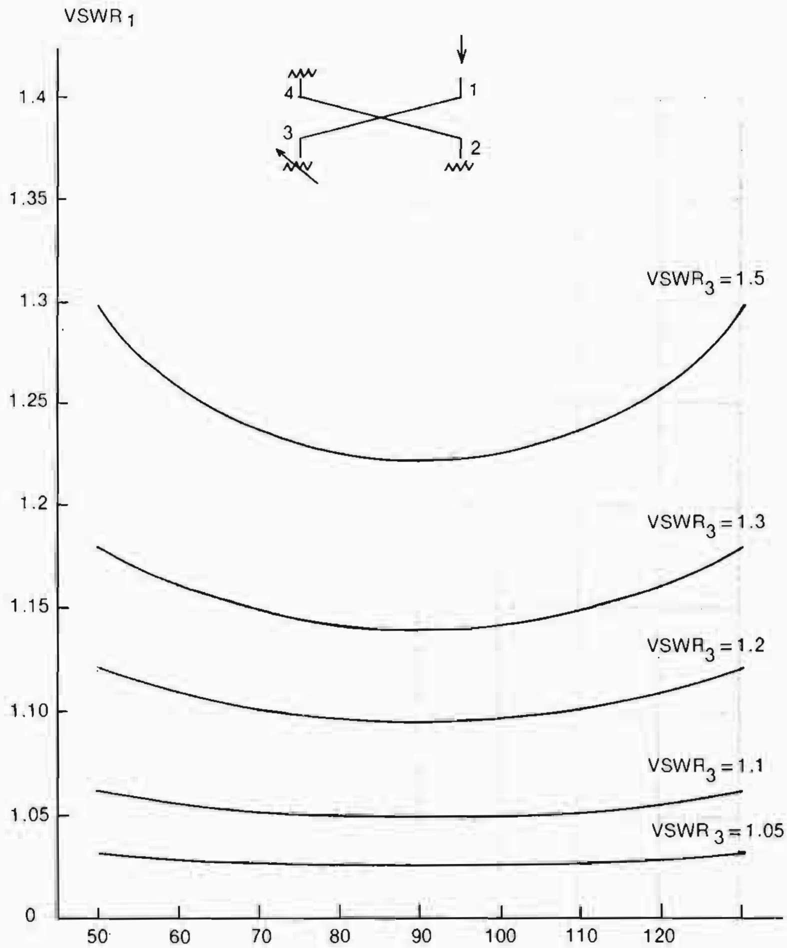


Fig. 55. VSWR of hybrid coupler.

output Ports 2 and 3 are terminated in their characteristic impedance when used as a power divider, or as long as Port 1 is properly terminated and both input Ports 2 and 3 have voltages of equal amplitude that are in quadrature.

4. The isolation between two ports, for example, Ports 1 and 4, depends heavily on the VSWR of the loads at Ports 2 and 3. If the load at Port 2 is an ideal load, the curve of Fig. 56 will then give the isolation as a function of the VSWR of the load at Port 3.

Example: If the load at Port 3 measures 1.05, then the theoretical isolation is approximately 38 dB. If the load measures 1.2, then the isolation cannot be better than 26.5 dB.

Couplers of this general mechanical form can be designed for a wide range of coupling factors and they are readily available for coupling ranges from 3 to 40 dB.

As power splitters or combiners, the 3-dB hybrid couplers are used predominantly but 4.70-dB hybrid couplers have been employed where the sum of three equal sources is combined by first summing two sources in a 3-dB hybrid and then combining the sum of the first two sources with the third source through a 4.70-dB hybrid coupler.

Fig. 57 shows the effect of two couplers connected "back-to-back." If two equal amplifiers are placed into the two connecting lines between the couplers, a nearly ideal arrangement for combining equal frequency sources is available. If, however, both amplifiers differ in phase, a reduction in output power will result, as Fig. 58 demonstrates. But, even an error of 60° (very unlikely to occur) will only cause a 25% reduction in power.

Inequality of amplifier gain, in other words different output powers, will have the effect shown in Fig. 59.

If one amplifier has only half the output power of the above, i.e., A-1 = 10 kw, A-2 = 5 kw, only 3% of the total 15 kw available is dissipated in the reject load, i.e., 450 watts. Ninety-seven percent is still transmitted to the load.

The conclusion is that hybrid splitters and combiners in the above arrangement are very tolerant of phase or power inequalities of the amplifiers operating in parallel. The second application for 90° hybrid couplers is the combining of sources of different frequency, especially where the two frequencies are not far apart and, therefore, well within the bandwidth of the coupler; e.g., in case of visual and aural transmitters of a TV transmission system.

Again, two distinct areas exist: (1) systems employing dual feed lines to the antenna, e.g., Batwing VHF antennas and (2) systems em-

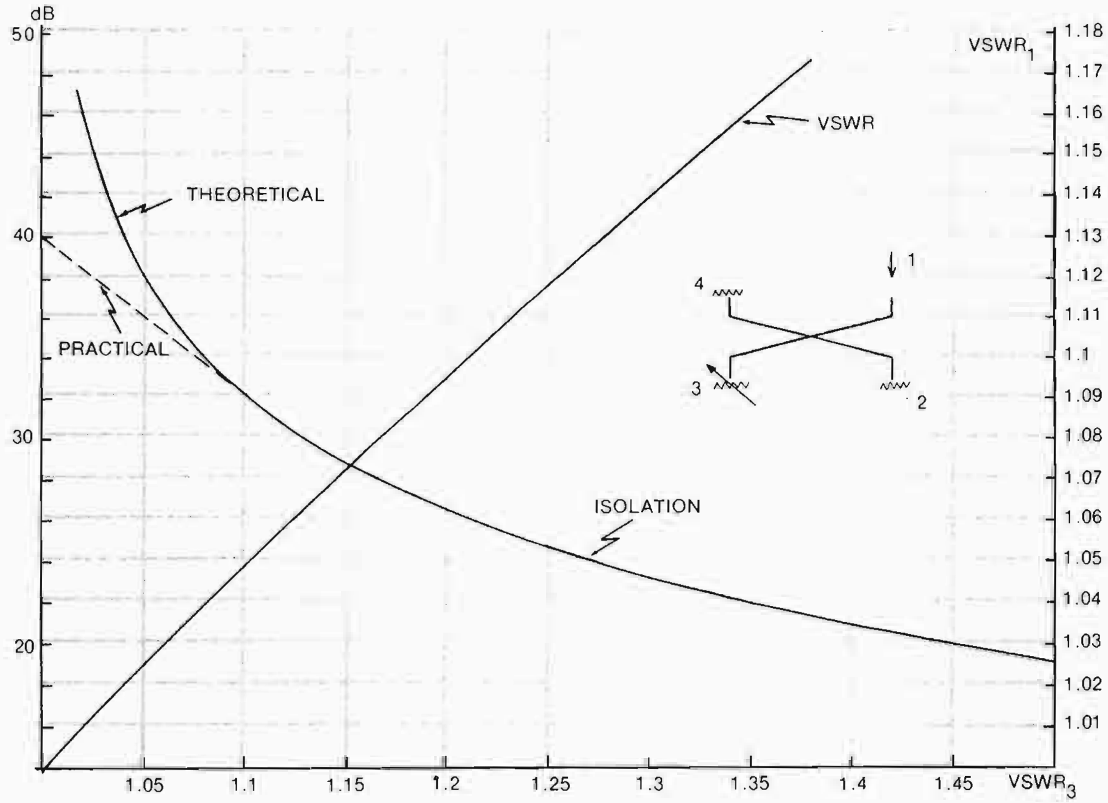


Fig. 56. Isolation of hybrid coupler.

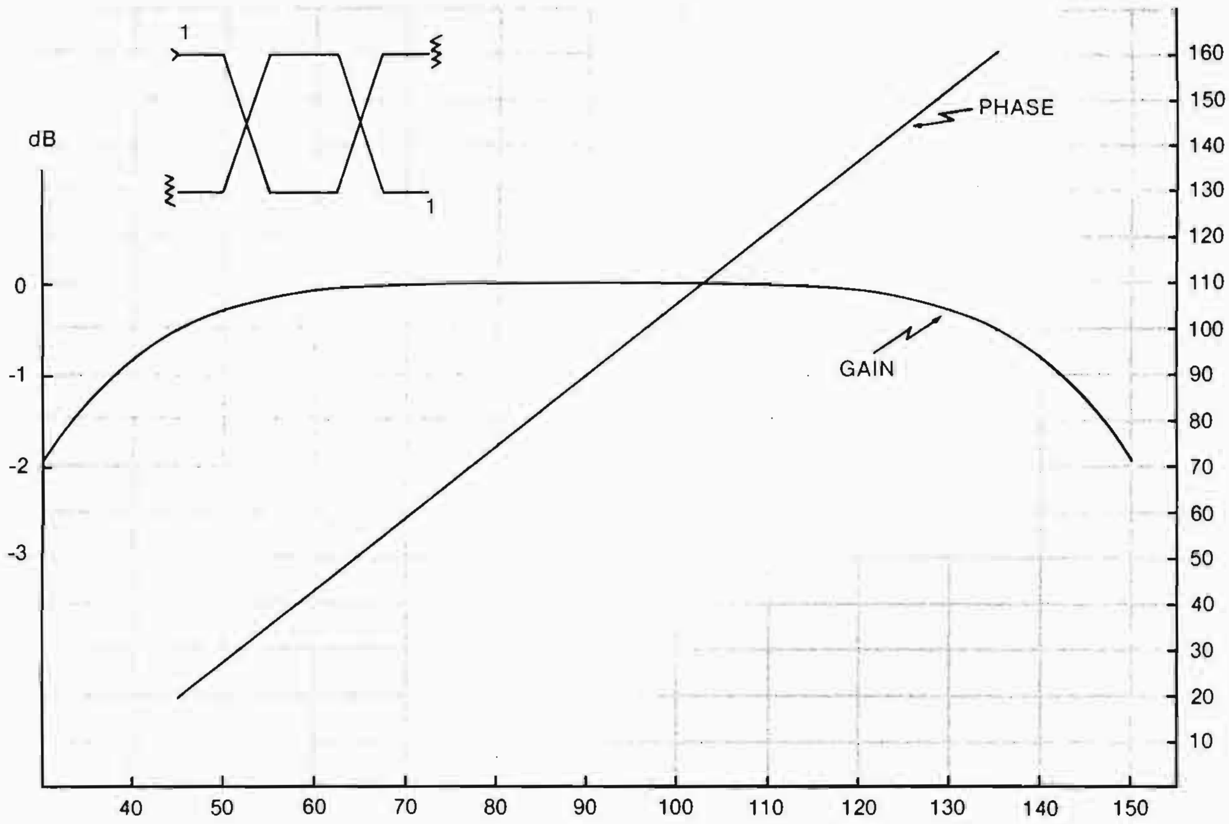


Fig. 57. 90° hybrid couplers as power splitters and combiners

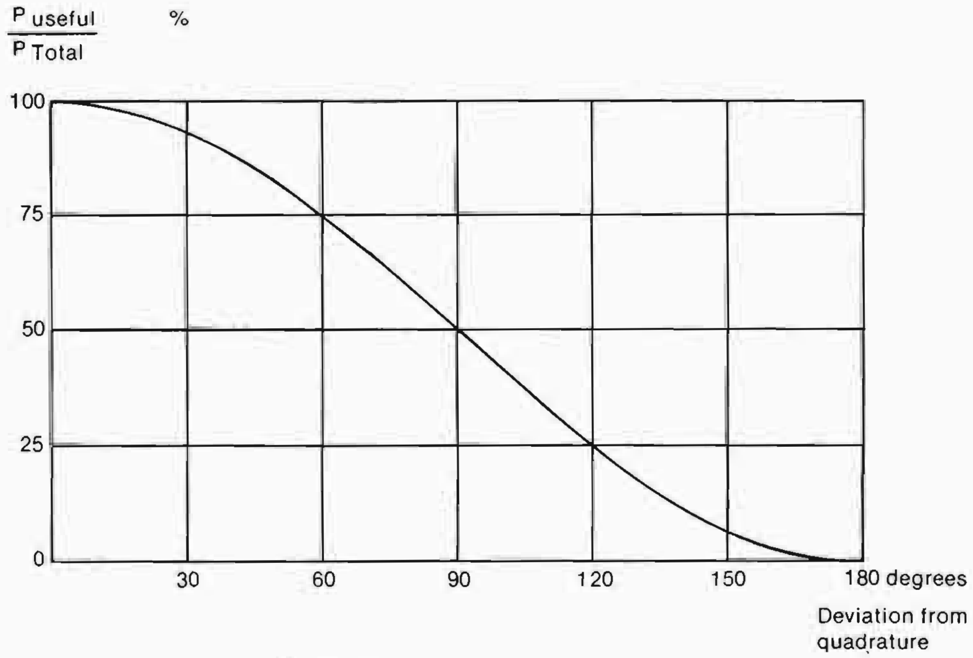


Fig. 58. Phase sensitivity, hybrid coupler.

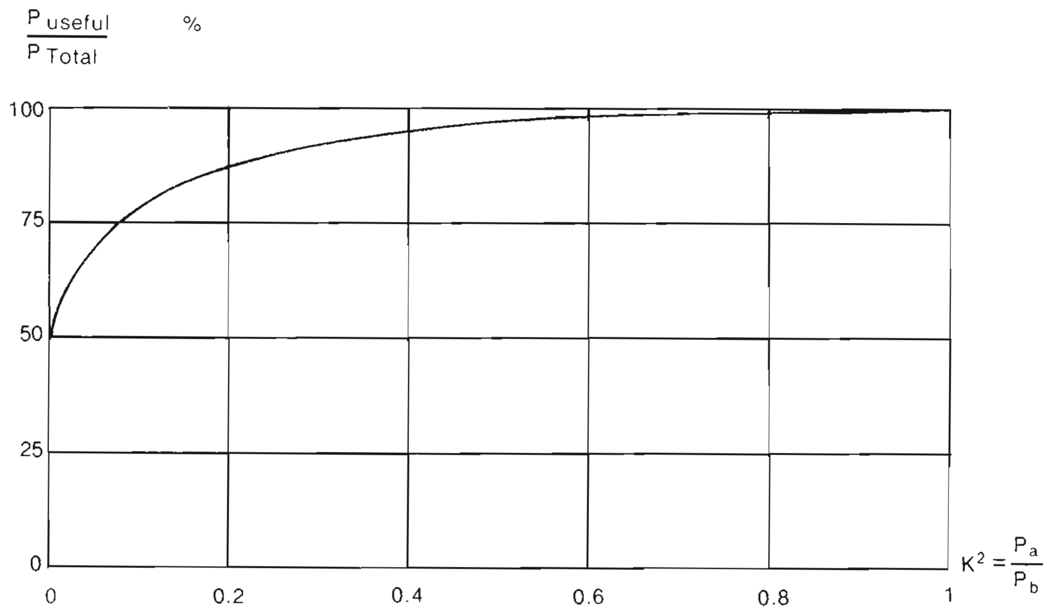


Fig. 59. Power imbalance in hybrid couplers.

ploying single antenna feed lines. The first application is very simple: Connect the visual transmitter to Port 1 of the 3-dB 90° hybrid, connect the aural transmitter to Port 4 of the same hybrid, and connect the two feed lines to the remaining Ports 2 and 3 of the coupler.

Isolation between visual and aural transmitters can be extracted from Fig. 56 if the antenna line VSWR is known, and figures of 30 dB are easily attainable, which will satisfy the requirement of any transmitting equipment. *Case 2:* "Single antenna feed line" is considerably more complicated and creates problems for which compensation must be applied elsewhere in the system.

A simplified schematic of a Visual/Aural Diplexer is shown in Fig. 60.

The aural transmitter is connected to Port 1 and the Antenna to Port 4. The Visual transmitter is connected to Port 4' and the reject load to Port 1'. Between the two 3-dB couplers, two high Q series resonant circuits are connected in parallel with the main lines. These are shown as a series circuit of L_1 and C_1. This combination is in resonance for the Aural Carrier Frequency.

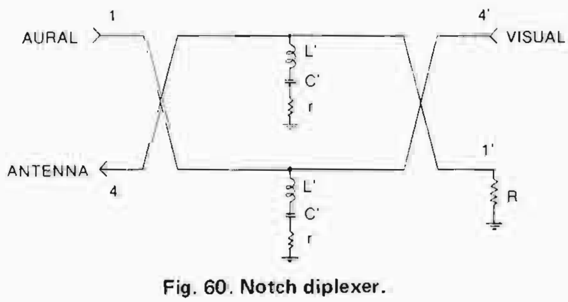


Fig. 60. Notch diplexer.

With the aural transmitter connected to Port 1, all power will be available at Port 4 since the series resonance circuits are a perfect short circuit for the Aural Carrier frequency, and no power will be transmitted to Port 1' or dissipated in the cavities.

Since a practical series resonant circuit cannot have infinite Q, the cavities can be represented with a small resistor r at aural carrier frequency. Some of the aural power will now be dissipated in this r and some will also be transmitted to Port 1'.

To have an idea how much power is dissipated in the cavities and what their equivalent resistance values are, the power at the reject port could be measured.

Suppose the Isolation between Ports 1 and 1' is 30 dB; this means that there is 1,000 times less power at Port 1' than applied at Port 1.

$$1000 \times \left(\frac{2r}{R + 2r}\right)^2 = 1$$

or:

$$r = \frac{R}{2(10\sqrt{10} - 1)}$$

since R is normally 50 ohms

$$r = \frac{25}{10\sqrt{10} - 1} = 0.81\ \Omega$$

The power dissipated in r is:

$$P_d = \frac{2rR}{(R + 2r)^2}$$

$$P_d\% = \frac{2 \times 0.81 \times 50}{(50 + 1.62)^2} \cong 3\%$$

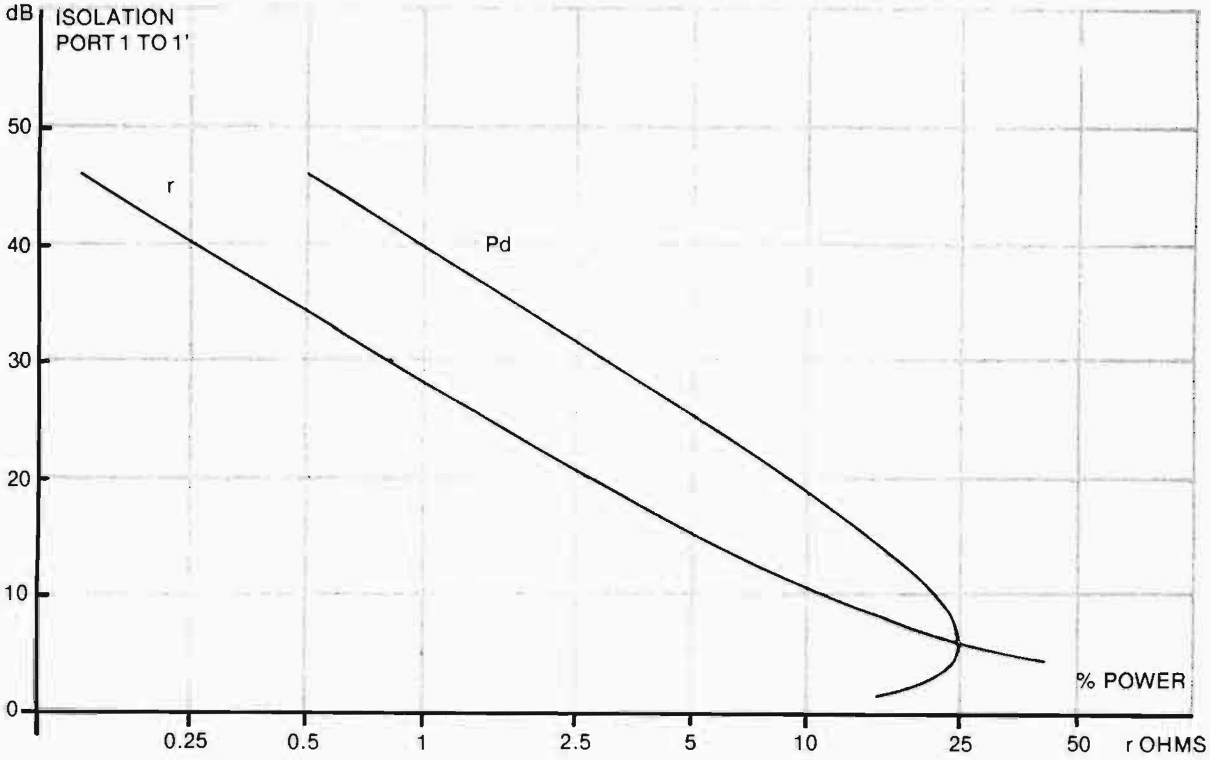


Fig. 61. Isolation of notch diplexer.

So that approximately 3% of the aural power is dissipated in each cavity.

Fig. 61 shows the power dissipated in each cavity versus the isolation of Port 1 to 1'.

The cavities have little effect for the visual carrier frequency and sidebands since they represent a high impedance (capacitive reactance) for these frequency. However, to offset whatever small effect the cavities may have, small inductances of equal, but opposite, reactances as the cavities at visual carrier frequency are installed in parallel with the cavities. To have the best possible circuit, these inductive reactances are so selected that the reactance versus frequency slope is complementary to these of the cavities.

If the ratio of voltages from Port 1 to Port 1' is compared with the notch depth, another interesting conclusion can be arrived at:

$$\frac{V_1}{V_1'} = \frac{R + 2r}{2r} \quad \text{for } \theta = 90^\circ$$

Cavity notch depth:

$$\frac{V_2}{V_2\,(1 - \epsilon)} = \frac{R + 2r}{2r} \quad \text{for } \theta = 90^\circ$$

Therefore, the notch-depth is equal to the ratio of the voltages at Port 1 and Port 1' or to the isolation of Port 1 to 1'.

The average notch diplexer will induce envelope delay for high video frequencies starting at approximately 2.5 MHz and reaching from 300 to 600 nanoseconds at 4.18 MHz, the band edge. The delay must be compensated elsewhere in the system. It can be accomplished at video frequencies, since the disturbance is restricted to the single sideband area of the television signal. For this purpose, a notch diplexer delay equalizer is generally added in, or ahead of, the visual exciter in a television transmission system.

TRANSMITTER COOLING SYSTEMS

It is the intent to provide some of the basic relationships needed to check or calculate the approximate cooling requirements for equipment of the type used in broadcast service. These relationships were simplified to make their use easy and quick. It is not intended that a major system be designed using these approximations.

Cooling by Forced Air

In broadcast equipment some of the power supplied is dissipated as heat. This requires that enough air must be passing through or over the unit to remove the heat, thereby maintaining the desired temperature. Air cooling is accomplished by pounds of air and not cubic feet per minute (CFM). The relationship between these two numbers is density (pounds per cubic foot). It is general practice to specify cooling requirements in CFM of standard air, thereby fixing the pounds per minute of air required. The density of standard air as measured at sea level and 70° F is .075 pounds per cubic foot. See Fig. 62 for correction factors to account for the specific altitude and temperature condition desired.

The basic relationship for calculating air flow requirements is:

$$CFM_{std} = \frac{3160 \text{ X } P}{\Delta T}$$

CFM is cubic feet of standard air.
P is kilowatts of power to be removed as heat.
ΔT is the temperature rise in degrees fahrenheit.
3160 is a constant.

To move this CFM from standard air to the specific condition desired divide CFM standard by the correction factor, Fig. 62.

Sample Problem: An electronic package dissipates 0.1 kw and the desired temperature rise for the air passing through the package is 15°F. The unit must operate at 100° F air inlet and an altitude of 2,000 ft. Find the CFM of air required.

$$CFM_{std} = \frac{(3160)(0.2)}{15} = 42 \ CFM_{std}$$

Correction factor from Fig. 62 for 100° F and 2,000 feet altitude is .880. Dividing the CFM_{std} by the correction factor it is found that 48 CFM is required.

The air flow required is now known, but it is also necessary to know what the resistance to this flow through the unit is. This resistance must be calculated or estimated for each element of the unit, see Fig. 63. Evaluate each element.

Unity Basis = Standard Air Density of .075 lb/ft^3
At sea level (29.92 in. Hg barometric pressure) this is equivalent to dry air at 70°F.

Air Temp. °F	Altitude in Feet Above Sea Level: 0	1000	2000	3000	4000	5000	6000	7000	8000	9000	10000
	Barometric Pressure in Inches of Mercury: 29.92	28.86	27.82	26.82	25.84	24.90	23.98	23.09	22.22	21.39	20.58
70	1.000	.964	.930	.896	.864	.832	.801	.772	.743	.714	.688
100	.946	.912	.880	.848	.818	.787	.758	.730	.703	.676	.651
150	.869	.838	.808	.770	.751	.723	.596	.671	.646	.620	.598
200	.803	.774	.747	.720	.694	.668	.643	.620	.596	.573	.552
250	.747	.720	.694	.669	.645	.622	.598	.576	.555	.533	.514
300	.697	.672	.648	.624	.604	.580	.556	.538	.518	.498	.480
350	.654	.631	.608	.586	.565	.544	.524	.505	.486	.467	.450
400	.616	.594	.573	.552	.532	.513	.493	.476	.458	.440	.424

AIR DENSITY RATIOS

Fig. 62. Air density ratios.

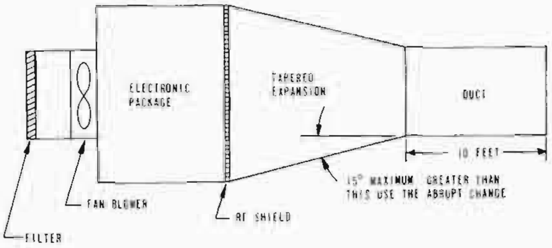


Fig. 63. Cooling system.

1. The filter is necessary to maintain a clean unit. The pressure loss across the filter may be determined by using the manufacturer's pressure loss curve. A properly designed filter will have 0.01 to 0.02 in. of water pressure loss clean and 0.1 to 0.2 in. of water pressure loss dirty.

2. The fan or blower is the driver for this system and must provide sufficient flow (CFM) and pressure (inches of water) to cool the electronics package.

3. The abrupt expansion from the fan/blower into the electronics package will cause a pressure loss. This loss is dependent on the velocity change and ranges from 0.01 to 0.02 in. of water.

4. Electronic package loss may be anything depending on construction and must be estimated or measured. For small open boxes, the pressure loss may be estimated at 0.05 to 0.10 in. of water.

5. Pressure loss across the outlet screen will depend on open area of screen. Screen wire used as RFI shield offers an average loss of 0.01 in. of water at a velocity of 300 feet per minute.

6. Tapered duct induces very low loss and unless air velocity is above 300 feet per minute may be ignored. Above 300 feet per minute allow 0.002 in. of water for each additional 100 feet per minute up to and including 600 feet per minute.

7. Duct loss should be kept to a minimum by using large, short ducts with a minimum of turns and diameter changes. A 90° turn is equal to ten (10) feet of straight duct. See Fig. 64 for approximate straight duct loss.

Pressure loss for the system is the sum of all element losses. The sample problem would appear as follows:

$$\Delta P_{total} = \Delta P_a + \Delta P_c + \Delta P_d + \Delta P_e + \Delta P_f + \Delta P_g$$

ΔP_a	- Dirty Filter	= 0.2	in. water
ΔP_c	Abrupt Expansion	= 0.02	in. water
ΔP_d	Electronic Package est	= 0.10	in. water
ΔP_e	RFI shield 300 FPi	= 0.01	in. water
ΔP_f	Taper 15° 500 FPi out	= 0.004	in. water
ΔP_g	Duct 10 ft. 5 in. @ 500 FPi	= 0.10	in. water
ΔP_t		0.434	in. water

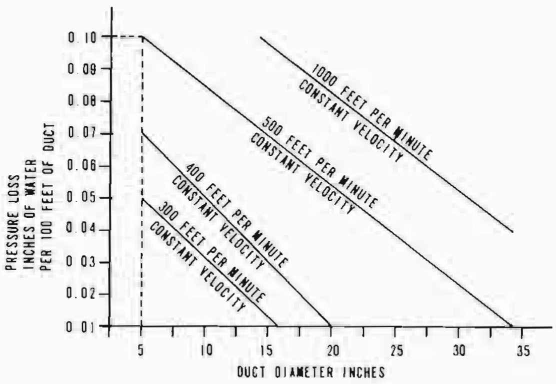


Fig. 64. Duct loss.

Correct this pressure loss for the condition of the sample problem. ΔP_t divided by correction factor equals ΔP required. $\Delta P_r = \frac{.434}{.880} = .493$ in. water. The fan or blower to cool this electronic package must move 48 CFM @ 0.5 in. of water. This is a good application for a muffin fan.

Broadcast equipment may be supplied with the blower or fan for cooling installed in the unit or the equipment may require the installation of the blower or fan external to the broadcast equipment. When the blower or fan is installed in the equipment, an auxiliary blower or fan will be required to provide for air losses in ducting if the hot air is to be ducted out of the building. The calculation for a transmitter exhaust duct system would be the same as the sample problem. To design the fan or blower the following information must be known:

a. CFM_{std} flowing,
b. Temperature of air @ outlet of equipment,
c. Altitude of site,
d. Pressure loss of exhaust duct system,
e. Negative pressure if required in exhaust duct,
f. Pressure loss of input duct system and filters.

Items a, b and e must be obtained from the equipment manufacturer while Items c, d and f must be evaluated for the case in point. More accurate data for duct losses may be obtained from heating and ventilating contractors.

Broadcast equipment using external air systems require a layout typical to the one in Fig. 65.

Air flow measurement outside a laboratory is very difficult if a high degree of accuracy is required. The basic tools required are: U-tube manometer, thermometer and some pressure drop curves for known system elements. Using the U-tube manometer, the static pressure across a known element such as a power tube may be measured. Comparison of this measured value with that same value of pressure loss on the curve gives a flow rate of air. Note here care must be taken to correct for temperature and altitude of incoming air to agree with the curves being used. A second check for air flow is to

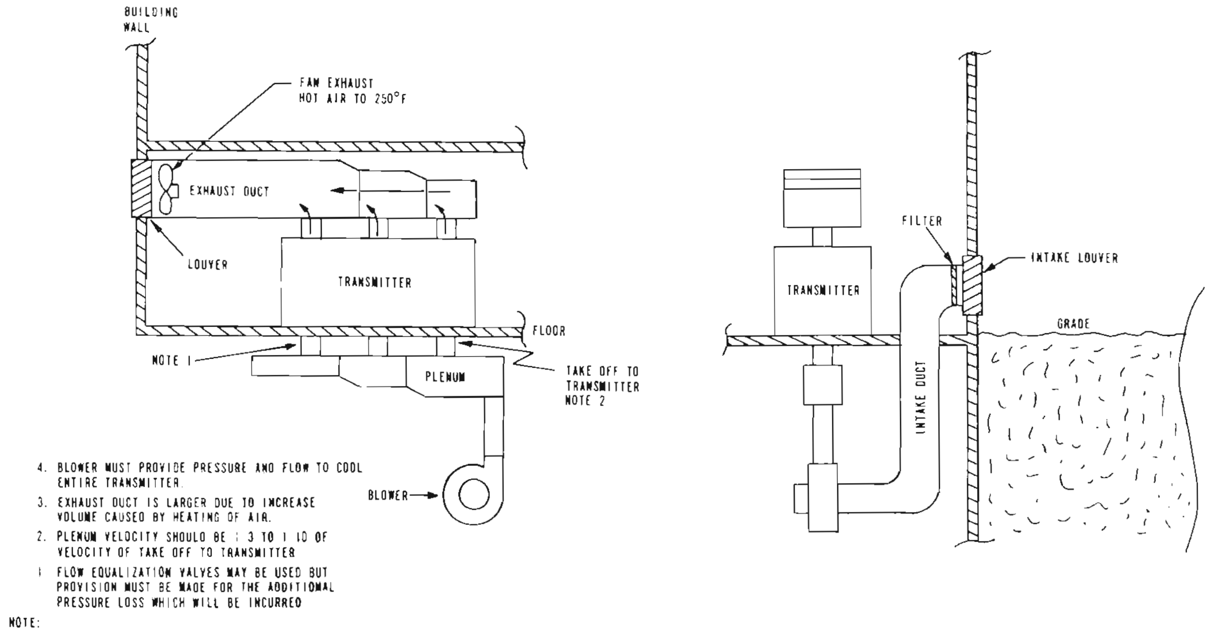


Fig. 65. Typical cooling system.

measure input and outgoing air temperatures across a known element and at the same time measure power dissipated in kilowatts. Knowing ΔT and P (in kilowatts) and using $CFM_{std} = \frac{3160 \times P}{\Delta T}$, the standard CFM can be arrived at. Comparison of this number with the one obtained by the manometer method provides a check. These numbers will not agree exactly, due to the many inaccuracies, but $\pm 15\%$ can be expected.

UHF TRANSMITTERS

Visual Exciter/Modulator

The Exciter/Modulator for UHF transmitters in principle departs little from the concept delineated in the VHF section. Other than the different treatment of the on-channel amplifier caused by the higher operating frequency and its up-converter, only two areas are modified to better serve requirements peculiar to a UHF transmitter employing klystron tubes as power amplifiers: In VHF which uses very linear tubes for power amplification, PC board No. 2 of the video portion provides for moderate correction of differential gain with a small ($\pm 3°$) capability for the differential phase correction.

With the substantially greater need for correction for both differential gain and differential phase in a klystron, this function is now performed by two different circuits. PC board No. 2 in the video portion will only correct for differential phase up to $\pm 10°$.

The need for much larger control range with steeper changes in rate make it desirable, if not mandatory, to place the corrector after the VSB filter in order to provide a wider bandwidth for the predistorted signal. Therefore, the linearity corrector is placed in the IF portion just ahead of the up-converter. With this approach four step correctors will allow a klystron to operate with a sync peak power of 10% below its saturated power.

Fig. 66 shows the schematic of the differential phase corrector. The correction is accomplished by phasor addition of in and out of phase voltages of the original video signal through an inductance (L2) and a resistor (R3), the latter being varied depending on the instantaneous video level. While this phase shift

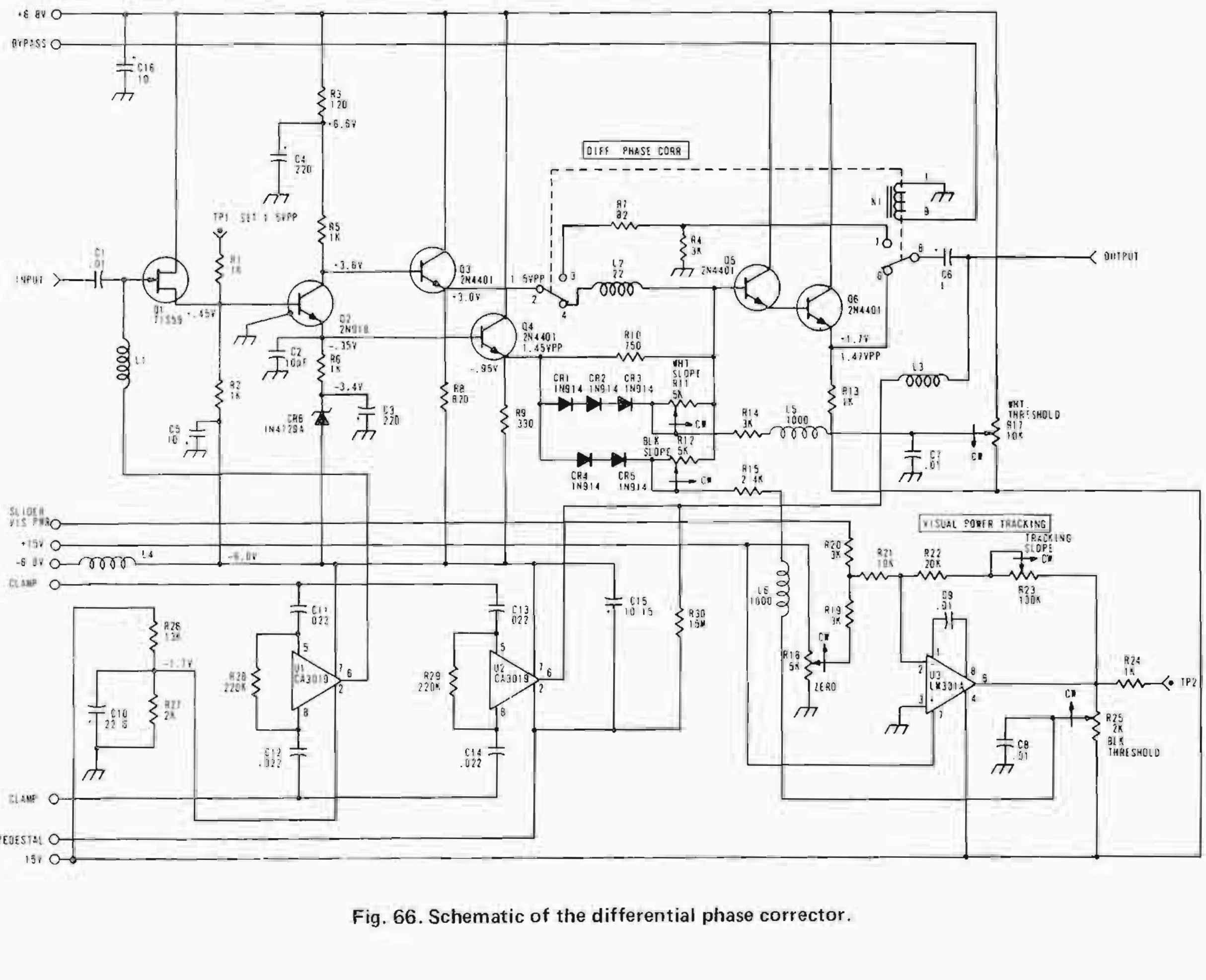


Fig. 66. Schematic of the differential phase corrector.

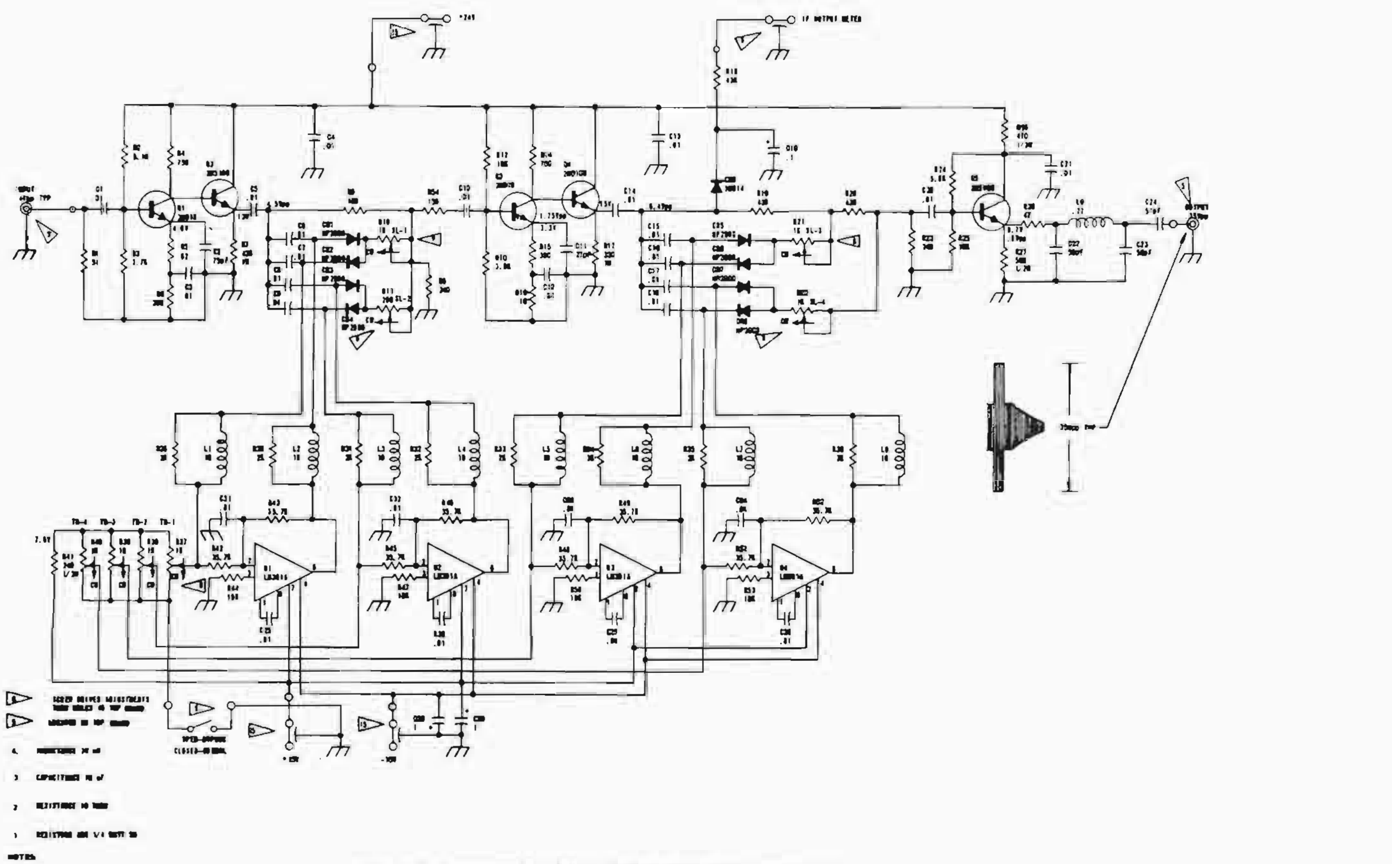


Fig. 67. Schematic of the differential gain corrector.

takes place no change in the video signal level occurs.

The circuit allows for a two step correction with adjustable slope and point of attack which has proved completely satisfactory in a practical system. A bypass switch allows the differential phase corrector to be switched in, or out.

Fig. 67 is a schematic of the differential gain corrector. It provides four control steps with the rate of change and point of attack adjustable for an optimum match. The circuit will work on the positive and the negative cycle of the IF waveform with a single adjustment, a feature that makes this adjustment easier and more straightforward.

Controls TH-4 and SL-4 set point of attack and slope for sync expansion. Controls TA-3 and Sl-3 perform the same function in the black area and TH-2 and SL-2, as well as TH-1 and SL-1 compensate for errors in the gray to white area. The circuit can also be bypassed by switch activation on the module. The operation of this unit can readily be understood from the schematic. Correction is supplied by voltage dependent control of two voltage dividers, R-8 and R-9 for white-gray and R-19, R-20, and R-23 for black and sync areas.

The generator for the pump frequency serves both mixers. The signal, which in the VHF Exciter/Modulator is obtained by multiplication by 2 for low band, or 4 for high band of the crystal oscillator, follows a totally different approach in the UHF Exciter/Modulator. This new approach is designed to accommodate the higher frequency of the carrier as well as the large number of individual channels.

The circuit centers around a step-recovery diode (SRD).[19] The advantage of this diode in this application is twofold: Every integral multiple of the drive frequency appears in the output and virtually no tuning of the device is required.

About 500 mw of drive is required which is obtained from the crystal oscillator through a conventional 60-MHz amplifier with approximately 10-MHz bandwidth.

The availability of every integral multiple of the input frequency allows design of the system with a crystal oscillator frequency variation of only 6 MHz and, therefore, the drive amplifier can be a nontuned broadband amplifier which drives the SRD diode. The tuning of the exciter to the selected channel involves merely the selection of the proper pump oscillator crystal and the tuning of two bandpass filters to the proper multiplier frequency and to the difference frequency of the up-converter. About 20 mw is available from the multiplier over a range from the 9th to the 14th harmonic when

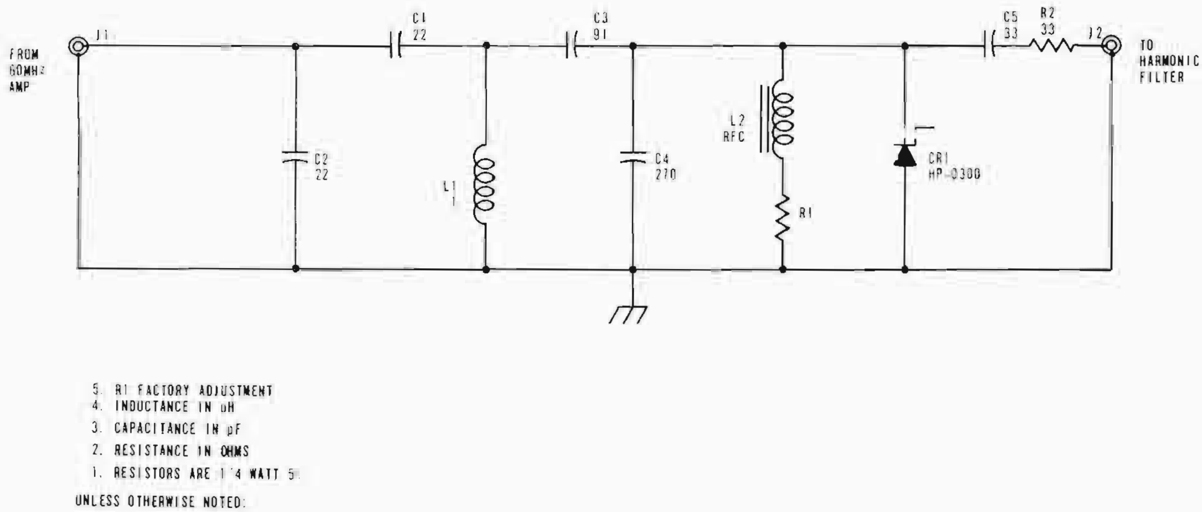


Fig. 68. Harmonic generator, schematic.

driven with approximately 500 mw. Channel 14, for instance, with a carrier frequency of 471.25 MHz, the pump is placed at 471.25 MHz plus 37 MHz equals 508.25 MHz. This is the 9th harmonic of the crystal frequency of 56.472222 MHz. The multiplier of 9 can be employed up to Channel 22, a 519.25 MHz carrier frequency leading to a crystal frequency of 61.8055 MHz. From Channel 23 on, a multiplier of 10 is used, leading to a crystal frequency of 56.225 MHz and so on.

It becomes evident that, if each integer in a multiplier is available, the crystal frequency variation, and consequently, the bandwidth of the drive amplifier is determined by the spacing of the TV channels.

Fig. 68, the schematic of the SRD harmonic generator shows that no variable, or tunable elements are present. The output is connected to a 7-pole bandpass filter which selects the wanted harmonic frequency and sufficiently rejects all other multiples of the drive frequency. This filter is adjusted for each channel selected and is identical to another bandpass which will be selected and adjusted to the difference frequency from the double-balance up-converter. However, the latter filter is tuned 37 MHz lower in frequency than the former.

A 3-dB 90° hybrid following the harmonic selector bandpass will split the signal in equal mounts for the up-converters in the visual exciter and in the aural exciter.

The SRD multiplier is often referred to as a comb generator since the display on the spectrum analyzer of its output spectrum into a resistive load resembles a comb.

Amplification of the output frequency from the up-converter is performed by a 43 dB gain, five-stage solid-state amplifier, similar to the one in the VHF exciter. The physical implementation of the circuit, however, differs greatly from the lumped circuit VHF model. Construction is in the form of a stripline which provides circuit elements in transmission line, or distributed form. A number of capacitors for tuning and bypassing are retained in lumped form. Fig. 69 is a photograph of a typical stripline amplifier. The aural exciter's IF portion,

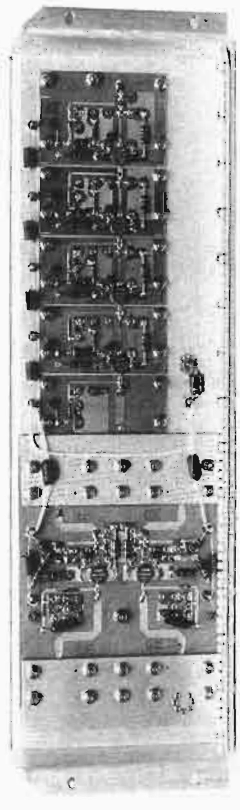


Fig. 69. Photograph of a typical stripline amplifier.

is identical to the VHF model. Only the up-converter and stripline amplifier are different, but are almost identical, except for biasing to comparable elements in the UHF visual exciter.

The visual and aural Exciter/Modulator produce 1 and 3 watts, respectively, sufficient to drive a power amplifier.

Power Amplifiers

Klystrons have found very widespread application in high power UHF TV transmitters. Excellent descriptions of their operation in broadcast service are available from several manufacturers. See References 20, 21, 22, 23, and Fig. 70.

In view of the many references available, only a short characterization of the klystron amplifier will be presented here. For visual service a five-cavity klystron resembles a stagger tuned, five-stage Class A amplifier. Its power gain is just under 10 dB per stage and, therefore, approaches 50 dB for five cascaded stages. Its bandwidth is attained by stagger tuning with the last cavity tuned to the carrier which contains, so to speak, the largest percentage of the total power.

The klystron has an input-output transfer characteristic defined by J_1 of the first order Bessel function which is very close to the first 90° of a sine wave. It will reach a point of maximum power, and if drive is increased beyond this point, the output will decline again.

In addition, when drive is varied from zero to a point reaching the selected output power, the input versus output phase varies greatly and in a nonlinear fashion, with delay rapidly rising when approaching saturated power.

To utilize a klystron for TV service both imperfections must be compensated by signal predistortion which at present is the only practical way to make the entire system linear in amplitude and phase.

While amplitude compensation is an absolute necessity, phase compensation is not required if the rest of the system has phase errors which are very small, and the klystron is not operated up to the saturated power level but only to a level of, 15 to 25% below saturated power. It becomes extremely important to devise methods

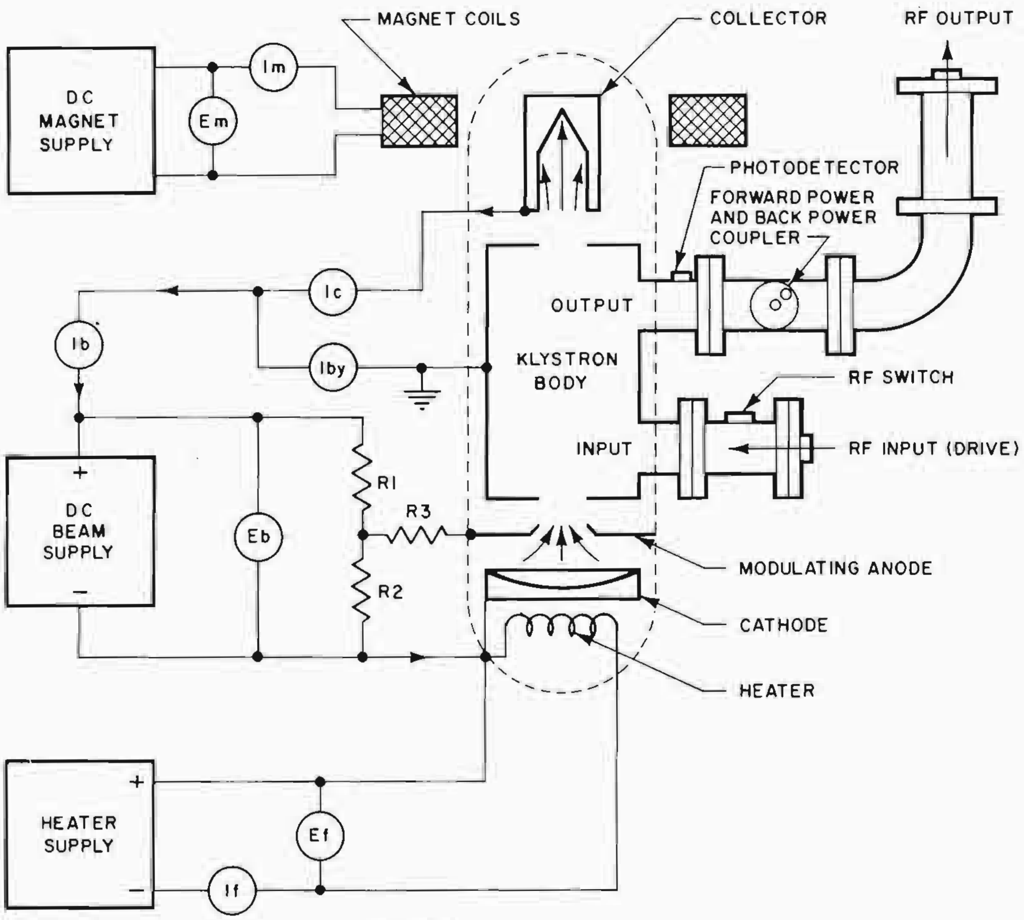


Fig. 70. Klystron amplifier.

of RF power measurements of great accuracy since a small error in power measurement in the wrong direction will force the klystron amplifier to operate closer to saturation, costing the user a loss in performance totally out of proportion with the small increase in power over that for which the station is licensed.

UHF TV stations should, therefore, pay strict attention to the accuracy and reliability of their output power measuring system. This is much less important for VHF transmitters which suffer little deterioration in performance if they are inadvertently (due to measurement inaccuracies) operated 10% to 20% above (true) rated power. From a practical point of view RF power can only be measured to an accuracy of about ±5% under field conditions and often ±10% uncertainty must be expected. Even under laboratory conditions, it becomes very difficult to be more certain than ±3%.

The beam current of a klystron is a function of collector voltage or the modulating anode—body voltage if the klystron is so equipped.

$$I = K \times V^{3/2}$$

Where: I = collector current
V = collector voltage
K = perveance

K has the dimension of $A/V^{3/2}$ power (in the nature of a conductance) and is determined by the physical dimension of the beam generating system of the tube.

By adding a modulating anode in the tube, the collector current can be varied to adapt the tube to different operating conditions, e.g., use as visual or aural amplifier.

Practical values for K lie between 0.1 and $10 \times 10^{-6}\ A/V^{3/2}$.

We can assume that for small changes in beam current the klystron efficiency will remain unchanged. Therefore, for small changes in the supply voltage, we see a 2.5 fold change in output power resulting from:

$$I = K \times V^{3/2}$$
$$P = K \times V^{5/2}$$
$$\Delta P \propto \sqrt[2]{V^5}$$

An increase of 1% of the supply voltage will increase output power by 2.5%. A klystron transmitter's power output (or any other power variation, e.g., hum) is very sensitive to supply voltage variation, and it is often necessary to regulate a supply line which for VHF applications would not require regulation.

Many klystrons are now equipped with a modulating anode which also follows the K X $V^{3/2}$ law (allowing reduced beam current) or in effect, the tube perveance, so that like tubes can be operated in visual and aural amplifiers running at reduced beam current in the aural socket. This way better efficiency for the transmitter can be obtained.

Klystron amplifiers require a considerable amount of supportive elements.

To focus the electron beam and define its path through the drift space to the collector, a magnetic field must be established through the tube. Electro magnets and permanent magnets can be used for this purpose.

Electro magnets are predominantly employed even though there is a finite resistance of the coil and additional dc power is used and must be removed by cooling. Typically, a 55-kw TV transmitter requires 3 kw of dc power for the magnet of each amplifier.

Two supplies provide the magnet current which must be carefully controlled therefore interlock protection is required since loss of the magnetic field (without immediately removing the beam voltage) will destroy the drift tube in a matter of a few seconds. No other source of electric power, except for perhaps 200 watts for filament are required.

Cooling, however, requires an elaborate mechanical system as three different methods are employed. The magnet and klystron body are cooled by running water. Filament seals are cooled by forced air. The collector is cooled by boiling away water (vapor phase cooling). This is done by providing water to the anode or the collector of the tube and allowing the water to boil away. Water removes large amounts of heat with low flow in this mode. The relationship is:

Kw dissipated =

$$.14644\,[(GPM_w)(212 - T_{in}) + 970\,(GPM_s)]$$

Kw dissipated is the amount of power in kilowatts. GPM_w is the gallon per minute of water flowing. T_{in} is the inlet water temperature in degrees Fahrenheit. GPM_s is the gallon per minute of water changed to steam.

Sample Problem: What will the GPM_s be in a system dissipating 182 kw with GPM_w equal 3 and T_{in} equal 120 °F?

$$182 = .14644\,[3(212 - 120) + 970\,(GPM_s)]$$

$$\frac{\dfrac{182}{.14644} - (3)(92)}{970} = GPM_s = .996\,GPM.$$

The boil-off rate is .996 gallon of water per minute.

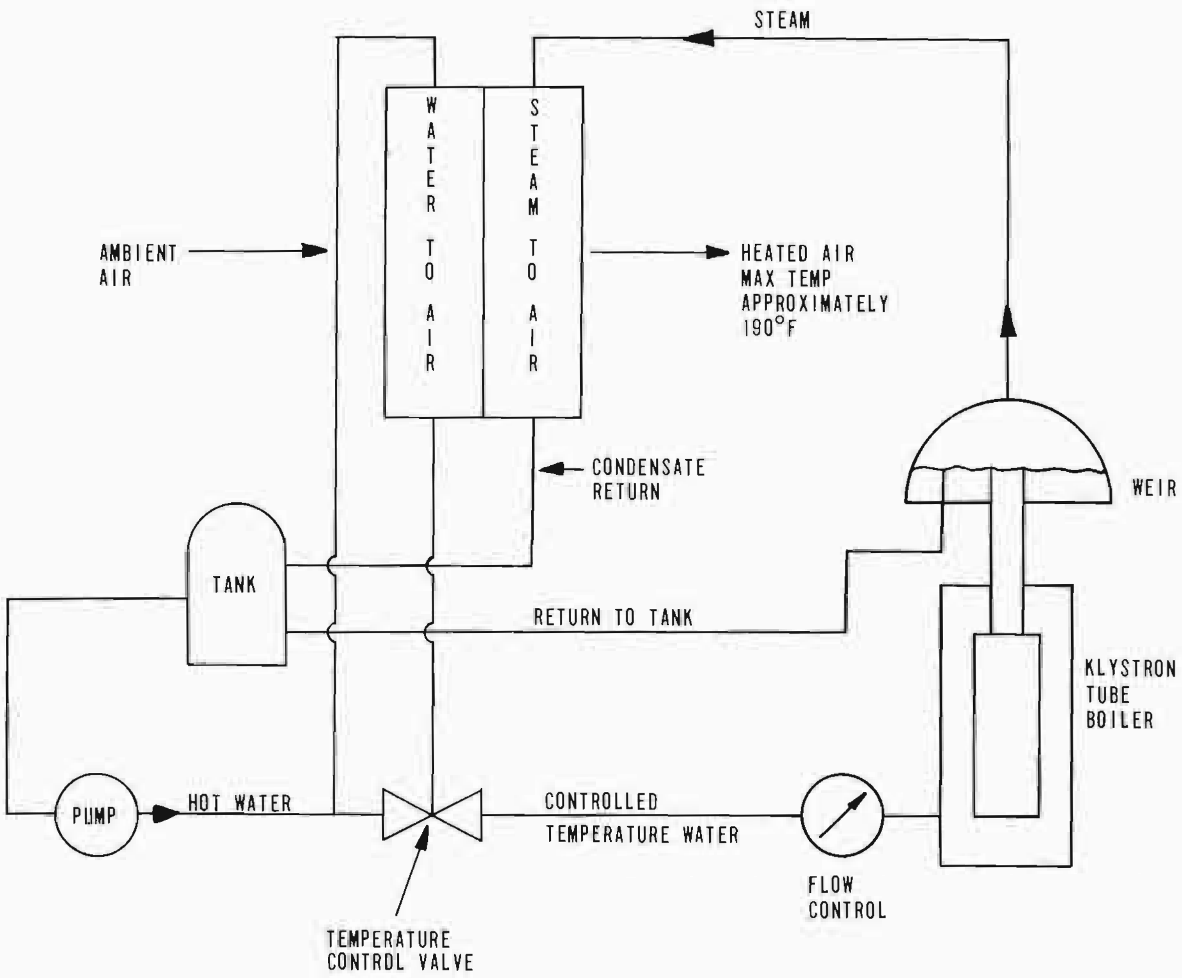


Fig. 71. Vapor phase cooling system.

The sample problem demonstrates how small the water flow required to remove large amounts of power, now a subsystem to supply water to the cooling system must be provided. This system generally consists of a blower, steam to air-condensing coil, water to air-cooling coil, a pump, and a temperature control valve to control the water temperature returning to the transmitter. See Fig. 71.

Major efforts in the design of a klystron amplifier goes into protective circuitry which will protect the expensive tube against damage from any possible malfunction. These protective circuits sense all supply voltages and some currents, they sense the RF load condition, a number of selected temperatures and coolant flow. These circuits, while not correcting the fault, will, however, remove the transmitter from service if certain limits of the sense quantities are exceeded in order to avoid destruction of the klystron. Additional protective devices are required for operator safety.

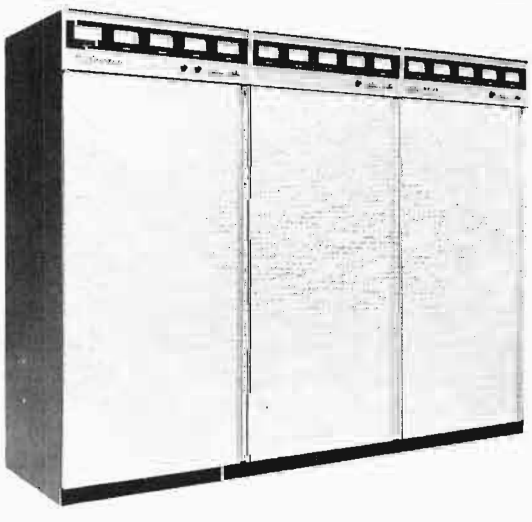


Fig. 72. Harris BT25H1 TV transmitter.

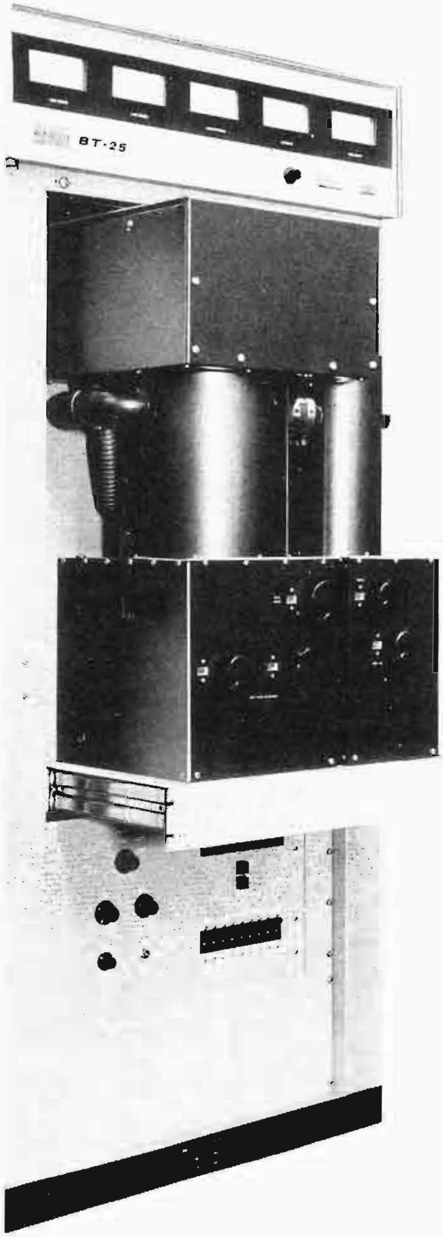


Fig. 73. 25 Kw VHF high band cavity.

TYPICAL TV TRANSMITTERS

Harris BT25H1 VHF TV Transmitter

Model BT-25H1 (25 kilowatts) high band VHF television transmitter features IF Modulation. The BT-25H1 consists of 3 cabinets: a 1.3 kilowatt exciter/driver, an aural power amplifier, and a visual power amplifier—plus an external power supply. The BT25H1 is FCC type accepted, and meets or exceeds CCIR specifications. Visual and aural exciters generate fully modulated low-level IF signals. The output of a common crystal controlled reference oscillator is used to raise the individual IF signals to the desired "on-channel" output frequency.

IF Modulation requires less circuitry to produce a fully processed visual signal. The ring modulator, with excellent linearity, permits modulation percentages to approximately 2% without compromising transmitter performance and eliminates most predistortion circuitry. This results in exceptional color performance and nearly perfect signal linearity. Even such colors as highly saturated yellow and cyan are faithfully reproduced with IF Modulation. Envelope delay compensation for the VSB filter is performed at the IF frequency. Continuously variable controls allow a precise delay correction not possible with conventional fixed-step systems.

Also, visual sideband shaping is performed on the IF frequency at milliwatt power levels, and provides precise shaping of both lower and upper sidebands.

The sideband filter used is a small removable module housed in the visual exciter.

Signal Flow (Visual): The visual exciter signal passes through two IPA stages: a Type 7289 planar triode and a Type 8122 tetrode, giving an output of approximately 100 watts. From the IPA, the signal goes to a power amplifier, an 8792 coaxial tetrode that delivers up to 1,300 watts drive through a circulator to the 8916

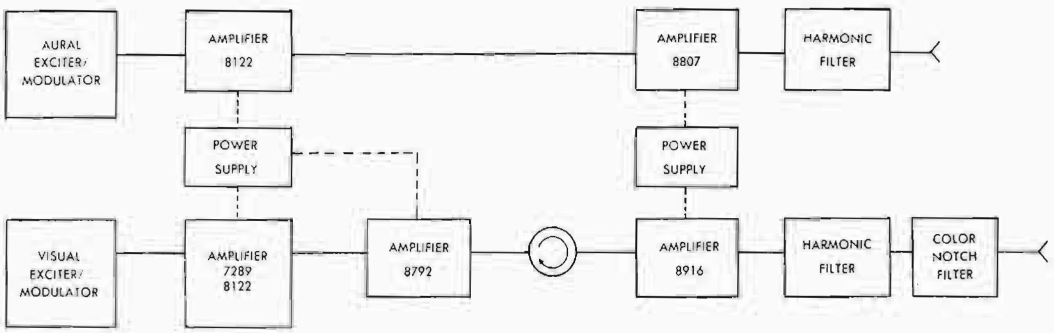


Fig. 74. Harris BT25H1, block diagram.

final visual amplifier. Harmonic filters and color subcarrier filters are externally mounted.

Signal Flow (Aural): The 10-watt output of the aural exciter drives an 8122 IPA. The IPA drives an 8807 aural final amplifier, which delivers up to 5 kilowatts average power output.

Control Logic: Complete and fool-proof control of all transmitter functions is achieved through the use of solid-state memory, timing, and logic circuits. A self-charging emergency power source is provided to maintain control logic memory during periods of power line failure.

The solid-state control logic and protective circuitry, in addition to commanding normal ac control functions, is also used to visually indicate, through pilot lights, the operating status of the transmitter system. The indicator lights allow easy isolation of circuit faults.

All control, metering and monitoring circuits have been designed specifically for remote control operation. The power controls are motor driven and all remote control and sample points are available at one common access point.

Accessibility: Total transmitter component accessibility is provided, front and back. Visual and aural exciters slide out and can operate independently from the transmitter outside the main cabinet. Various exciter circuits, such as oscillators, modulators, and processing circuitry, are of modular construction and can be removed for maintenance or replacement.

Easy-to-read, eye-level, 4-in. meters are used to monitor required transmitter functions. The meter panel is of double-hinged construction for convenient fold-down access during maintenance.

All components of the BT-25H1—including visual and aural exciters, intermediate power amplifiers, final power amplifiers, sideband filters and power supplies—are designed for ease of maintenance.

SPECIFICATIONS

Visual Performance

Power Output: BT-25H1: 25 kilowatts. (Measured at the output of the notch diplexer, if used.)

Output Impedance: 50 ohms. Output connector: (BT-25H1) 3-1/8" EIA standard.

Frequency Range: 174 to 216 MHz (Channels 7 through 13).

Carrier Stability: ±250 Hz (maximum variation over 30 days).

Regulation of RF Output Power: Better than 3%. (Black to White Picture.)

Amplitude Variation of Output: Over one picture frame: less than 2%. (Measured at blanking level.)

Visual Sideband Response:

+4.75 MHz and higher	−20 dB or better
Carrier to +4.18 MHz	+0.5, −1 dB
Carrier	0 dB reference
Carrier to −0.5 MHz	+0.5, −1 dB
−1.25 MHz and lower	−20 dB or better
−3.58 MHz	−42 dB or better

Frequency Response versus Brightness: ±0.75 dB (measured at 65% and 15% of modulation. Reference 100% = peak of sync).

Visual Modulation Capability: 3% or better.

Differential Gain: 0.5 dB or better (maximum variation of subcarrier amplitude from 75 to 10% of modulation. Sub-carrier modulation percentage: 10% peak-to-peak).

Linearity (Low Frequency): 0.5 dB or better.

Differential Phase: ±3° or better (maximum variation of sub-carrier phase with respect to burst for modulation percentage from 75 to 10%. Sub-carrier modulation percentage: 10% peak-to-peak).

Signal-to-Noise Ratio: −50 dB or better (RMS) below sync level.

Envelope Delay:

.05 to 2.1 MHz	±40 ns	Referenced to standard FCC curve
at 3.58 MHz	±30 ns	
at 4.18 MHz	±70 ns	

k Factor (2t): 2%.

Pulse Response (12.5t): 10% or less peak-to-peak disturbance of Baseline referenced to height of 12.5t pulse.

Video Input: Bridging loop through input with 33 dB or better return loss up to 5.5 MHz, 75 ohms impedance.

Harmonic and Spurious Radiation: −80 dB.

Aural Performance

Power Output: BT-25H1: 5 kilowatts. (Measured at the output of the notch diplexer, if used.)

Audio Input: +10 dBm, ±2 dB into 600 ohms.

Input Impedance: 600/150 ohms.

Pre-emphasis: 75 microseconds.

Frequency Response: ±0.5 dB relative to pre-emphasis (30-15,000 Hz).

Distortion: 0.5% or less after 75 microseconds de-emphasis with ±25 kHz deviation, 0.7% after 50 microseconds de-emphasis with ±50 kHz deviation.

FM Signal-to-Noise Ratio: −60 dB relative to ±25 kHz deviation.

AM Signal-to-Noise Ratio: −50 dB relative to 100% modulation (measured after de-emphasis).

Output Impedance: 50 ohms, output connector: 3-1/8" EIA standard.

Frequency Stability: ±250 Hz (maximum over 30 days).

Service Conditions:

Ambient Temperature: −20° to +50° C.
Ambient Humidity Range: 0 to 100% relative humidity.
Altitude: Sea level to 7,500 feet.
Physical and Mechanical Dimensions: BT-25H1: Transmitter size, 94-1/2" wide, 31-1/2" deep, 72" high. Power supply: 36" wide, 24' deep, 60' high.
Weight: BT-25H1: 3,235 lbs. Power supply: 950 lbs.
Electrical Requirements: 208/240 volts (±11 volts) 3 phase, 50/60 Hz. (Special voltages available.)
Power Factor: 100% nominal.
Regulation: 3%.
Phase Unbalance: 2%.

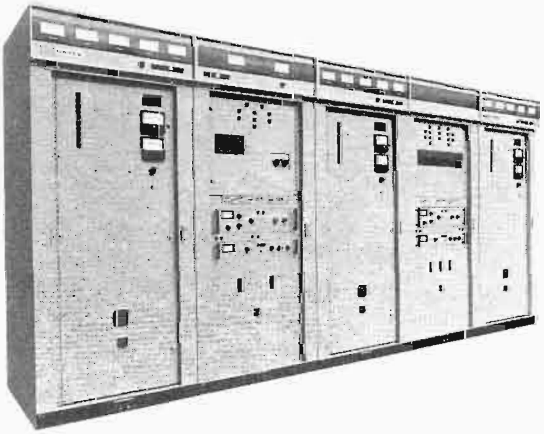


Fig. 75. Harris BT60U TV transmitter.

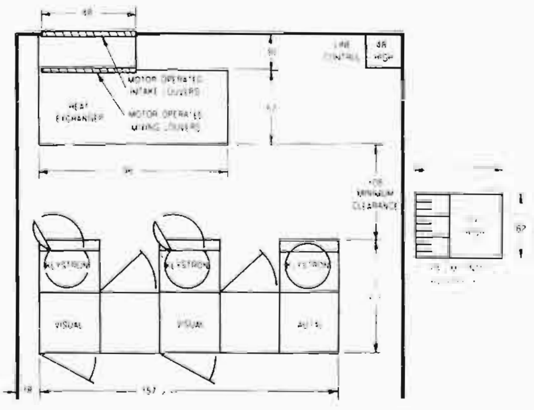


Fig. 76. Floor plan, BT60U.

Power Consumption: BT-25H1.
Average picture: 44 kw.
Black picture: 55 kw.

TYPICAL TV TRANSMITTER

Harris-Model BT-60U

Employing IF Modulation, Harris' FCC type accepted BT-60U provides redundancy of the visual amplifiers for on-the-air protection.

Each of the visual amplifiers operates at a 30-kw power level, and the outputs are added together in a hybrid combiner to produce a total output of 60 kw. If one visual stage should fail, output power will automatically drop to 25% of total output power, with no carrier interruption. The defective stage may be bypassed through an optional patching system so that power may be increased to 50% during service time. In case of an aural amplifier failure an optional kit allows a visual stage to be substituted.

The BT-60U includes two visual klystron cabinets, one aural klystron cabinet, two control cabinets, one high-voltage power supply, and one unitized heat exchanger.

The high power visual and aural amplifiers use identical klystrons. The visual amplifiers operate at 30 kw each, and the aural klystron is capable of producing up to 12 kw at the output of the diplexer. The BT-60U is completely factory tested at full-rated power to assure performance to specifications.

Standby visual and aural exciter/modulators, plus an automatic switcher, are available as options for additional protection.

Low-Level Vestigial Sideband Filtering: The VSB filter consists of lumped tuned circuits and operates at the 37 MHz IF frequency, which affords temperature stability far exceeding any type of on-channel VSB.

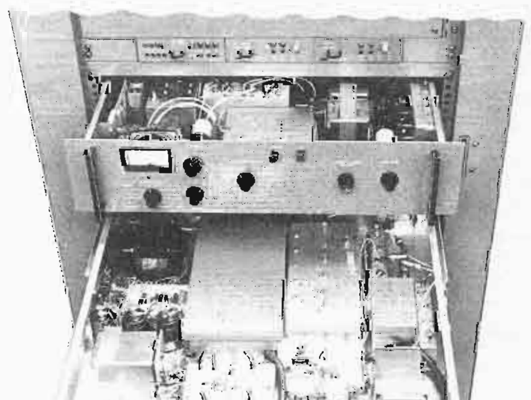


Fig. 77. Harris BT60U transmitter exciter/modulator.

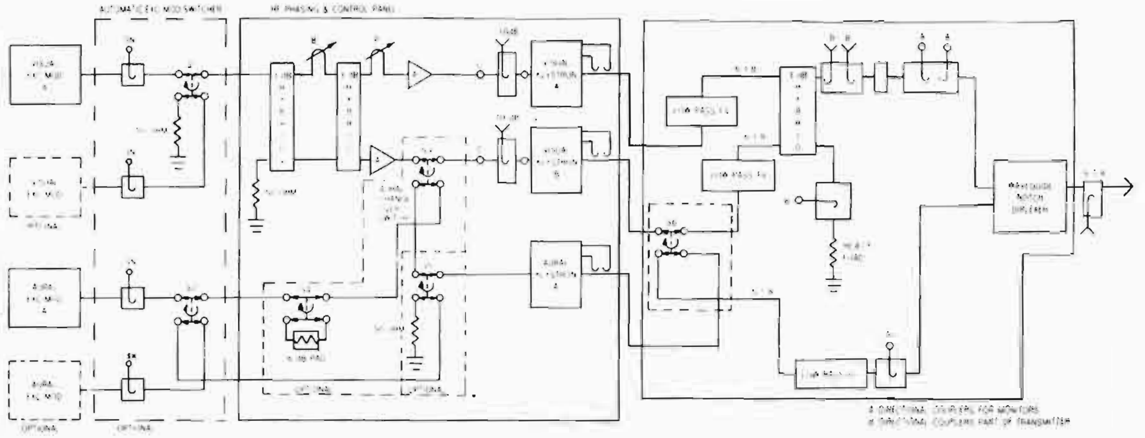


Fig. 78. Harris BT60U transmitter block diagram.

Complete color correction of VSB envelope delay is provided by stable, active filter circuits operating on the IF frequency. This allows asymmetrical correction to be made on either side of carrier to completely compensate for the VSB's nonsymmetrical delay characteristics. Direct on-carrier modulation systems can only provide symmetrical delay compensation and cannot equal the IF color performance.

Klystrons: New high gain klystrons offer substantially higher efficiency over older models. The klystrons contain five internal cavities and amplify the exciter outputs to the proper power levels. Vapor cooling is employed, with one unitized, completely self-contained heat exchanger provided. The klystrons are mounted in special mechanical assemblies which pivot to allow easy installation.

Remote Control: Control circuit functions, metering, and monitoring are designed specifically for remote control operation.

Simplified Assembly, Installation: Harris "building block" concept, in which major components are housed in functional modules, greatly simplifies transmitter assembly and installation, and allows later power increases if desired.

SPECIFICATIONS
(CCIR Specifications Available)

Visual Performance

Power Output: 60 kw, peak of sync (FCC and CCIR systems M, N).

Output Impedance: 50 ohms. Output connector: 6-1/8" EIA Standard.

Frequency Range: 470-890 MHz (Channels 14 to 83; CCIR bands IV/V).

Carrier Stability: ±250 Hz (maximum variation over 30 days).

Regulation of RF Output Power: (Black to white picture): 3% or less.

Variation of Output: Over one frame: less than 2%.

Visual Sideband Response: (exclusive of diplexer)
+4.75 MHz and higher .... −20 dB or better
Carrier to +4.18 MHz........ +0.5, −1 dB
Carrier...................... 0 dB reference
Carrier to −0.5 MHz......... +0.5, −1 dB
−1.25 MHz and lower .... −20 dB or better
−3.58 MHz.............. −42 dB or better
Corner frequencies scaled to meet CCIR standards.

Frequency Response versus Brightness: ±0.75 dB (measured at 65% and 15% of modulation. Reference 100% = peak of sync).

Visual Modulation Capability: 3% or less.

Differential Gain: 0.5 dB or better (maximum variation of sub-carrier amplitude from 75% to 10% of modulation. Sub-carrier modulation percentage: 10% peak-to-peak).

Linearity (Low Frequency): 0.5 dB or better.

Differential Phase: ±4° or better (maximum variation of sub-carrier phase with respect to burst for modulation percentage from 75% to 10%. Sub-carrier modulation percentage: 10% peak-to-peak).

Signal-to-Noise Ratio: −50 dB or better (RMS) below sync level.

K Factors: 2t 2%, 12.5t less than 10% baseline disturbance, 20t 3%.

Envelope Delay:
.05 to 2.1 MHz ... ±40 ns
at 3.58 MHz...... ±30 ns
at 4.18 MHz...... ±70 ns
Referenced to standard FCC curve

Video Input: Bridging, loop through input with −30 dB or better return loss up to 5.5 MHz, 75 ohm system.

Video Input Level: 1.0 Volt peak to peak ±3 dB, sync negative.
Harmonic Radiation: −80 dB.

Aural Performance

Power Output: 6 kw to 12 kw, at diplexer output.
Output Impedance: 50 ohms. Output connector: 6-1/8" EIA standard.
Audio Input: +10 dBm, ±2 dB into 600 ohms.
Input Impedance: 600/150 ohms.
Pre-emphasis: 75 microseconds.
Frequency Response: ±0.5 dB relative to pre-emphasis (30-15,000 Hz).
Distortion: 0.5% or less after 75 microseconds de-emphasis with ±25 kHz deviation.
FM Noise: −59 dB relative to ±25 kHz deviation.
AM Noise: −52 dB relative to 100% modulation (measured after de-emphasis).
Frequency Stability: ±250 Hz (maximum variation over 30 days).

Service Conditions

Ambient Temperature: +2° to +50°C. (36° to 122°F).
Ambient Humidity Range: 0 to 95% relative humidity.
Altitude: Sea level to 7500 feet.
Physical and Mechanical Dimensions: Transmitter: 157-1/2" W X 63" D X 72" H.
Weight: 6550 lbs. Power Supply: 72" W X 62" D X 59" H. Weight, 9200 lbs. Heat Exchanger: 96" W X 52" D X 80" H. Weight: 4000 lbs.
Electrical Requirements: Power input: 440/460/480 Volts, 3 phase, 60 Hz, 4 wire (380 Volts 50 Hz available on special order). Power consumption: 20% aural, 290 kw. 10% aural, 263 kw. Regulation 3%. Phase unbalance: 2%. Power Factor: 100%.

Klystrons

Types:
Low Band (Channels 14-29, 470-566 MHz), VA 946H.
Mid Band (Channels 30-51, 566-698 MHz), VA 947H.
High Band (Channels 52-83, 698-890 MHz), VA 948H.

PRECISE 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 (cochannel interference) geographical separation and radiated powers are carefully selected.

Nevertheless, considerable cochannel interference was encountered in many locations and investigation showed that additional means were available to reduce cochannel interference by making use of optical or perceptive elements in the system. In other words, steps were taken not to alter the strength of the interferring 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.[24,25,26,27,28]

It turned out, with horizontal and vertical sweep ratios fixed, the Visual carrier frequencies of the interferring stations was the parameter which had to be controlled.

Cochannel 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 cyclicly as a function of the difference in frequency of the interfering carriers (Fig. 79). The interference is most visible when the carriers were offset by multiples of the line frequency (15,734 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. The 10 kHz offsets were chosen to provide approxi-

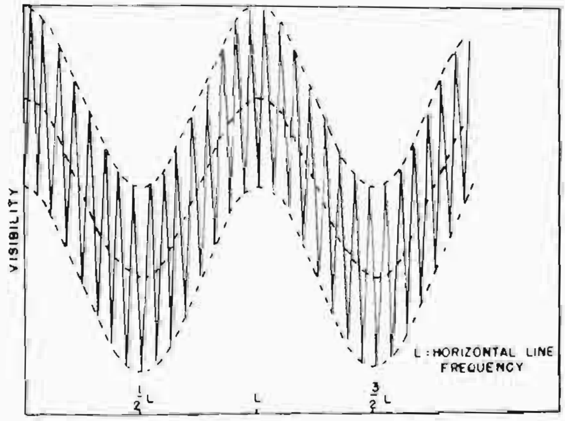


Fig. 79. Cochannel interference.

mately equal reduction of the interference patterns for any number of stations in geographical proximity, see Figs. 80 and 81.

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 10,010 Hz. Since the carrier frequency tolerance is ±1,000 Hz, the precision offset is compatible with existing regulations. In a three station arrangement, the stations will be offset by 10,010 Hz and 20,020 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 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 IF Modulation transmitters, the modulated carrier is obtained by up-converting the modulated IF signal. Gates transmitters employ two crystal-controlled oscillators—one to generate the IF and one to generate the Master Oscillator frequency. After multiplication, the Master Oscillator signal is mixed with the modulated IF signal. The difference frequency is then the visual carrier frequency. The multiplication factors are X2 for VHF Low Band and X4 for VHF High Band.

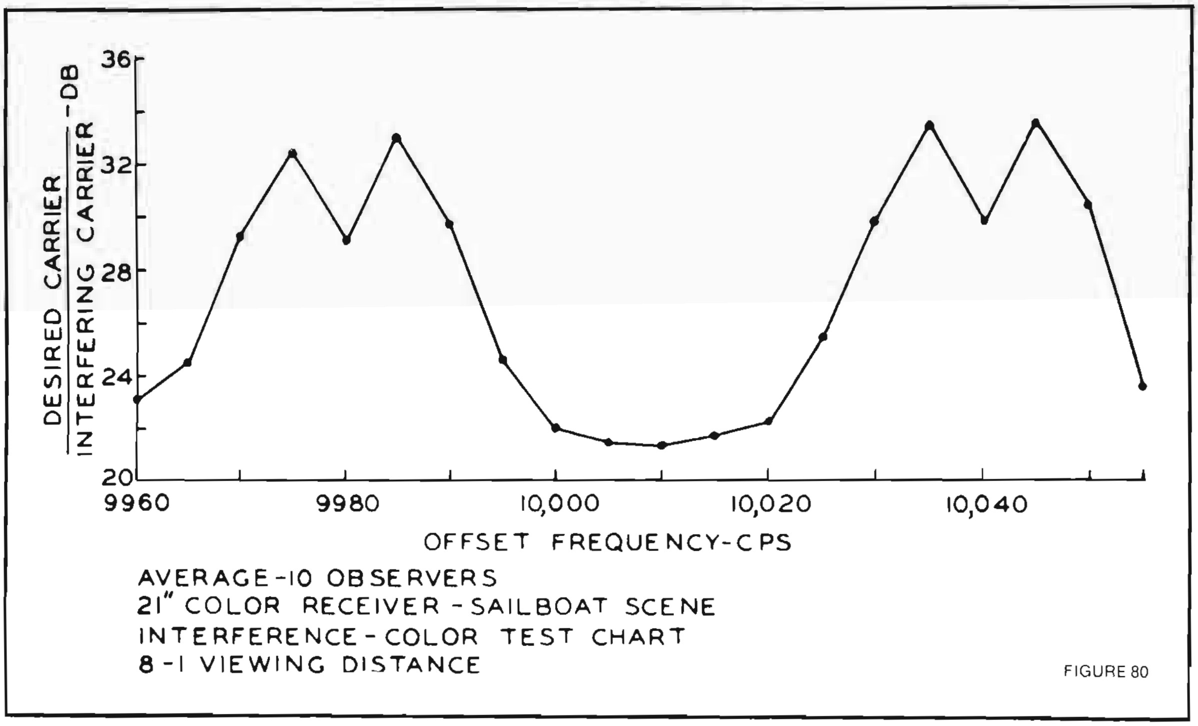


Fig. 80. 10 kHz offset pattern.

$$f_{VCAR} = J X f_{MO} - f_{IF} \quad \begin{matrix} J = 2 \text{ VHF Low Band} \\ J = 4 \text{ VHF High Band} \end{matrix}$$

Since the carrier frequency is derived from two independent oscillators, precise control of the carrier frequency must allow for frequency drift in both oscillators. One possible way to achieve control is to phase-lock each oscillator to an extremely stable oscillator. Another method, which requires less circuitry, is to sample the carrier frequency and correct only one of the oscillators to maintain the correct frequency, see Fig. 82.

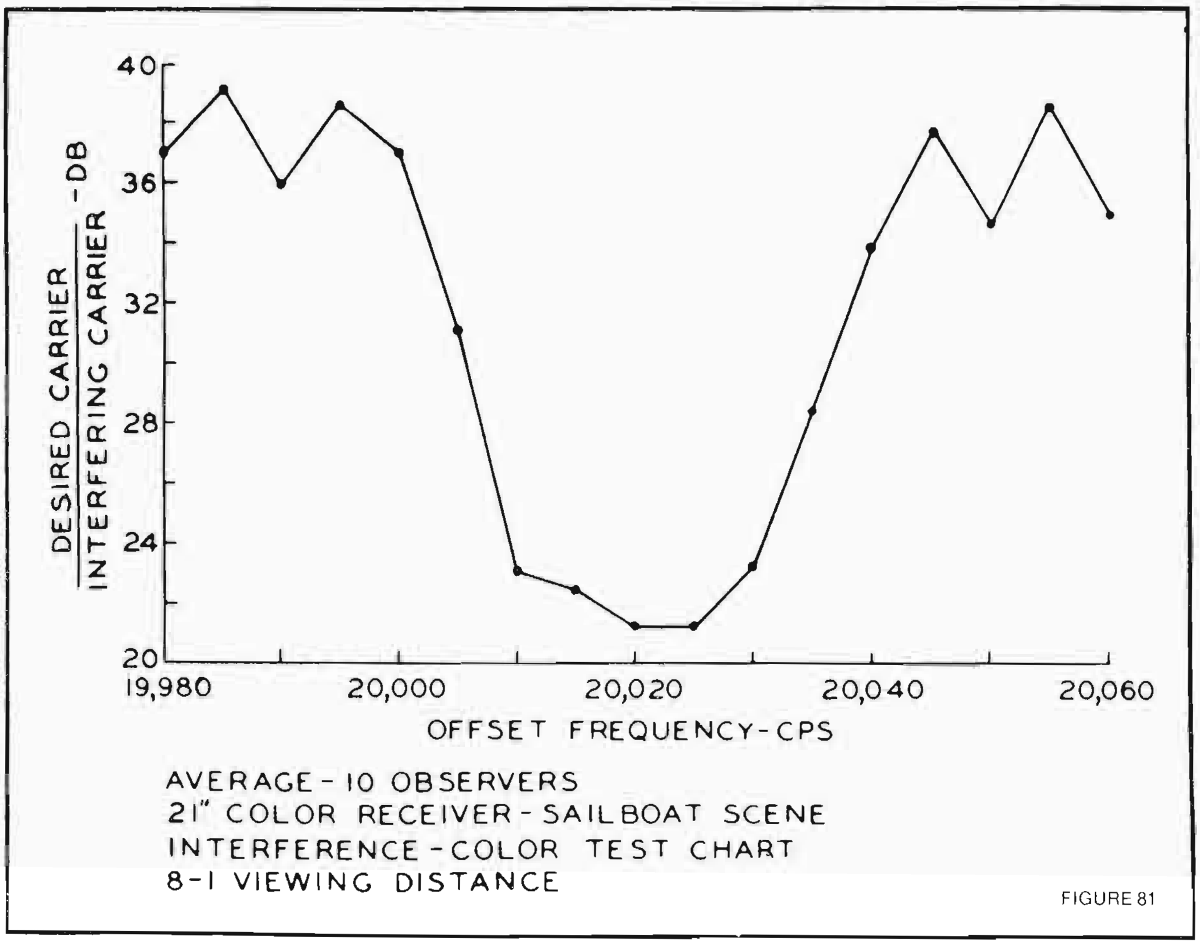


Fig. 81. 20 kHz offset pattern.

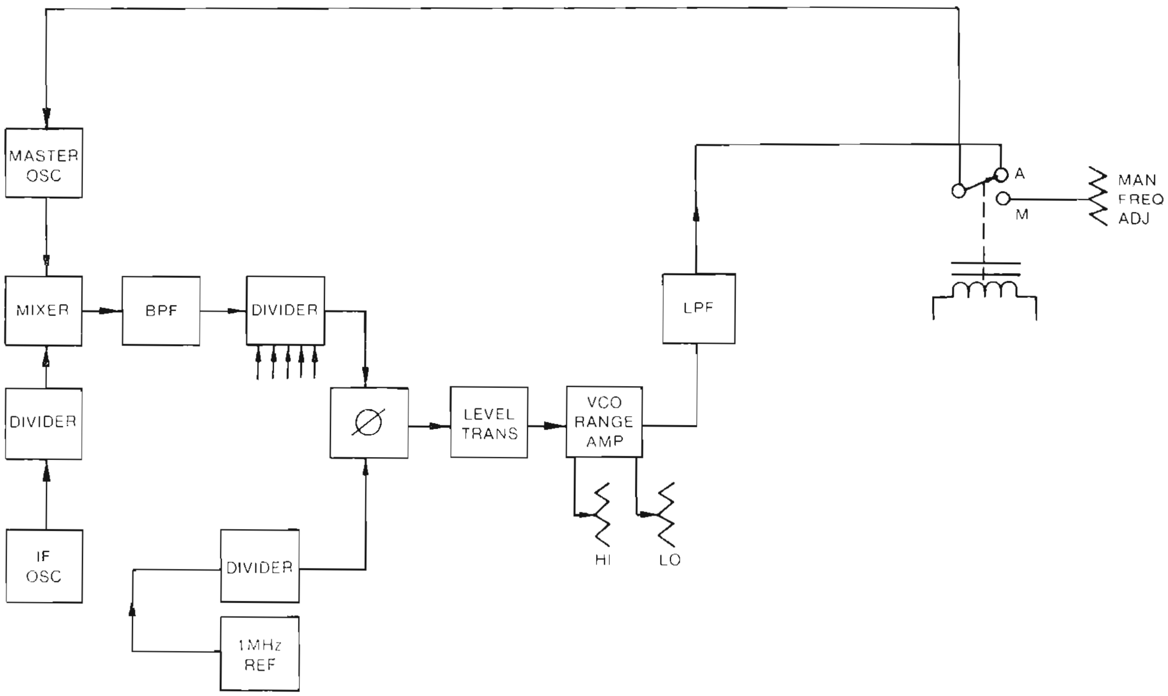


Fig. 82. Precision frequency control block diagram.

The Precise Frequency Control uses the latter method. The Master Oscillator is used as the variable frequency oscillator (VFO) in a phase-locked loop. Although the oscillator is crystal-controlled, its frequency can be pulled slightly by a varactor diode. A sample of the carrier frequency is compared with the signal from an extremely stable oscillator in a frequency/phase comparator. The output of the frequency/phase comparator is the error signal applied to the varactor diode in the Master Oscillator.

To obtain an unmodulated sample of the visual carrier, both the IF Oscillator and Master Oscillators are sampled. For FCC Channels 2-13 the Master Oscillator frequency ranges from 46 MHz to 62 MHz. The IF Oscillator is fixed at 37 MHz. Using digital devices the IF sample is divided by exactly 2 for VHF Low Band and by 4 for VHF High Band. After division, this subharmonic of the IF is mixed with the Master Oscillator sample. A bandpass filter at the mixer output port selects the signal whose frequency is the difference between the Master Oscillator and the subharmonic of the IF Oscillator. The difference frequency is then an exact subharmonic of the visual carrier frequency.

$$f_{\mathrm{d}} = f_{\mathrm{mo}} - \frac{f_{\mathrm{IF}}}{J} = \frac{f_{\mathrm{vcar}}}{J} \quad \mathrm{J} = 2,4$$

The carrier sample is frequency divided by a six-stage decade counter. The decade counters are programmable so that any integral division ratio up to 10^6 can be obtained. The programming can be easily changed to provide the correct division ratio for each channel of operation. This way precision control equipment can be manufactured completely before its channel application is known, leading to a considerably more economic method of production. Furthermore, it becomes possible to use a readily available 1 MHz precision reference for practically all possible VHF and UHF channels without the need to procure precision oscillators for uncommon frequencies. Similarly the output of an extremely stable reference oscillator is digitally divided and compared with the carrier sample in a phase frequency comparator. After filtering, the dc error voltage is applied to the varactor diode in the Master Oscillator. To prevent excessive frequency pulling before lockup, the error voltage is limited to a preset range.

Sensors are employed to disable the loop in the event of any malfunctions. When the loop is broken, the Master Oscillator returns to its normal crystal-controlled operation. Although precision control may be lost because of some malfunction, the transmitter carrier frequency will still remain under crystal control and within FCC limits.

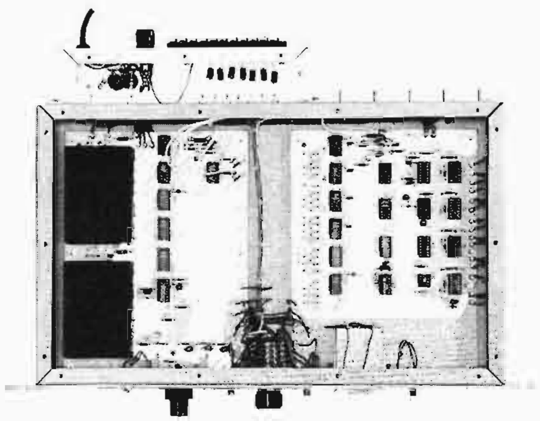


Fig. 83. Precision frequency control, top view.

By phase-locking the Master Oscillator to a stable reference oscillator, the Master Oscillator acquires the stability of the reference source. Crystal-controlled oscillators today can achieve stabilities of less than a few parts in 10 billion (10^{10}) per day. Stabilities of this magnitude mean that a transmitter can be maintained within a few Hertz of a desired frequency for several months. Experimental results have indicated that frequency differences between transmitters can vary as much as ± 5 Hz from the precision offset before cochannel interference becomes noticeably worse. This means that two television stations can minimize their cochannel interference without the requirement to frequently adjust the transmitter frequency. When adjustments are required for the system described in this paper, the transmitter frequency can be changed by adjusting the frequency vernier controls on the reference oscillator. See Figs. 83-84.

Determining when transmitter adjustments are necessary is not an easy task. Measuring transmitter frequencies as high as 200 MHz with 1 or 2 Hz precision is very difficult.

To ascertain the carrier frequency of a station operating under precision control, assuming a

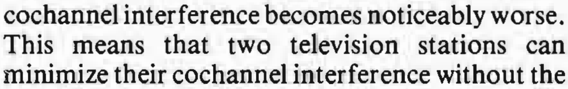


Fig. 84. Precision frequency control, front view.

measuring accuracy 10 times better than the quantity to be measured, leads to the following accuracy requirements:

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

No frequency counter presently on the market will fulfill these requirements directly.

Standard frequency transmissions from station WWVB, located at Fort Collins, Colorado, can be used as primary reference. Transmissions from WWVB run 24 hours a day without interruption at 60 kHz with 16 kw radiated power, vertically polarized and can be received in the entire continental U.S.A.[28]

The transmitted signal is modulated in amplitude by a digital code transmitting at 16 kw for the high logic state and 1.6 kw during the low logic state. This time code is irrelevant for our purpose here.

We are only interested in the carrier itself. This carrier is advanced in phase by 45° at 10 minutes after the hour and returned to 0° phase at 15 minutes after the hour which serves as identification for the station.

The accuracy of WWVB's carrier is a few parts of 10^{-12}.

The signal from WWVB is available 24 hours in the entire U.S.

However, measurements with high accuracy should only be made during periods when the entire path from Fort Collins to the measuring site is in daylight. Figs. 85 and 86 illustrate the effects of transmission irregularities during nondaylight hours. They show the rate of change in phase of a signal received from WWVB versus a local precision reference. It can be assumed that the local

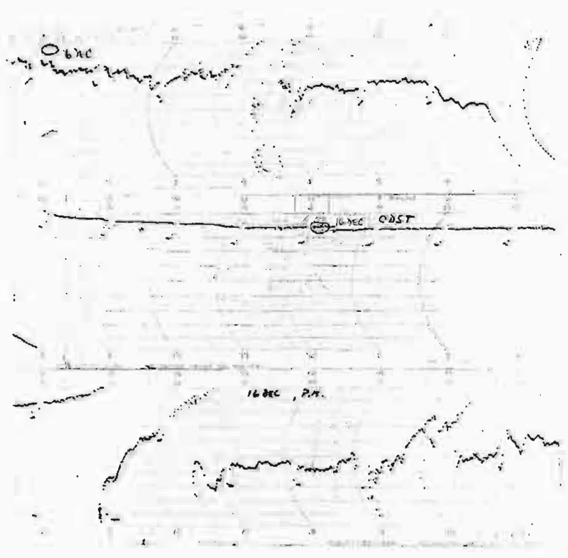


Fig. 85. WWBV phase recording 16-DEC-1974.

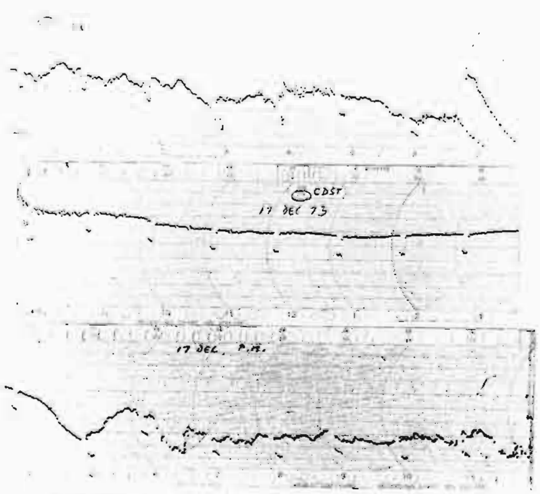


Fig. 86. WWVB phase recording 17-DEC-1974.

source did operate in a consistent manner, day or night, and that a phase departure during nighttime conditions is a result of path disturbance from the WWVB carrier. These disturbances are not always present, but best results are obtained during daylight hours. Note the phase advance from 10 to 15 minutes after the hour which serves as identification of the station.

Two approaches to measure the carrier frequency of a television station operating under precision control will now be described. The first method requires that modulation be removed from the visual carrier and that the aural transmitter be disabled sometime during daylight hours. If this is undesirable, a second method is shown which avoids the above service interruption (leaving, however, a small element of uncertainty).

The first method requires an HP-117A VLF comparator with modification HO1-117A and a HP5245L/5253B Frequency Counter. With the VLF comparator installed and operating (following the instructions provided with this unit), the 1 MHz output from the HP-117A, which is now phase-locked against WWVB, is utilized as an external time base reference for the HP5245/5253 Frequency Counter. With the counter in the frequency measuring mode and using a 10-sec. sampling period, the carrier frequency of the TV transmitter can now be measured, see Fig. 87.

With the aural transmitter turned off and the visual modulation removed, the transmission line sample of the visual transmitter should be fed to the input of the counter. At least ten "10-sec. periods" should be measured, sampled, and recorded for comparison. One should not see more variation in frequency (now displayed to 0.1 Hz resolution) than approximately 1 Hz, plus the usual 1 county ambiguity of the frequency counter. If the frequency indicated is not within ± 2.5 Hz,

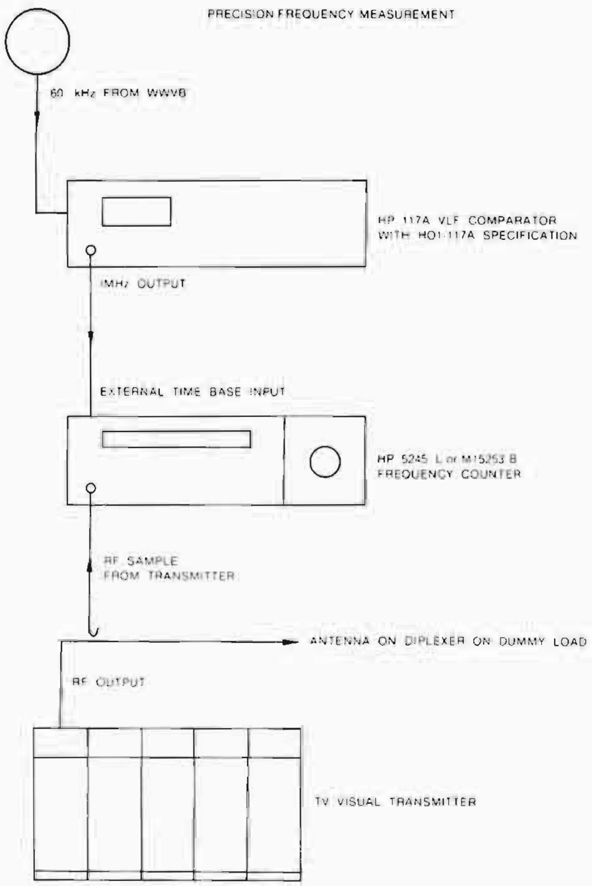


Fig. 87. Measuring system, precise frequency.

the reference standard of the transmitter's precision control equipment should be adjusted. One should keep in mind that any adjustment of the reference standard of the precision control equipment sees a considerable time lag in the phase-lock loop system such that adjustment in the reference standard should be made only in very small frequency increments. Most precision reference standards have calibrated incremental adjustment capability. After an adjustment sufficient time must be allowed for the precision control Equipment of the transmitter to place the transmitter in exact phase-lock to the reference source which may take up to 30 min. Therefore, if adjustment of the carrier frequency of the transmitter is required several measurements should be made at 30-min. intervals to bring the transmitter frequency to within 2.5 Hz of the assigned frequency.

A second method can be used when a service interruption during daylight hours cannot be tolerated. In this case the HP5245M/5253B frequency counter is recommended which has a time base with a lower aging rate.

Preparatory to the frequency measurement of the TV transmitter initial measurements of the carrier frequency (during nonservice hours) should be made using the counter's internal standard to establish proper frequency control, lockup and consistent frequency indication. Next, during daylight hours the exact frequency and drift-rate of the HP counter should be established against WWVB, again utilizing the VHF comparator by recording the phase difference between the 1 MHz phase-locked reference of the VLF comparator against the 1 MHz external output of the frequency counter (which is 1/5 of the internal 5 MHz standard. A recording over several hours should clearly establish the exact frequency of the reference standard as well as it's drift rate. Once the error is established the frequency measurement of the previous night can be recomputed or the internal reference standard of the counter can be adjusted to be within one times 10^{-9} of the wanted frequency and then measurements can be repeated the next night. During the 12-hour waiting period from the daylight measurement of the reference standard to the nighttime measurement of the transmitter's carrier frequency, the precision reference of the HP counter may drift two or three parts of 10^{-10} which will leave the system accuracy well within the limits stated above.

It becomes obvious that the above approaches can be varied in several ways. One should, however, always remember to let precision crystal oscillators operate without any interruption. Any interruption, no matter how short, will (within our frame of reference, speaking in units of 10^{-10}) radically change frequency of any crystal oscillator.

It has been shown that the considerable reduction of areas of cochannel interference between two or three stations operating on the same assigned channel can be utilized in modern television transmitters employing precision control equipment. This advantage can be utilized and maintained without undue efforts and expenditure of funds by employing a low frequency comparator and a standard frequency counter which will allow the user to trace the accuracy of the frequency of his visual carrier to the primary standard of the United States at the National Bureau of Standards in Boulder, Colorado.

AUTOMATIC POWER CONTROL

The output power of visual and aural TV Transmitters may change, resulting from variations of the supply voltage, changes in ambient temperature and for other reasons. Operators are used to adjust the transmitters' output power from time to time to maintain it constant within certain limits, but definitely within the +10, −20% limit established in the FCC Rules and Regulations.

It seems logical to relieve the operator of this tedious task by maintaining output power of both transmitters automatically. In a transmitting system which produces fully processed RF signals at

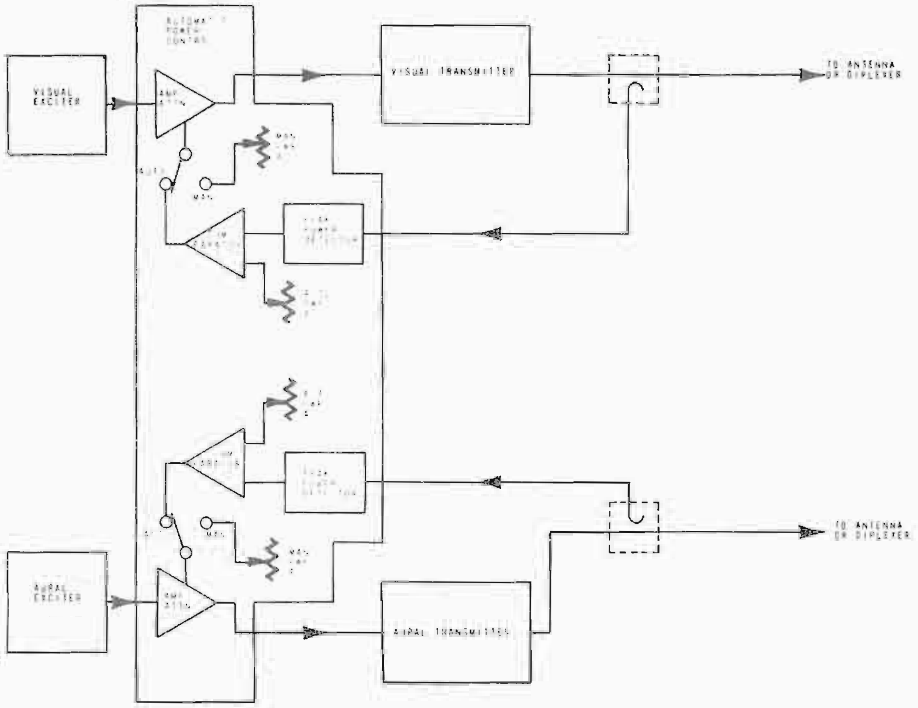


Fig. 88. Block diagram of an automatic power control.

low power levels, a straightforward method offers itself and has since found application as a standard product. Fundamentally, a controllable attenuator is inserted in the transmitter chain at approximately 1-watt power level which will be so adjusted to make the output power constant. The Block Diagram of such a unit is shown in Fig. 88. Three distinct elements are found: an output sensing device, a comparator producing a control signal, and a pin diode attenuator.

The sensing device in the case of the visual transmitter samples the output power during the peak-of-sync period and produces a dc voltage analogous to the power output. In similar fashion, the aural transmitter output is sampled and a dc voltage proportional to it's average output power is produced. In the comparator the dc analog voltages in each case are compared with a stable (but settable) dc reference. The dc output voltage from the comparator, which is proportional to the difference between the sampled dc and the reference voltage, will now steer an electrically controllable attenuator changing system gain to adjust the output power to the desired 100% output.

The control loop can be broken if it is desired to control output power of the transmitter manually.

Automatic power control units employing the above principles can maintain the output power of a visual or aural transmitter to within $\pm 2\%$ of a wanted level with system variations (open loop) of up to $\pm 20\%$. Over this control range no noticeable change in the control signal can be detected, including maintenance of all modulation levels in the visual signal. In part, this is due to a unique type of hybrid attenuator employing dual diodes. This type of attenuator maintains input return loss to narrow limits over its entire control range. Fig. 89 shows a simplified schematic of this type of current controlled attenuator. The automatic power control unit presently manufactured by Harris Corporation contains a pair of control units in a single rack-mounted chassis and is so arranged that it controls either the visual and the aural transmitter of a single television transmitter, or both visual transmitters when they are operated in parallel.

Fig. 89. Schematic of a pin diode attenuator.

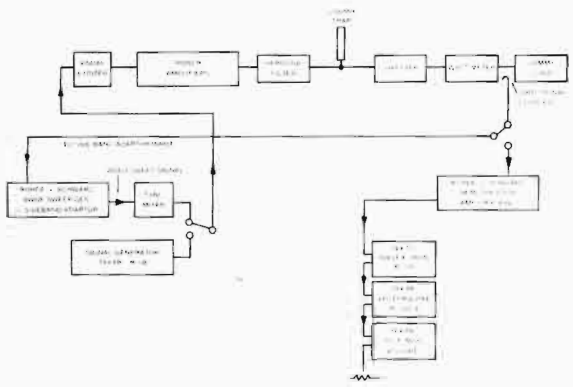


Fig. 90. Test set-up, performance measurements.

PERFORMANCE MEASUREMENTS

Measurements of system performance are generally required after completion of the construction of a new station. When filing a license application with the FCC these measurement data are required in addition to a list of statistical and numerical information related to the transmitter, the station and the antenna system (Proof of Performance). The visual transmitter and its test equipment is arranged as shown in Fig. 90. The following measurements were made on a 220-kw UHF TV transmitter, the first transmitter of this power level ever to be placed in service. Frequency response measured with a double-sideband diode is shown in Fig. 91, while the same response plotted from the scope display of a Rohde & Schwarz SWOF Sideband Analyzer is shown in Fig. 92. In both cases the RF was sampled after the diplexer.

Further measurements were made to show compliance with applicable specifications related to differential gain, differential phase, low frequency response and signal-to-noise ratio. Fig. 93 shows differential phase and gain, Fig. 94 characterizes the transmitter's behavior at low frequencies by showing the vertical interval which occurs at a field rate of 60 Hertz. Visual signal-to-noise ratio through a flat system was – 57 dB RMS (referenced to the peak of sync Voltage).

Performance of the aural transmitter is shown in Fig. 95 which indicates frequency response and Fig. 96 showing harmonic distortion. FM noise of the system is – 64 dB and AM noise is – 58 dB.[29]

The above performance data are generally sufficient for FCC filings in connection with a license application.

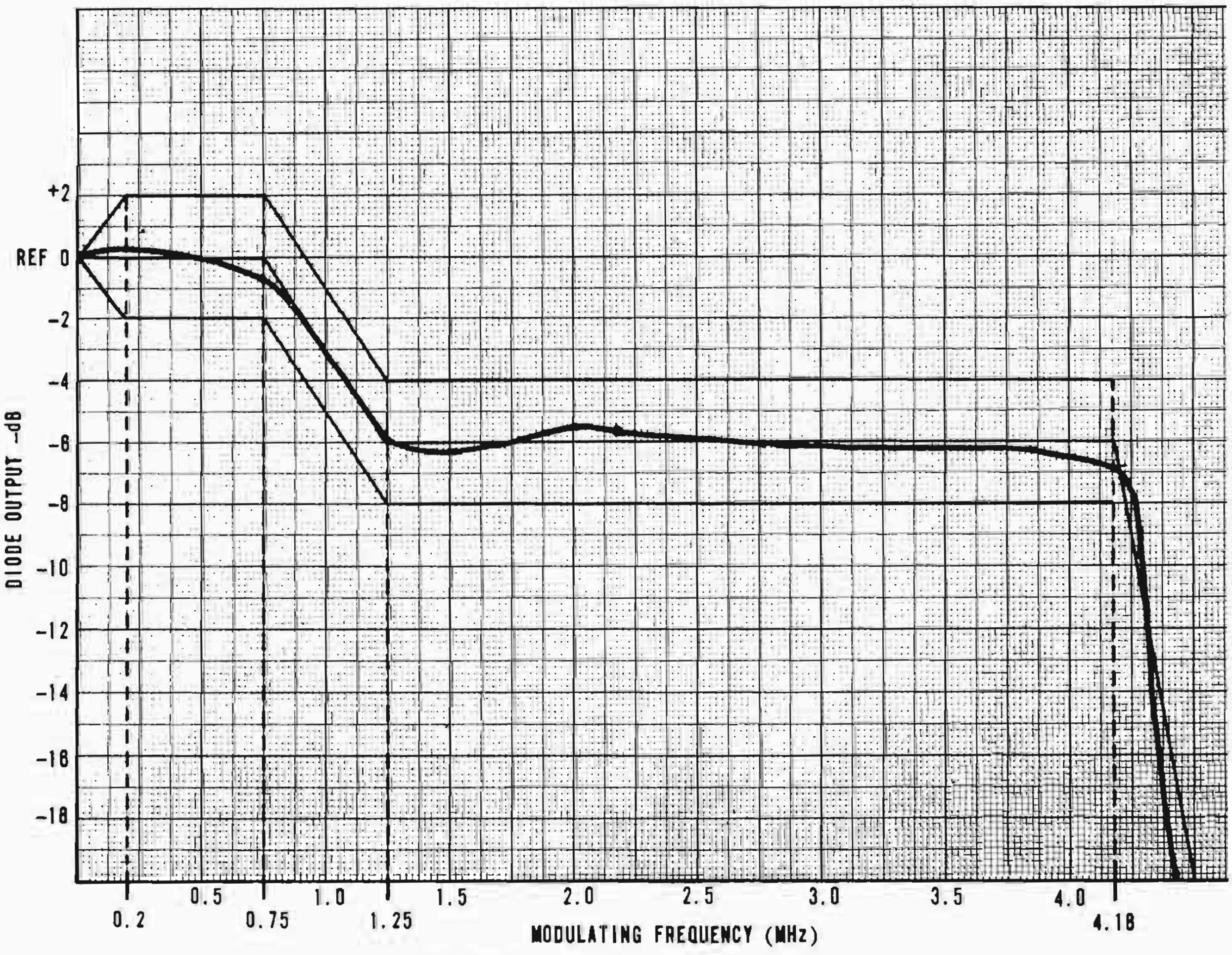


Fig. 91. Double sideband response diode.

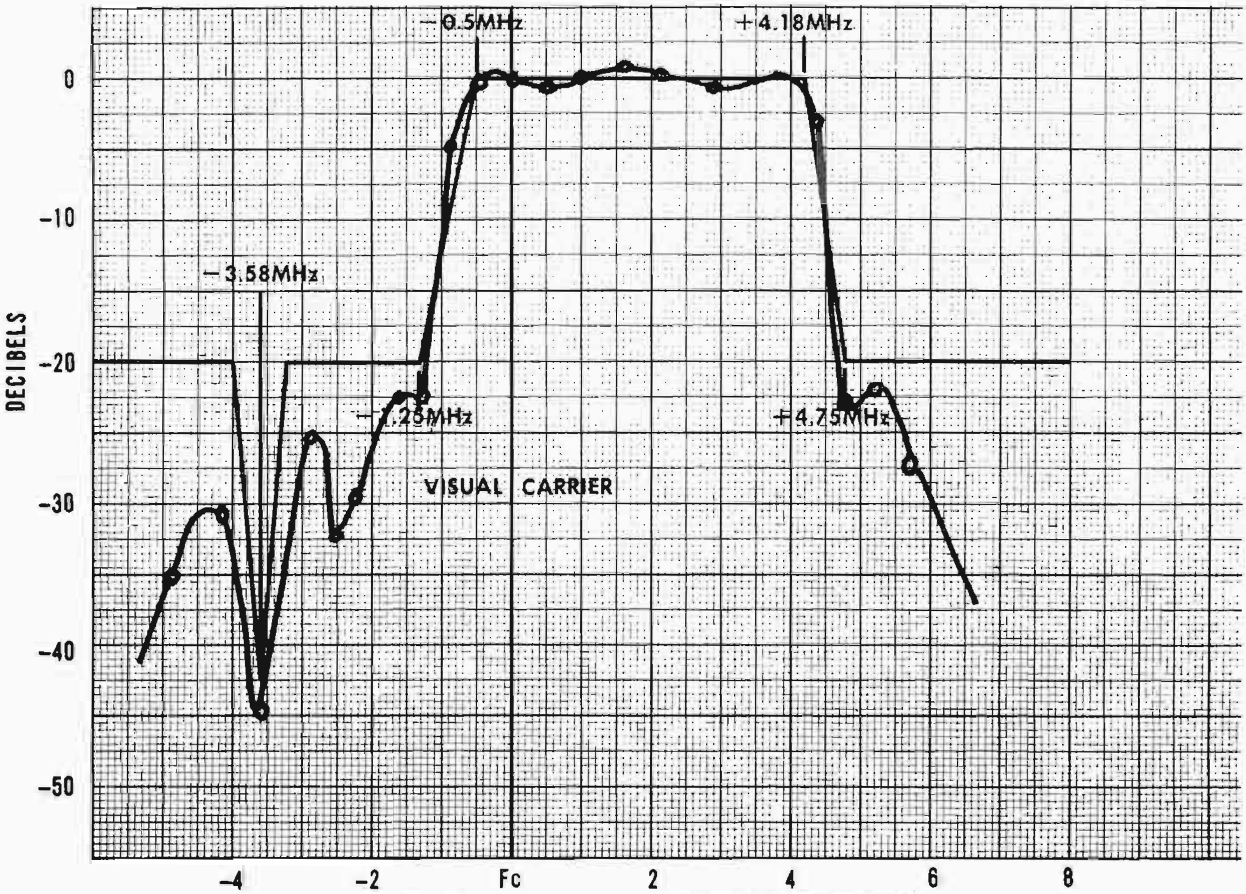
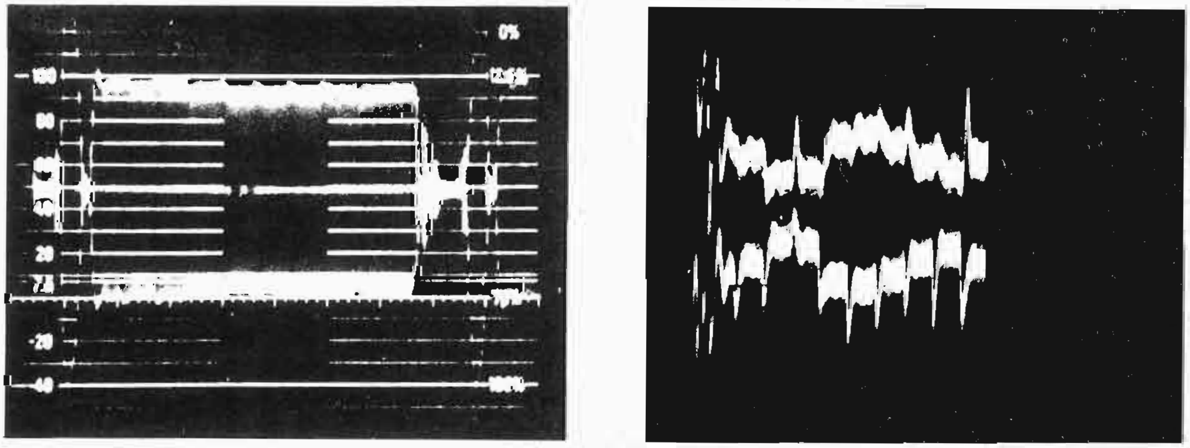


Fig. 92 Frequency response thru Rohde & Schwarz SWOF.



3.5% Fig. 93. Differential gain and phase 2.3%

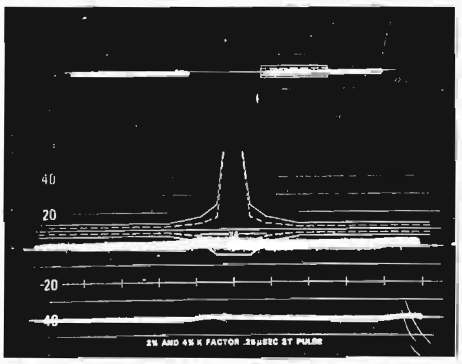


Fig. 94. Low frequency response (vertical interval).

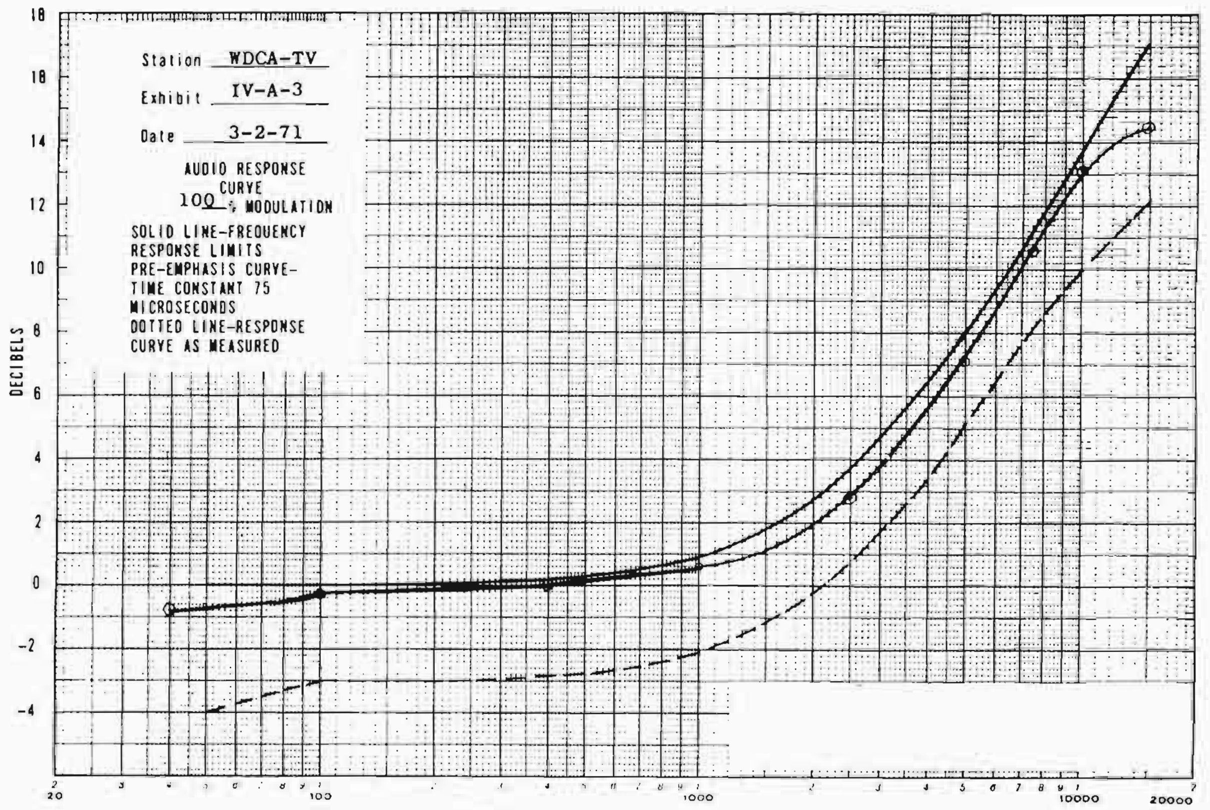


Fig. 95. Frequency response, aural transmitter.

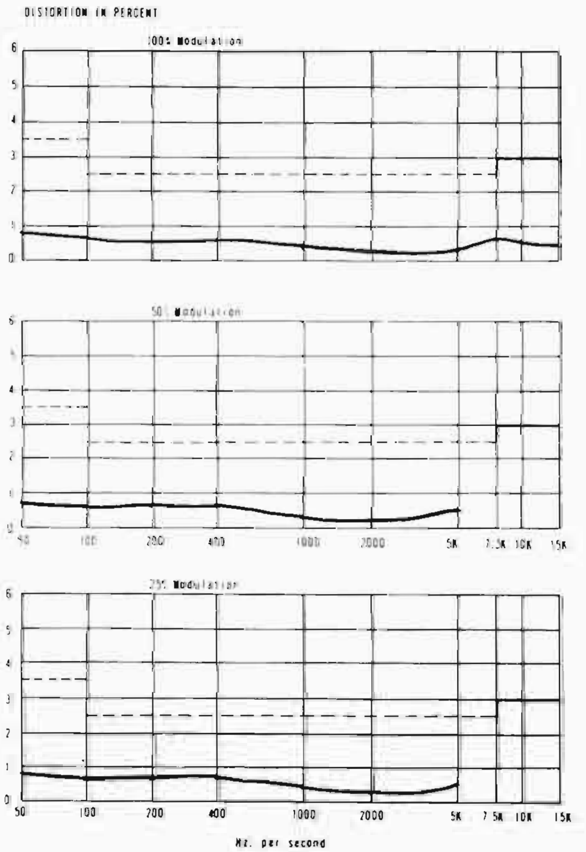


Fig. 96. Harmonic distortion, aural transmitter.

Further information, however, is needed to ascertain optimum performance in a color system. Most importantly, time coincidence between luminance and chrominance information must be shown by measuring envelope delay typically as shown in Fig. 97.

Very similar answers can be obtained from a test signal according to FCC Rules and Regulations Paragraph 73,699, Fig. 15. Typical results (employing a signal very similar to the above mentioned FCC composite test signal) are displayed in Fig. 98.

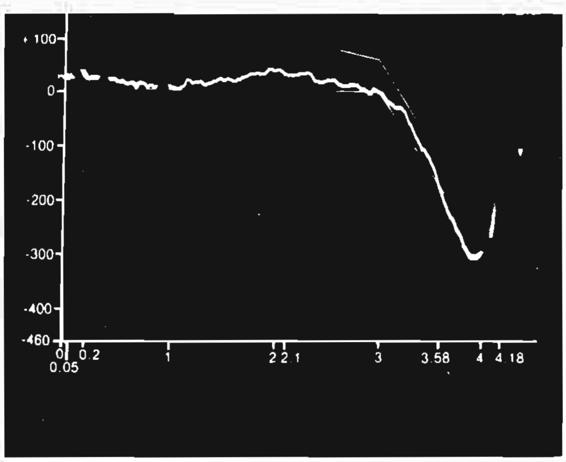


Fig. 97. Swept envelope delay.

One must keep in mind that many performance characteristics of a transmitter, for instance, frequency response and envelope delay, can vary with picture level; in other words, they are slightly different when measured at black as compared to white. To assess this deficiency many other test signals can be varied to produce high or low average picture levels. This is often accomplished by transmitting the test signal on one line to be followed by four lines which are either black or white.

An alternate very useful approach is to view the various test signals during the vertical interval, which allows the remainder of the available time to be operated at predetermined picture levels with transitions either at line or field rates.

Numerous excellent descriptions of the various techniques are available from test equipment manufacturers and many excellent papers have been published in the various journals, which can be consulted for Reference 30.

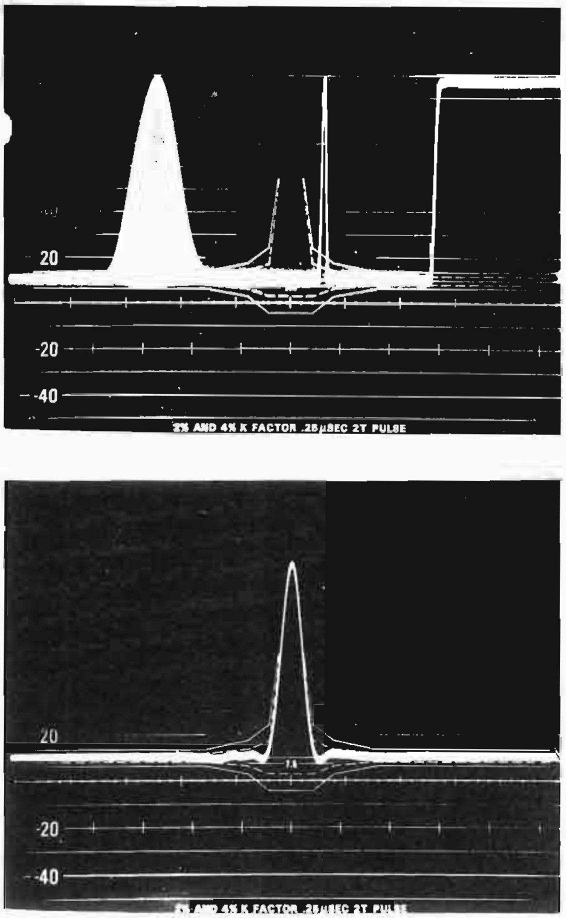


Fig. 98. Composite test signal response (2t, 20t).

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Television Transmitters

Addendum

T. M. Gluyas
RCA Broadcast Systems

Television transmitters in use in the United States today can be classified, broadly, into two basic types. One employes modulation at some moderate RF power level followed by linear power amplifiers and a high-level vestigial sideband filter at the transmitter output. A block diagram of such a transmitter is shown in Fig. 1.

The second type employes modulation at an IF frequency followed by a low-level IF vestigial sideband filter, an upconverter to final frequency at low power, and a chain of linear amplifiers on the final frequency to reach the desired power level. A block diagram of a system of this type is shown in Fig. 2. Signal levels shown in Figs. 1 and 2 are illustrative of common practice, but levels in specific designs may vary from these values significantly.

There is a third type which is a mixture of the first two. It employes modulation at IF frequency, an upconverter to final frequency but at comparatively high power (4 w), followed by final frequency linear amplifiers to the desired output power. The main difference between this variant and the basic system shown in Fig. 2 is that a low-level IF VSBF is not employed. Instead, a high-level final-frequency filter is used. A block diagram and a more complete description of a transmitter of this type will be covered later as part of a discussion of UHF-TV transmitters.

There are, as might be expected, variations of each basic type and many different circuit arrangements are possible for each major block of the diagrams. Some of these variations will be described and compared in later paragraphs.

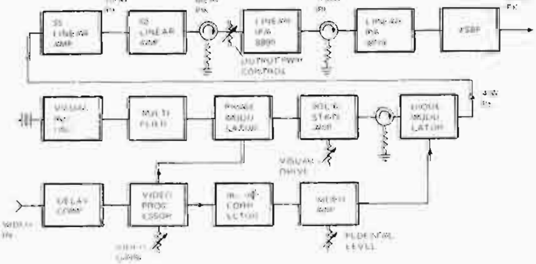

Fig. 1. Visual transmitter with modulation and VSBF at final frequency. Block diagram.

The basic Chapter 19, preceding this addendum, described fundamental principles and performance requirements of television transmitters in general. It provided a fairly complete description and many circuit details of television transmitters employing modulation at IF frequency. The design philosophy of these transmitters was stated and reasons were advanced for selecting the concepts and circuits chosen.

Similarly, in this addendum to Chapter 19, circuits will be described which are used in transmitters with final frequency modulation and high-level sideband filtering.

DESIGN CONSIDERATIONS

The prime considerations in design of a television transmitter are good technical performance and performance stability, reliability, maintainability and low cost—both capital cost and operating cost. The equipment designer must make some tradeoffs to achieve a proper balance among these and other factors, based upon personal or corporate judgment. Finally, equipment performance, reliability, operating cost, etc., depend not only upon the choice of circuit and packaging concepts made by the designer but also upon excellence in design details, product quality, and the operating and maintenance skills of the broadcaster.

Certain block diagrams or circuit choices do offer fundamental possibilities of improved per-

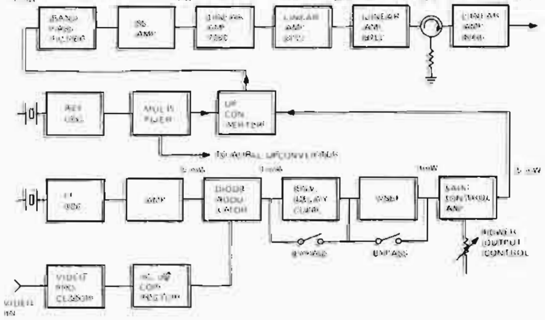

Fig. 2. Visual transmitter with modulation and VSBF at IF. Block diagram.

formance of one characteristics or another, but often at the expense of other desirable factors as illustrated in the following examples.

Modulation Levels

It is possible, and less expensive, to achieve better modulation linearity in a 1-milliwatt diode ring modulator at IF frequency than in a higher power diode modulator or grid bias modulator at final carrier frequency. On the other hand, low-level modulation requires a greater number of wideband linear amplifiers and, in general, more numerous tuning controls and adjustments.

When modulating at a higher level, linear amplifiers after the modulator are exchanged for cw amplifiers preceding the modulator. These need not be wideband and can be operated at nearly saturated output power for improved power output stability. An arrangement of this type was selected by RCA for the design of transmitters to be described in later sections of this addendum.

The following table lists modulation levels and associated linear amplifier gains required to reach a power of 25 kw for each of the three modulation methods in common use in VHF-TV transmitters. The ring modulator is followed by a small IF linear amplifier, an upconverter and, as for the other modulators, a succession of wideband final-frequency linear amplifiers. The 80 dB gain figure listed in the table for the ring modulator system makes allowance for a 6 dB upconverter loss.

Modulation method	*Mod. output power*	*Linear amp. gain required to reach 25 kw*
Ring modulator	0.001 w	80 dB
Switching mode diode modulator	2 w	40 dB
Grid bias modulator	1000 w	14 dB

Low Level versus High Level VSBF

It is believed to be less expensive to use a low-level (IF) vestigial sideband filter, and it is possible to design a more compact system than when using a high-level VSBF. However, the size advantage is minimal when a high-level notch-type aural-visual combiner is required, and there are some advantages to consider in favor of a high level filter.

Among these advantages are that it permits reduced transmitter power consumption, simultaneous good luminance and chrominance linearities, and a reduction of out-of-channel intermodulation products when a VSBF and other bandwidth limiting circuits are not included between the linearity corrector and the visual PA.

A linearity correction circuit generates distortion products that are out of phase with those generated in nonlinear modulators and imperfect "linear" amplifiers. Ideally, all the distortion products cancel out before reaching the PA output, thereby eliminating intermodulation products and restoring signal linearity at all modulating frequencies. This cannot happen as effectively if some of the distortion products are lost between the correction circuit and transmitter output in a low-level VSBF or cascaded linear amplifiers of limited bandwidth.

A lower power consumption with high-level filtering (or IF filtering followed by linearity correctors) is possible by operating the PA in Class B, with idling current near cutoff. Improved plate efficiency is achieved by accepting some PA non-linearity and compensating for this in video linearity correction circuits.

Fig. 3 is a set of spectrum analyzer photographs that illustrate the advantage of carrying the distortion products of the linearity corrector through the system to the PA input. In-channel intermodulation products cannot be seen as they are masked by useful sidebands but out-of-channel components can be seen. Second harmonic distortion of the 3.58 MHz color envelope (or third order intermodulation of main carrier and color subcarrier) gives rise to out-of-channel signals displaced 7.16 MHz from visual carrier. Fig. 3a is the output of the IPA (in this case a grid bias modulated stage) with linearity correction (differential gain correction) applied. Fig. 3b is the output of the PA ahead of the VSBF and Fig. 3c is the output after a high-level VSBF.

Without linearity correction for the PA, the 7.16 MHz components of Fig. 3b would be nearly as large as those of Fig. 3a. The indicated improvement is 12 dB and 15 dB for the upper and lower 7.16 MHz components. Of this amount, the PA circuit selectivity accounts for 3 dB and 6 dB, respectively. Even greater improvement would occur except for some bandwidth limitation of the IPA/PA interstage coupling network.

Another argument for a high-level filter at the transmitter output is that it is less apt to depart from specified attenuation limits through component aging or manipulation of adjustments than is an IF filter comprised of coils and capacitors.

Video versus IF Delay Equalizer

The design choice between video and IF (or RF) delay equalizers involves both theoretical and practical considerations.

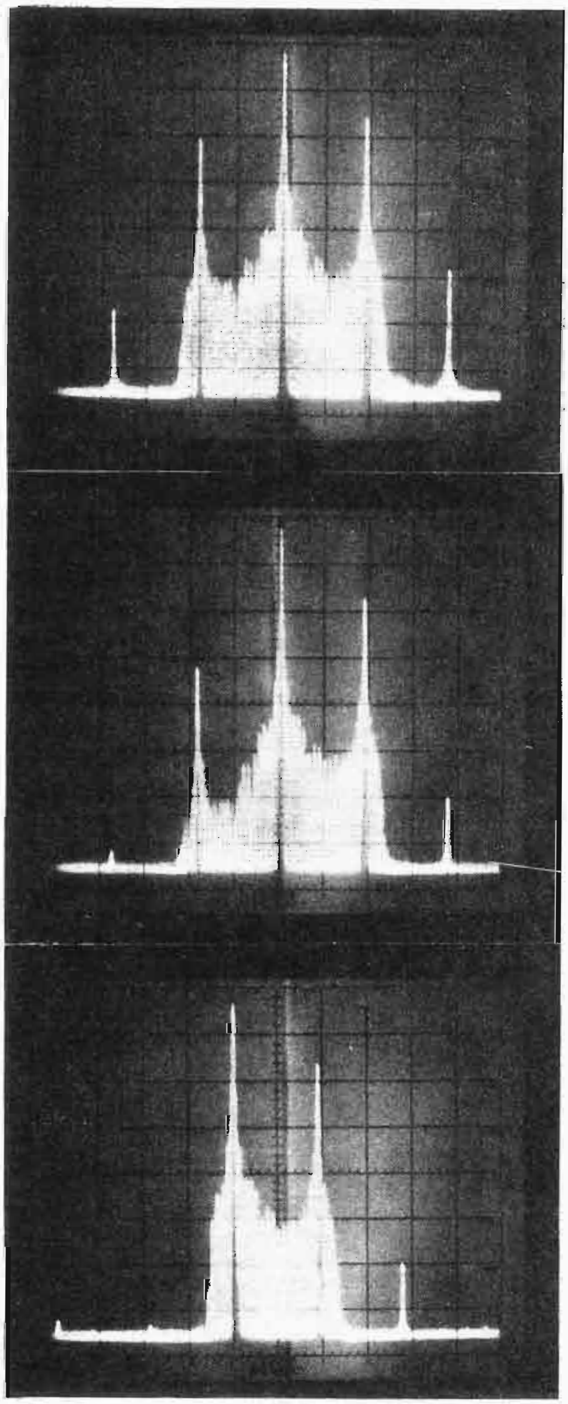

Fig. 3. RF spectrum of VHF-TV transmitter, transmitting color bars.
a. IPA (grid bias modulated amp.) output.
b. PA output ahead of vestigial sideband filter.
c. PA output after vestigial sideband filter.

For frequencies in the region of 1 to 4.18 MHz, since only one sideband is transmitted, there is no theoretical difference in system performance whether envelope-delay correction is carried out at video frequency or at IF frequency and both methods are in common use. The choice depends on the frequency range that permits the best practical design for a specific overall system.

For video frequencies in the region of 0.01 to 1.0 MHz (visual carrier ± 1 MHz) both sidebands are transmitted and the delays of upper and lower sidebands relative to carrier are different. A rigorous solution to delay correction in this frequency band would require that correction take place at IF (or RF) so that upper and lower sidebands could be equalized separately. In practice it makes little difference whether equalization is at baseband (video) or at IF. The system envelope delay of transmitter plus demodulator, for any one demodulator, can be equalized exactly by transmitter video delay equalization. The delay equalization may not be quite correct for other demodulators or receivers, but the residual delay error will be quite small. (See References 1 and 2).

Video delay equalizers have the advantage that required coil Qs are lower and more practical to attain, and the adverse effect of stray couplings (parasitic elements) is less than at IF.

One video equalizer, to be described later, employs multiple, switchable, delay sections with specific settings and corresponding delay characteristics defined by graphs or tables in an instruction manual.

Solid-State versus Vacuum Tube Linear Amplifiers

Solid-state circuits have, by degrees, replaced vacuum tube circuits in sequential transmitter designs until only the higher power RF amplifier stages still employ vacuum tubes.

In most designs, when transistors have replaced tubes in RF circuits, the circuits have been made wideband and/or fixed tuned. Consequently, solid state transmitters not only are more reliable but also require a minimal number of adjustments.

Reliability is not assured by the mere utilization of transistors since many things can, and have, caused transistor failures, especially in first-generation transistor circuits and in high-power applications. However, there can be little doubt that a properly designed solid-state circuit is more reliable than its vacuum tube counterpart when only a moderate number of transistors are required to take the place of each tube.

VHF-TV TRANSMITTER CHANNELS 2-6

Transmitters described in general terms in the introduction were of two basic types. Following is a more detailed description of the first type employing modulation at some moderate RF power, followed by linear amplifiers and a high-level VSBF at the transmitter output. Examples of this transmitter type are the RCA low-band model (Ch. 2-6) employing grid bias modulation at a

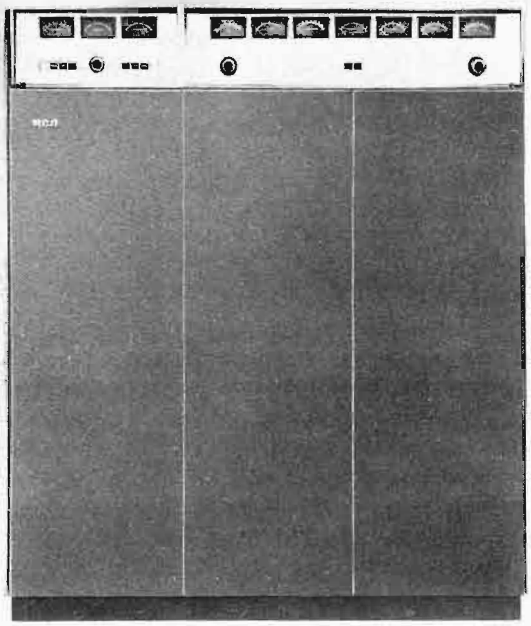

Fig. 4. RCA TT-25FL lowband VHF TV transmitter.

level of approximately 1 kw and the high-band model (Ch. 7-13) employing diode modulation at a level of approximately 2 w. The low-band design will be described first. Some elements of the low-band design also apply to the high-band design, to be described later.

A photograph of an RCA TT-25FL low-band transmitter appears in Fig. 4. It is a 25 kw transmitter for Channels 2-6. The photograph shows only the front line cabinets and not the complete transmitter which also includes a power supply cabinet and a vestigial sideband filter. All these assemblies are drawn to scale in the floor plan, Fig. 5. The power supply is approximately the size of the amplifier cabinet. It can be placed alongside the front line cabinets, but more often it is placed elsewhere as illustrated in the typical floor plan. An internal view of the power supply cabinet is shown in Fig. 6. Elements of the power supply unique enough to warrant description will be covered in later paragraphs.

A block diagram of the transmitter is shown in Fig. 7. The transmitter circuit is all solid-state except for the visual modulated stage and PA, and the aural IPA and PA. A total of five tubes are employed: quantity-3 8791-V1, quantity-1 3CX3000A7 and quantity-1 3CX20,000A7.

Circuits performing the functions titled in the block diagram will be described, with emphasis on the functions that are performed differently than in IF modulated transmitters which were described in the basic Chapter 19.

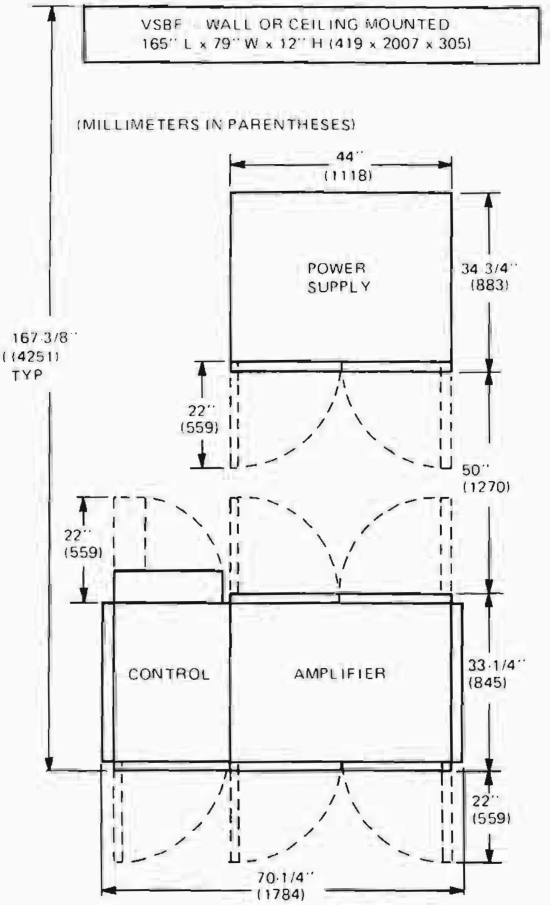

Fig. 5. Typical floor plan, TT-25FL transmitter.

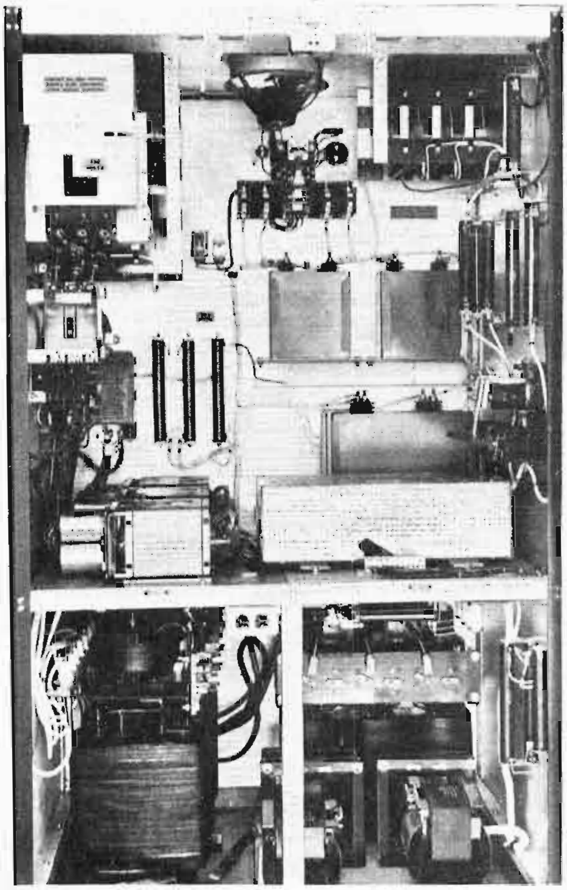

Fig. 6. Power supply cabinet, doors removed.

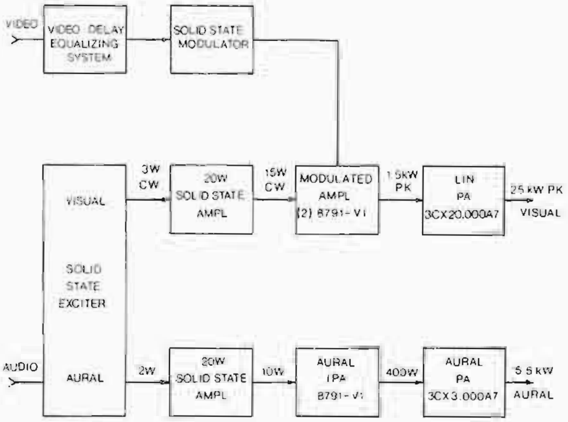

Fig. 7. TT-25FL transmitter, functional diagram.

Video Envelope-Delay Equalizer

With reference to Fig. 7, the video signal enters the transmitting system at the video envelope delay equalizer which is divided into the functional segments shown in the block diagram, Fig. 8. Each section of the equalizer can be switched out, as shown, for test purposes, but all are connected for normal programming. The envelope-delay equalized low-pass filter limits the maximum video frequency response to -20 dB at 4.75 MHz and above to meet the FCC rules for transmitter amplitude versus frequency response.

The low-frequency equalizer corrects mainly for the envelope delay distortion introduced by the vestigial sideband filter plus a lesser amount introduced by the transmitter bandpass coupling circuits. It is adjustable to any of 72 preset curves by combined operation of two switches—a 4-position switch and an 18-position switch. These curves, tabulated in the instruction manual, have different slopes and curve break points and are designed to match the delay of a variety of transmitter systems and different transmitter adjustments to within approximately 25ns.

The high-frequency equalizer corrects for the relatively small (approx. 50ns) high-frequency delay distortion introduced by the bandpass coupling circuits of the transmitter. However, for the many transmitting systems where an aural-visual notch diplexer is employed, introducing rapid visual signal cutoff between 4.18 MHz and 4.5 MHz, several hundred ns of additional high-frequency envelope-delay error are introduced. The delay equalizer allows for these extremes by providing delay correction ranging from 0 ns to 1,000 ns at 4.18 MHz. A choice of 39 curves is provided by combined operation of two switches—a 3-position switch and a 13-position switch.

The receiver equalizer provides -170 ns envelope-delay predistortion at 3.58 MHz for "typical" receivers as required by par. 73.687(a)(5) of the FCC rules. It must be kept "on" for normal transmission but should be switched "off" when the demodulator sound trap is turned "off" for transmitter performance measurements.

Video Circuits (Solid-State Modulator)

The video circuits of the transmitter connect after the video envelope-delay equalizer. Fig. 9 is a simplified block diagram of the video circuits.

Video input amplifier. A combination of a differential input amplifier and a fast-acting clamp circuit eliminate hum and low-frequency noise from the video input signal. The differential amplifier rejects any common mode noise on the input signal rising from possible ground loop problems in the coaxial cable connection between terminal equipment racks and the transmitter proper.

The first clamp circuit not only reduces input signal hum and noise but it also reinserts the dc component of the video signal as required ahead of the differential phase and gain correctors. This makes the applied corrections fall at specific brightness levels on the video amplifier transfer characteristic independent of picture signal APL.

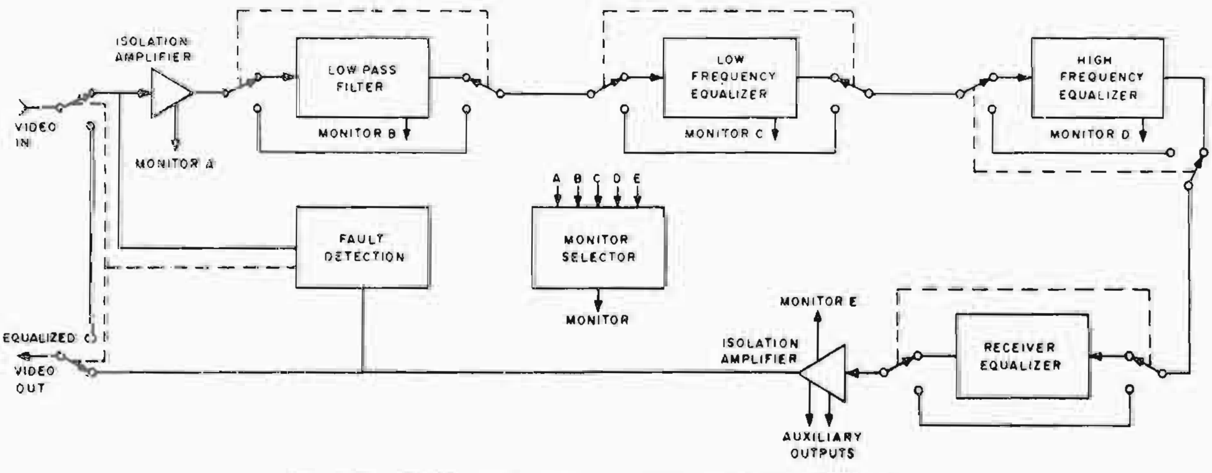

Fig. 8. Type TTS-1 video delay equalizer, functional diagram.

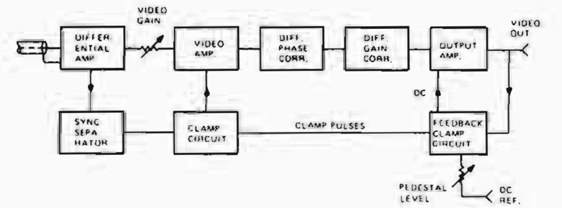

Fig. 9. Video amplifier and processor, simplified block diagram.

Differential phase and gain corrector circuits are connected in tandem in RCA VHF-TV transmitters, and can be adjusted with negligible interaction between the two. A wide range of control is available, but a generous number of controls permits precise correction (complementary matching) of any ordinary differential phase or gain error likely to be encountered.

A separate but similar circuit controls sync stretch.

Output amplifier. With reference to Fig. 9, the dc component is removed (AC coupled) in the output amplifier and then promptly reinserted by a feedback clamp circuit referenced to the video output signal. This holds the pedestal level constant at the amplifier output, independent of any long-term drift accumulated in the cascaded video amplifiers.

The output stage is a complementary symmetry video amplifier capable of delivering up to 70 Vpp to the grid bias modulated RF amplifier. Approximately 40 Vpp normally is required for 25 kw at the transmitter output.

The bandwidth of the video amplifier, including the output amplifier, exceeds 8 MHz. Consequently, video harmonics generated in the nonlinear differential gain correction circuits are carried through to the grid bias modulated amplifier, as required for proper correction of modulated amplifier and "linear" amplifier nonlinearities.

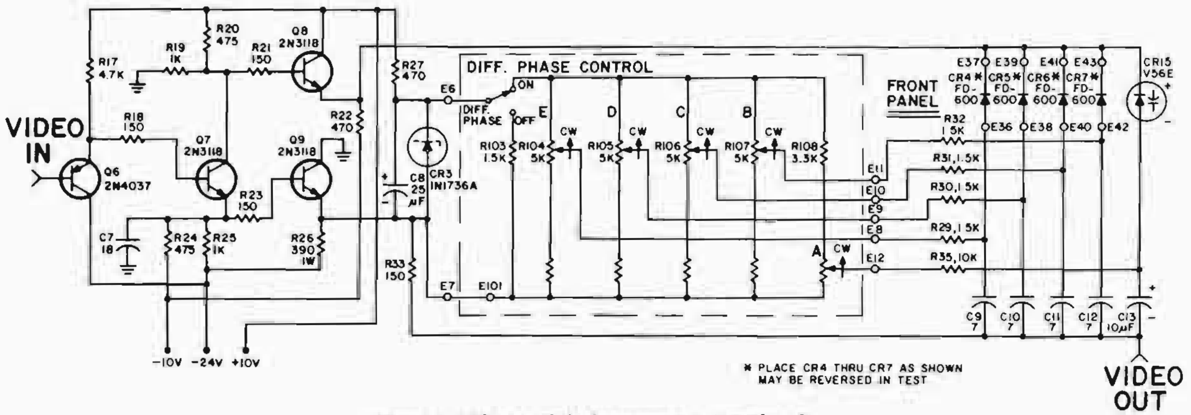

Fig. 10. Differential phase corrector circuit.

The differential phase corrector circuit, shown in Fig. 10, operates in the following manner. Video output is extracted from a pair of push-pull emitter followers. The principal output is from one emitter through resistor R33. To this is added a small capacitive-coupled output from the opposite emitter through any or all of four gated capacitors C9-12 and a varicap CR-15. The added capacitive coupled signal shifts the color subcarrier phase by up to $\pm 2°$ for each gated capacitor and $+8°$ maximum for the varicap. The brightness level at which the capacitors are electronically connected is determined by controls B through E and the effectiveness of the varicap is controlled by A.

The differential gain control, or more properly "linearity control" circuit modifies the video amplifier gain as a function of instantaneous picture brightness. This is accomplished by gating a succession of five resistors across the coupling circuit between two transistors. The resistors have different values and each can be gated at any desired brightness level, thereby achieving a variety of smoothly changing transfer functions.

Grid Bias Modulated RF Amplifier

The output of the video amplifier (solid state modulator) is connected to the grid bias modulated RF amplifier as shown in the simplified schematic, Fig. 11. The push-pull arrangement provides a way of introducing the video at an RF

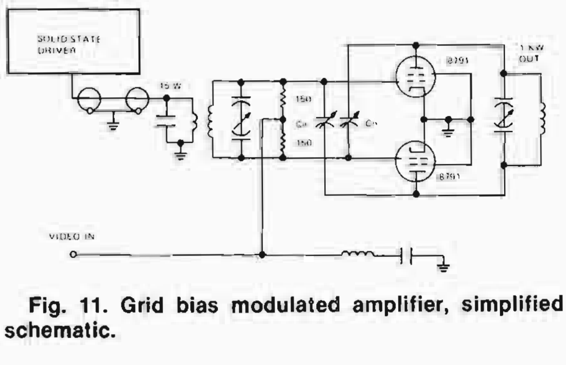

Fig. 11. Grid bias modulated amplifier, simplified schematic.

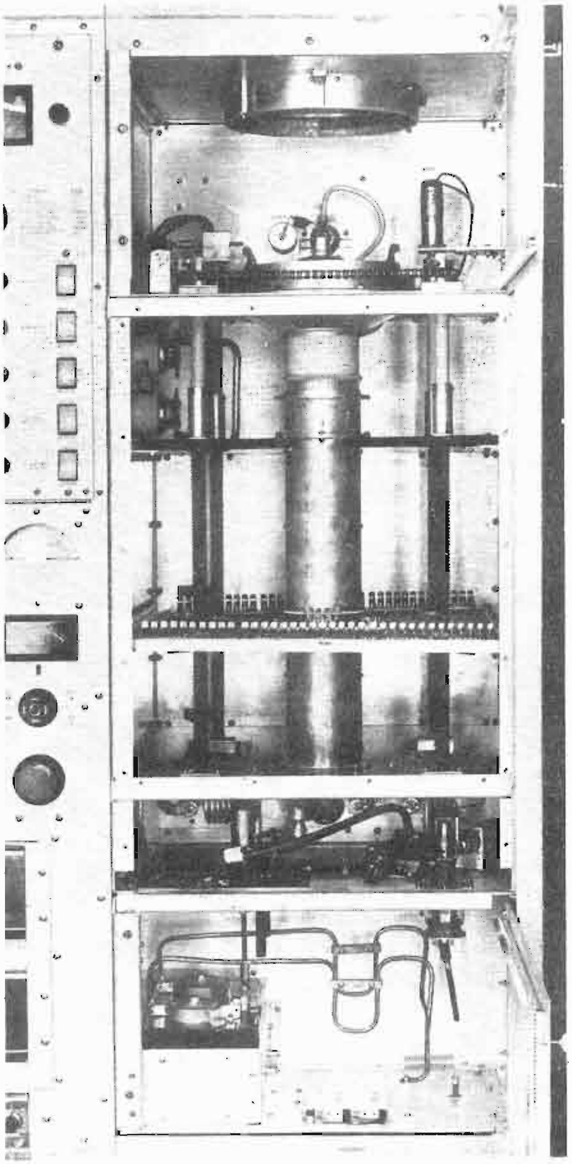

Fig. 12. Visual modulated amplifier and PA.

null point and avoids burdening the video system with the added capacitance of an RF bypassing and filtering scheme. It also enables wideband neutralization by virtue of the simple capacitance bridge neutralization circuit.

A very special point is that the input circuit, although carrying only a cw signal, nevertheless is a wideband circuit just as if the stage were operated as a visual linear amplifier. As a result, any residual feedback or load modulation caused by varying input impedance of the modulated stage, is independent of modulating frequency and does not produce waveform distortion. Residual modulation, if any, is very small because of improved neutralization, better tubes with lower input circuit loading, and the tubes are not driven into grid current, even at sync peak level. These circuits do not exhibit the waveform distortions sometimes encountered in older designs.

Visual Power Amplifier

The visual PA is a hi-mu zero-bias grounded-grid triode linear amplifier. The use of a triode has notable advantages—and like most engineering choices, some disadvantages. In this case, the advantages strongly outweigh the disadvantages.

The zero-bias triode eliminates the requirement for: a regulated bias supply; a regulated screen grid supply; associated metering, control, and protection circuits; and screen grid bypassing and tube contact pieces in the cavity. These eliminations result in a simple and reliable circuit.

The disadvantages of a triode as compared with a tetrode are: somewhat higher output capacitance and therefore a lower gain-times-bandwidth product; higher feedback capacitance (not troublesome on Ch. 2-6); and higher grid current.

Fig. 12 is a photograph of the visual modulated amplifier and visual PA. The front panel of the PA cavity has been removed to make the interior visible. The PA anode is bypassed to the top of the cavity to eliminate the anode radiator-to-ground capacitance from loading the circuit and consequently reducing the bandwidth.

It may be hard to interpret the cavity circuit of Fig. 12 as a grounded-grid circuit. Clearly, the grid terminal is not physically grounded as it can be seen where the spring fingers connect at the bottom of the anode-to-grid ceramic insulator. If the circuit is renamed a "common-grid" amplifier instead of a "grounded-grid" amplifier, the circuit operation should become clear. The terminology parallels that of a transistor amplifier which may be called a grounded base amplifier or a common base amplifier, interchangeably.

The PA input circuit (cathode-to-grid) is inside the copper pipe in the center of the cavity. The PA output circuit primary (anode-to-grid) is in the space inside the box and outside the pipe. Just as in a conventional grounded-grid circuit, there is no electromagnetic coupling between input and output circuits.

Aural Exciter

Fig. 13 is a functional diagram of the aural exciter and low power solid-state amplifiers.

The varicap controlled FM oscillator is in a temperature controlled oven. It operates at

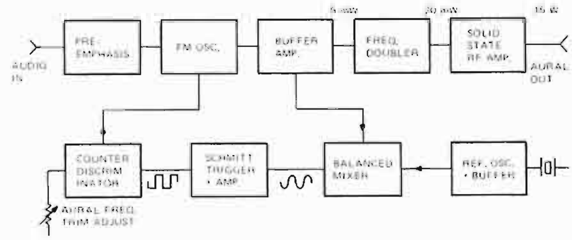

Fig. 13. Aural exciter and low power amplifiers, functional diagram.

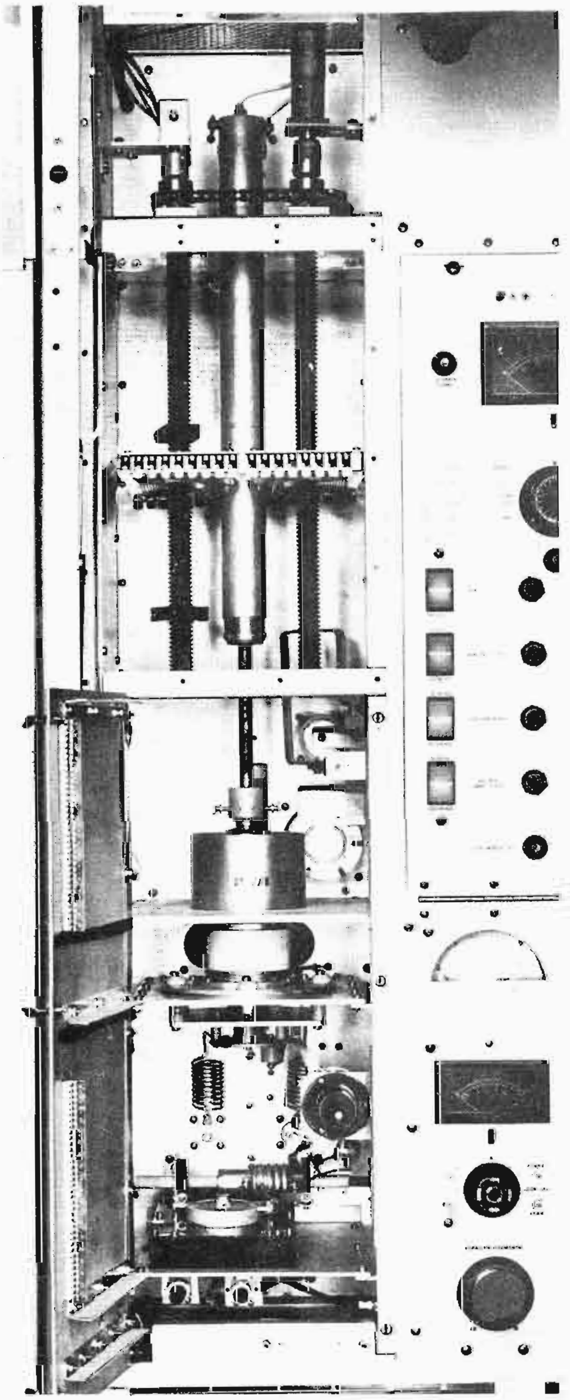

Fig. 14. Aural IPA and PA assembly.

one-half carrier frequency. One varicap controls the aural frequency deviation and a second controls the center frequency through an AFC loop.

A crystal controlled reference oscillator is offset from the FM oscillator by a convenient counting frequency of 150 kHz. The 150 kHz difference frequency is obtained in a balanced mixer then changed to a square wave and applied to a counter type discriminator. The discriminator develops a dc output voltage directly proportional to frequency. The developed voltage is offset by a highly stabilized dc reference so that the net output is zero when the counting frequency is exactly 150 kHz. The control voltage swings plus or minus if the FM oscillator should tend to drift above or below the assigned frequency.

The loop gain of the AFC system is 48 dB which ensures that the FM oscillator center frequency is closely held with respect to the temperature controlled crystal reference oscillator.

A succession of broad-band solid-state amplifier stages raises the aural signal to a level of approximately 15 w to drive the two tubes in the aural transmitter—the IPA and PA tubes.

Aural IPA and PA

A photograph of the aural IPA and PA enclosure with the doors open is shown in Fig. 14. The Type 8791 V1 IPA tube operates grounded cathode employing lumped constant input and output circuits.

The PA, employing a 3CX3000A7 triode, operates grounded-grid with a coaxial cavity output circuit. Unlike the visual PA, the aural grid is directly bypassed to the cabinet, forming a ground plane to isolate input and output circuits. This is practical because, in the case of the aural PA, the capacitance of the tube anode radiator to ground is of no concern. Consequently, the aural cavity is simpler in construction than the visual cavity.

Power Supply

The power supply cabinet (Fig. 6) contains a high voltage supply for the visual PA and an intermediate high voltage supply for aural PA, visual IPA and visual PA stages. Both power supplies employ three-phase full-wave silicon rectifier circuits.

A distribution transformer and three constant voltage transformers rest on the shelf above the high voltage transformers. Another transformer, in the transmitter control cabinet, isolates the control circuit ladder from the incoming 230 v, 3 Ø power line for safety and provides 117 v ac to circuits requiring the lower voltage.

The constant voltage transformers regulate the filament voltage to all vacuum tubes to a stability of 1 percent for extended tube life. This is an important feature. In the absence of the many other things that cause catastrophic tube failure or premature performance degradation, wearout in a tube with thoriated tungsten filament is caused by loss of emission through depletion of the carbide layer. It has been calculated that the theoretical tube life is reduced 2/1 by a 3 percent excess of filament voltage over the life span of the tube.

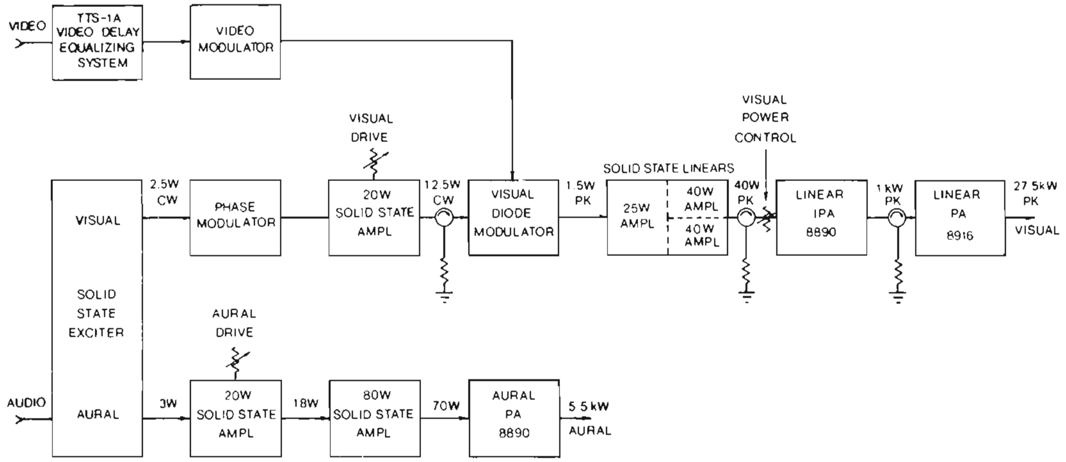

Fig. 15. TT-25FH transmitter, functional diagram.

Filament power supplies for tubes with directly heated cathodes are dc to minimize hum which can originate, in part, from magnetic deflection of the electron stream within the tube. The hum has high harmonic content and the magnitude is not stable with time. Consequently, hum bucking circuits are less effective than eliminating hum at the source through the use of 3-phase full-wave dc filament supply.

VHF-TV TRANSMITTER CHANNELS 7-13

In overall size and external appearance, the RCA highband transmitter (Ch. 7-13) is nearly identical to the lowband transmitter (Ch. 2-6) previously illustrated (Fig. 4).

The left-hand rack contains control circuit logic, exciter, modulator, solid-state aural amplifiers, solid-state visual linear amplifiers, and the solid-state power supplies. In brief, solid-state circuits are located in the left-hand rack and power tube circuits are located in the right-hand rack.

The right-hand rack contains the three cavities for the only three tubes in the transmitter, motor-driven cavity tuning controls and dc power supplies (located in the air plenum) for the power tube filaments.

A block diagram of the transmitter is shown in Fig. 15.

The envelope delay equalizer, video circuits and solid-state exciter are similar to the corresponding blocks in the lowband transmitter which have been described.

Visual Diode Modulator

A simplified schematic of the visual modulator is shown in Fig. 16. Two diode bridge modulators are employed, connected in quadrature, and they are operated at comparatively high power—not in the square law region.

The diodes are biased by the video signal and pass more or less of the RF cycle, depending upon the instantaneous video potential. The capacitance of the diode assembly is neutralized by coil Ln so that negligible signal is passed when the diodes are biased off. The diode modulator operates with an insertion loss of approximately 5 dB at sync peak.

The tuned circuits L1C1, L2C2 are tuned to visual carrier frequency to suppress harmonics generated in the modulation process.

The input impedance of each diode assembly varies widely over the brightness range (instan-

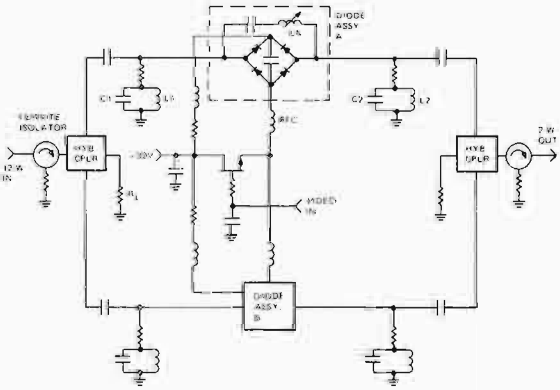

Fig. 16. Video diode modulator, simplified schematic.

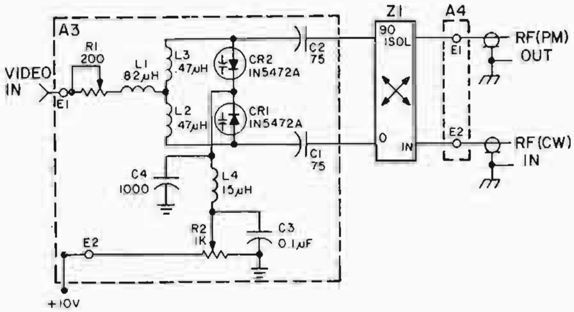

Fig. 17. Phase modulator, simplified schematic.

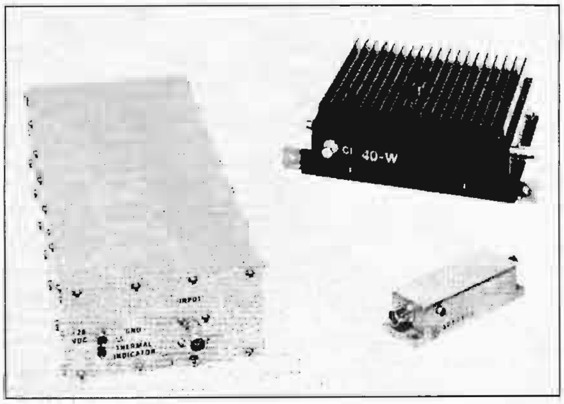

Fig. 18. Solid-state amplifiers: 20 w, 40 w, 80 w.

taneous video signal) and this would cause an undesired modulation of the input signal, except for the hybrid coupler. If the two diode assemblies are identical, and the hybrid is well balanced, the varying reflections from the diodes end up in the reject load R_L. Therefore, the input impedance to the coupler remains a constant good match to the 20-watt solid-state carrier source. The ferrite isolator removes residual reflections due to imperfections in the hybrid coupler or minor differences between the diode assemblies.

Phase Modulator

The purpose of the phase modulator is to correct for transmitter incidental phase modulation. It accomplishes this by introducing phase modulation into the visual carrier (cw) path, ahead of the visual modulator, of the correct amplitude and polarity to cancel unwanted phase modulation occurring elsewhere in the transmitter. The result is a transmitted signal essentially free of any incidental phase modulation at all modulating frequencies.

Adverse effects of uncorrected incidental phase modulation, which are avoided by the design of this transmitter, are:

- Noise ("buzz") in the aural output of intercarrier sound receivers. However, incidental phase modulation is only one of several possible causes of this problem.
- When differential phase has been corrected using a demodulator with envelope detection, it is then incorrect for synchronous detection.
- Envelope-delay correction required is different at different brightness levels. One result is that "sync spikes" may be excessive.

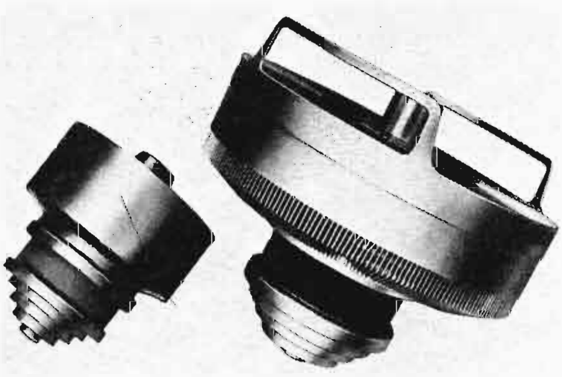

Fig. 19. Power tubes employed in TT-25FH, tube types 8890 and 8916.

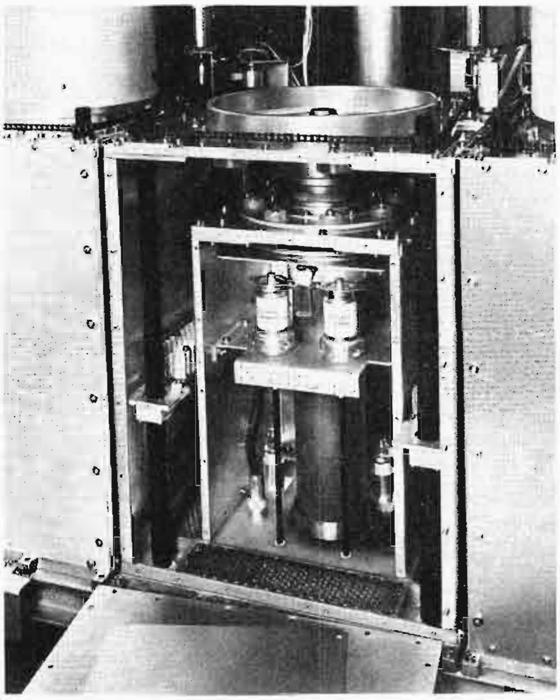

Fig. 20. View inside RF amplifier cavity.

Fig. 17 is a simplified schematic of the phase modulator which operates as follows. A cw signal at visual carrier frequency is applied to the input port of a 3 dB hybrid. The normal output ports are connected to varicaps. Since these present nearly identical reactive loads, the signals at the output ports are reflected and combine in the "isol." port; and this becomes the phase-modulated RF output signal.

The phase shift through the 3 dB hybrid depends upon the magnitude of the two reactive loads which in turn is controlled by the instantaneous value of the video signal applied to the varicap. The video input signal is shaped in circuits similar to a differential gain corrector to control the degree of phase modulation versus brightness level.

Solid-State Amplifiers

Both aural and visual solid-state amplifier stages have approximately 80 w capability, but the designs are different. The visual linear amplifier stage is, in fact, two separate 40 w linear amplifiers operated in parallel and combined by means of input and output 3 dB quadrature phase hybrids. Each 40 w amplifier uses a pair of transistors, operated in push-pull.

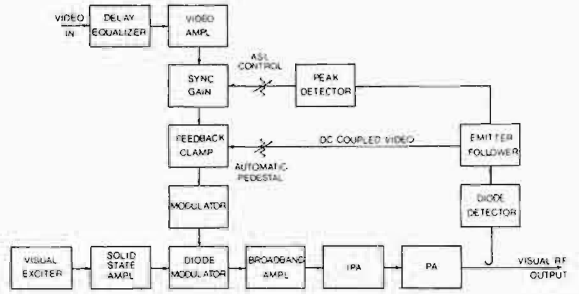

Fig. 21. Automatic power control functional diagram.

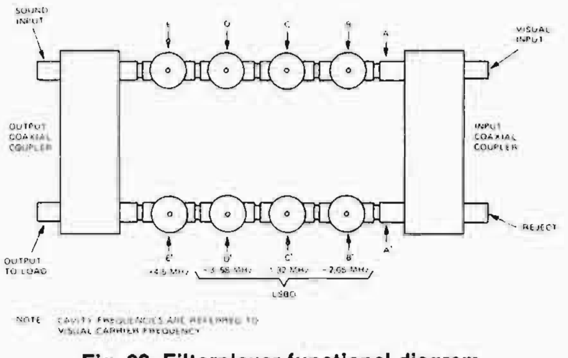

Fig. 23. Filterplexer functional diagram.

Fig. 18 is a picture of the 20 w (cw or aural), 40 w visual linear, and 80 w aural solid-state amplifiers employed in the transmitter. They are essentially "black-box" devices with no adjustments and need not be accessible in the transmitter except to facilitate removal if required for replacement or bench repair.

IPA and PA Power Tube Stages

There are only three tubes in the transmitter. They perform the following functions: visual IPA, visual PA, and aural PA. The visual IPA and aural PA employ the same tube type so only two tube types are required—an 8890 and 8916, shown in Fig. 19. Both are tetrodes. Unlike the case of the lowband transmitter, no triode tube is available for operation on Channel 7-13 with adequate power, gain, low feedback and good tuning characteristics in a practical Channel 7-13 cavity. Modern tetrodes permit wideband operation with the high gain necessary for compatible operation with the moderate-power solid-state amplifiers already described.

All three tubes are used in the same basic RF cavity, one of which can be seen in Fig. 20. The front plates of the inner and outer boxes of the cavity have been removed to provide a view of the internal construction and also to illustrate that disassembly for maintenance or repair is not especially difficult.

An assembly change is made to adapt the cavity to one tube type or the other, using interchangeable anode contact assemblies. A further variation is that the secondary plate circuit of all three cavities are different. This does not show in the photograph because the secondary circuits are comprised of assemblies of coaxial line sections mounted on the rear side of the cavities.

The cathode driven input circuit is a pi network, coupling between a 50-ohm RF drive source and the somewhat lower input impedance of the power amplifiers.

The pi network is adjusted by two vacuum variable capacitors that can be seen in the photograph. They are simply adjusted for minimum reflected power under normal operating conditions.

Automatic Power Control

The transmitter includes automatic power control circuits which operate as shown in the block diagram, Fig. 21.

A sample of the visual output signal is detected and the sync peak and pedestal levels are compared to reference levels. The error voltages from

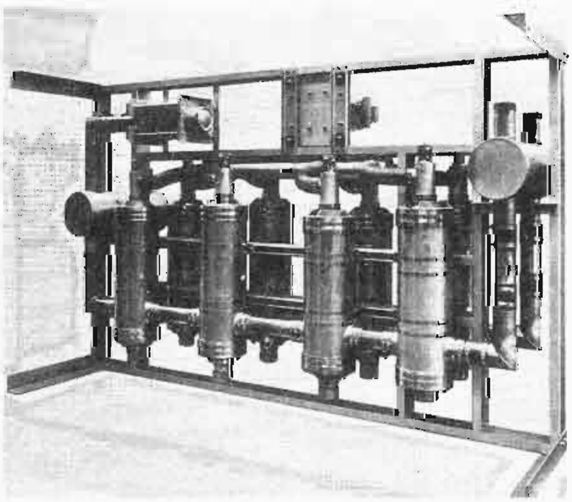

Fig. 22. A 50 kw high-band VHF filterplexer.

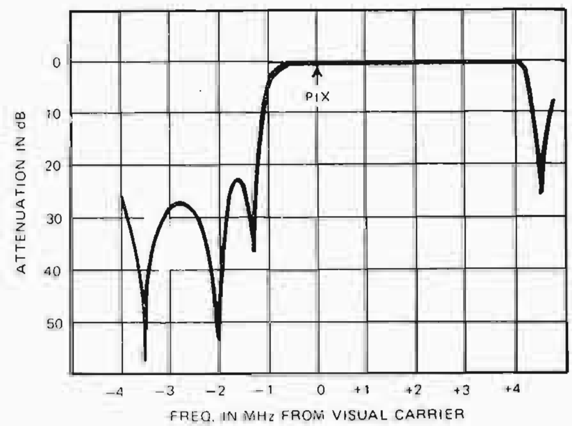

Fig. 24. Filterplexer attenuation characteristic.

the comparators are applied to sync gain control and feedback clamp circuits to control visual peak power and pedestal level, respectively. Thus the transmitter licensed power and the correct sync-to-picture ratio are automatically regulated. This capability is especially useful in contemporary remote control operation and will be useful in future automatic transmission systems.

Switches are included for turning off either or both automatic control circuits as an aid in initial setup of transmitter operating levels.

VSBF or Filterplexer

A harmonic filter is connected at the output of the aural PA; and a harmonic filter plus a vestigial sideband filter are connected at the output of the visual PA. Then aural and visual signals are combined.

If a turnstile antenna is employed with a two-transmission line feed system, then aural and visual transmitters can be combined in a broadband bridge diplexer and there is no need for a 4.5 MHz notch filter in the visual output system. On the other hand, if an antenna is selected with a single transmission line feed system, then aural and visual signals must be combined in a notch diplexer following the visual VSBF.

An alternate arrangement that takes less space, is less expensive, and often results in better performance is to combine the functions of VSBF, and notch diplexer in one assembly called a filterplexer. Fig. 22 is a picture, Fig. 23 is a functional diagram, and Fig. 24 is the attenuation characteristic of a high-band 50 kw filterplexer.

In the functional diagram (Fig. 23), observe that there are four pairs of resonant cavities connected between an input and an output 3 dB coaxial coupler. The resonant frequencies of the four pairs of cavities correspond to the four notch frequencies in the attenuation curve, Fig. 24.

With reference to Fig. 23, the visual signal splits in the 3 dB coupler into two equal signals, one delayed from the other by 90°. When the signals encounter a pair of cavities, the signal frequencies corresponding to the cavity resonant frequencies are reflected. The signal delay in one path is increased by an additional 90° in the return trip through the coupler. Consequently the reflected frequencies are routed to the reject load and dissipated. Signals which are not reflected combine in the output coaxial coupler and arrive at the output port.

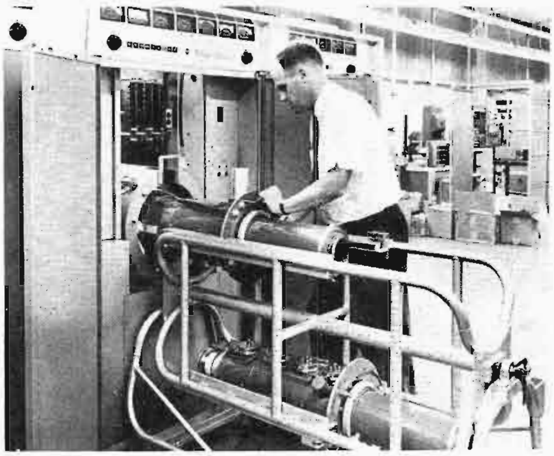

Fig. 26. Replacing a 30 kw 4-cavity klystron.

The aural signal, applied to one port of the output coupler, is reflected by a pair of cavities. The reflected signals combine and also arrive at the output port where both visual and aural signals are combined and routed to the antenna or to a test load.

UHF-TV TRANSMITTER

The vast majority of modern UHF-TV transmitters in service in the United States employ internal cavity klystrons manufactured by Varian Associates. The usual power output ratings of these klystrons for visual service are either 30 kw or 55 kw per klystron. Transmitter power output ratings are in multiples of these values with power outputs ranging from 30 kw to 220 kw. Except for power supply and cooling components, there is little difference in transmitter configuration, size,

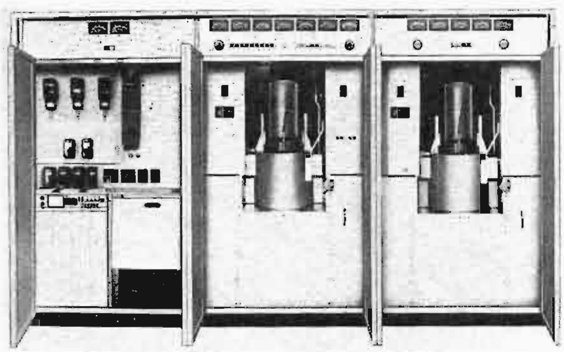

Fig. 25. RCA TTU-30C, 30 kw UHF TV transmitter, front line cabinets.

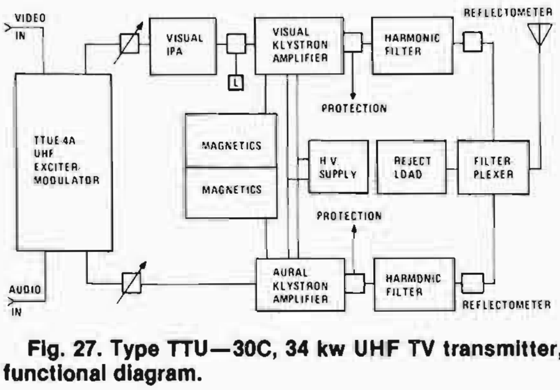

Fig. 27. Type TTU—30C, 34 kw UHF TV transmitter, functional diagram.

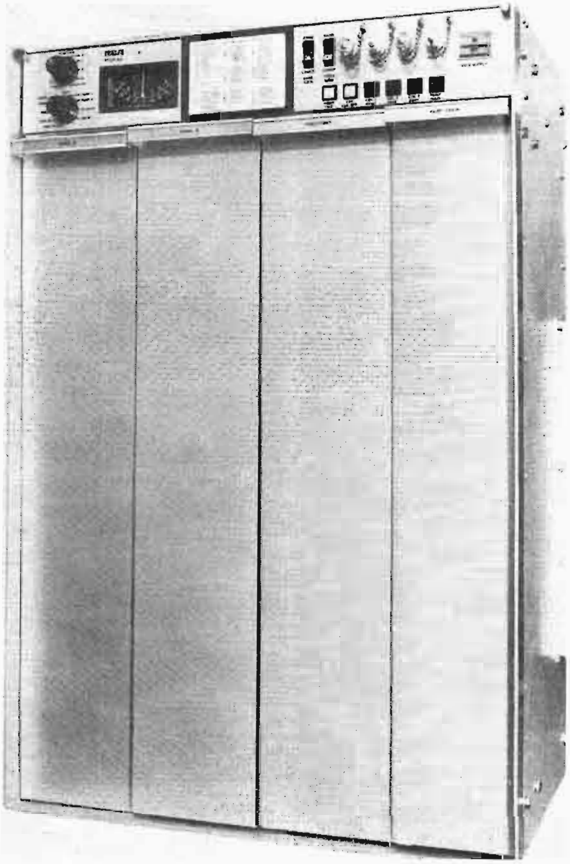

Fig. 28. UHF TV solid-state exciter-modulator.

or appearance whether 30 kw or 55 kw klystrons are employed.

A further classification of klystrons that can be made is into 4-cavity and 5-cavity types. The klystron power gain is just under 10 dB per cavity or roughly 40 dB and 50 dB for 4-cavity and 5-cavity klystrons, respectively.

RCA UHF-TV transmitters use 4-cavity 30 kw klystrons in some models and 5-cavity 55 kw klystrons in others. Five-cavity klystrons eliminate the need for a 10 w solid-state linear amplifier preceding the klystron, shown in one of the diagrams to follow. However, four-cavity klystrons

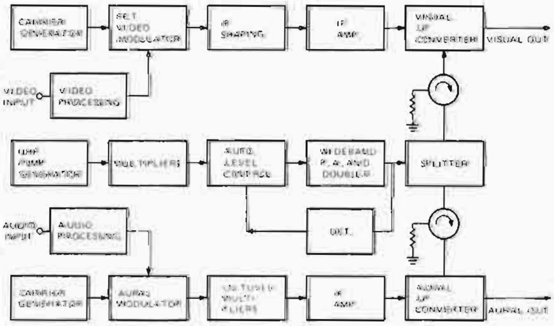

Fig. 29. Exciter-modulator, functional diagram.

are less expensive to replace and consequently permit slightly lower operating cost. A short discussion of power costs, which is a more significant element of operating cost, may be found in a subsequent paragraph.

Fig. 25 is a picture of the front line cabinets of an RCA TTU-30C, 30 kw UHF-TV transmitter with the cabinet front doors open to provide an internal view. The all-solid-state exciter is in the lower left corner of the left-hand cabinet. In each of the other two cabinets can be seen a 30 kw vapor cooled klystron, mounted in its magnet assembly. This may be pivoted forward for easy removal of the klystron as shown in Fig. 26.

A block diagram of a 30 kw transmitter is shown in Fig. 27. The transmitter is comprised of an exciter, a visual solid-state IPA, two klystrons, power supplies, a filterplexer, and some lesser subsystems and components.

Solid-State Exciter Modulator

An external view of the exciter modulator is shown in Fig. 28 and the related block diagram is shown in Fig. 29. There are four vertical slide-out drawers organized into the following basic functions: aural IF, visual IF, video processor, and pump chain. Each drawer houses a number of plug-in modules and insofar as practical, each module performs one basic function. The module performance can be metered on a multi-meter by combined operation of a nine-position switch and a ten-position switch. The concept is to minimize downtime if a fault should occur by rapid location and replacement of the defective module.

Visual and aural modulation occur at IF. The IF visual carrier frequency is 45.75 MHz and the aural frequency is 50.25 MHz. The aural frequency is higher than the visual frequency as a result of operating the pump circuit below the frequency of the UHF TV output signal—a desirable condition for maximum stability of the upconverter.

Video processing. The video processing circuits perform the same functions as those carried out in the VHF circuits, previously described. The circuits are similar. However, the quantitative linearity correction requirements are different for a UHF klystron transmitter, and the video signal level required to operate the FET modulator is

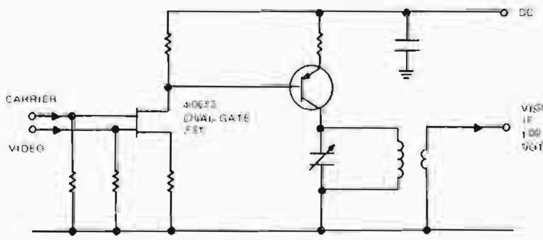

Fig. 30. FET visual modulator, simplified schematic.

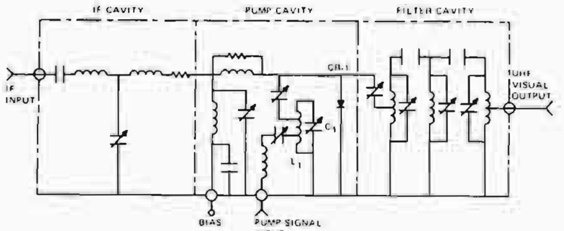

Fig. 31. Upconverter, simplified schematic.

less than required for VHF grid bias or diode modulators.

The UHF processor linearity correction requirements are for small or negligible white level differential gain correction, moderate black level correction and approximately 100 percent sync stretch. Klystrons are purposely operated close to saturated power for best efficiency. This produces substantial sync compression to be overcome by operation of the sync stretch circuits.

FET visual modulator. Fig. 30 is a simplified schematic of the IF visual modulator employing a dual-gate field-effect transistor. The operation is extremely simple. Video is applied to one gate and IF visual carrier to the other. The drain output is a double-sideband amplitude-modulated IF visual signal with a nearly linear modulation characteristic.

Active IF filter. An IF shaping filter follows the visual modulator. It serves two functions. First, it restricts the bandwidth and establishes the initial vestigial sideband response characteristic. The final steep cutoff at the edge of the channel is established by a high-level filterplexer following the klystron.

The second function of the active filter is to provide a trim adjustment for the exciter or overall transmitter frequency response. The exciter normally is adjusted for flat frequency response within a tolerance of 0.25 dB across the channel.

The active filter has overall unity gain and 50 ohm input and output impedances.

Following the active filter, the signal is amplified to a level of 2 w, then padded to 1 w at the input to the visual upconverter.

Visual and aural upconverters. A parametric upconverter employing a microwave varactor diode was selected to convert from IF to final frequency because it is extremely linear over a wide range of power output, and it has a power gain of approximately 6 dB. This compares to a typical loss of 5 or 6 dB for a conventional diode mixer. Consequently, the desired final frequency visual output power of 4 w is achieved with an IF power of only 1 w.

No final frequency linear amplifiers are required to drive a 110 kw transmitter employing a pair of 5-cavity visual klystrons.

A simplified schematic of the upconverter is shown in Fig. 31. CR-1 is the mixer varactor diode. L_1 C_1 is the pump frequency tank circuit. The other circuit elements in the "IF cavity" and "pump cavity" blocks are required for: isolation between pump frequency and IF frequency; impedance match between the upconverter diode and the IF and Pump circuits; and control of intermodulation products.

Residual harmonics, or intermodulation products substantially removed from the desired output frequency, are eliminated by the three-pole bandpass filter in the "filter cavity" block of the diagram.

Another identical parametric upconverter handles the aural signal at an output level of 0.8 w. A common pump chain drives both upconverters.

Pump chain. Three temperature compensated crystal oscillators (TCXO) are employed, one each for: visual IF carrier generator, aural IF carrier generator; and UHF pump generator. Temperature compensation is precise from 0 degrees to +45 degrees C. No ovens or crystal heaters are required.

Of the three crystals, the stability requirement is most severe for the pump generator. The pump chain TCXO unit operates with a power output of 5 milliwatt on a frequency between 11 and 18 MHz, depending upon the channel assignment. Then, through a succession of wideband amplifiers and diode frequency multipliers, a power in excess of 20 watts is developed at the output of the pump chain.

The pump chain output is sampled, detected, and the resulting dc is used in an automatic level control circuit to maintain constant pump power at the splitter input.

The splitter is a 6 dB coupler which delivers 15 w pump power to the visual upconverter and 5 w pump power to the aural upconverter.

Solid-State IPA

The solid-state IPA is a 10-watt broadband fixed-tuned linear amplifier with a bandwidth approximately equal to the tuning range of a klystron. It is the "black box" device in the upper right-hand corner of the left cabinet in Fig. 25.

The output of the visual IPA connects to the klystron input through a ferrite circulator to isolate the IPA from load changes when tuning the klystron input circuit. No aural IPA is required.

Klystron Efficiency and the Anode Pulser

Power costs of operating UHF TV transmitters is substantial and there is strong incentive to reduce it.

In the first place, very high power transmitters are required to provide equivalent service to that provided by typical VHF-TV transmitters.

A second factor contributing to high power costs for UHF transmitters is the mediocre beam power efficiency of klystrons compared to the plate efficiency of VHF PA power tubes. The former have, until recently, operated at a sync peak level efficiency of approximately 35 percent compared to approximately 55 percent for VHF power tubes. Recently, high efficiency 30 kw klystrons have gone into service and this ameliorates the problem somewhat. These have a typical saturated efficiency of approximately 43.5 percent, corresponding to 40.5 percent sync level efficiency.

A third factor, more important than the second, is that the klystron input power is independent of average picture level (APL) whereas the tube input power is essentially directly proportional to signal output level. At mid-characteristic (average power = 1/4 sync power) a VHF tube would be twice as efficient as a UHF klystron on the basis of this factor alone.

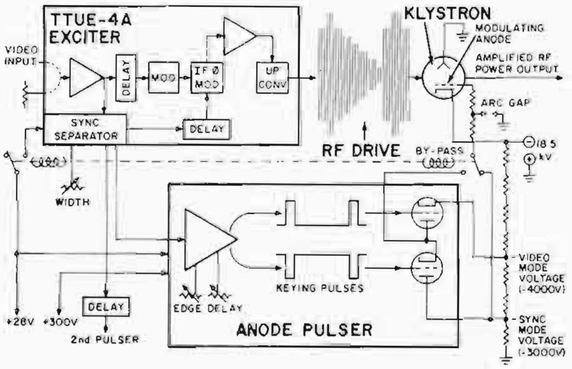

Fig. 32. Anode pulser system, functional diagram.

A system has been described (See Reference 3) for reducing the average power requirement of visual klystrons. In the new system, the klystrons are operated with increased RF drive and lower modulating anode voltage during all but the synchronizing interval. Then the modulating anode is pulsed to a higher voltage during the sync interval. A functional diagram of the anode pulser system to accomplish this is shown in Fig. 32.

The mod. anode pulser system produced an improvement in klystron beam power efficiency of 20 percent—further closing the gap between VHF power tube and klystron operating efficiencies.

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