SECTION 3: TRANSMITTERS

3.1

AM Transmitters

George W. Woodard, P.E.*
Radio Free Europe/Radio Liberty
Washington, DC

Broadcasting to the general public began with the process known as Amplitude Modulation near the start of the second decade of this century. Then, as now, the system of modulation chosen for transmission was heavily dependent on practical and economic aspects of receiver technology. The evidence is clear that amplitude detection was the only known practical method of signal demodulation when the ideas of radio communication began to be formulated in the late 1800's and early 1900's. It is apparent from some of the earliest technical writings on radio communication that a general mathematical knowledge of angular modulation, i.e.; frequency and phase modulation, did not exist until the mid 1920's.[1] These modes of transmission and the necessary receiver technology were not to be proven practical until long after amplitude modulation had become the standard of the radio communication and broadcast technological art. Amplitude modulation in broadcasting is today more accurately referred to as DSB-FC-AM, an acronym for Double SideBand-Full Carrier-Amplitude Modulation, however, the abbreviation "AM" has been generally accepted to describe the mode of transmission currently used throughout the world in the standard broadcast Amplitude Modulation band of 535 to 1605 kHz in North America, and slightly wider in other parts of the world. From almost the very beginning of broadcast technology, it was known that AM was not a very efficient mode of transmission, either in spectrum usage or in transmission of intelligence. The pioneers in AM broadcast transmitter and receiver engineering technology had a deep sense of responsibility toward developing this "marvelous medium" to enhance the lives of the general public with informative, educational, and entertaining programs. Some of the early communication technology pioneers in the U.S. and abroad had great vision for both the technical and programming aspects of public broadcasting.[2]

As time and technology progressed, engineers could no longer ignore the problems of transmission efficiency. When, in the 1930's, the telephone industry was planning a major switch from DSB-FC-AM in favor of Single SideBand-Suppressed Carrier (SSB-SC) for both its long distance wire and wireless services because of its higher transmission efficiency and channel capacity[3], the broadcast industry was necessarily committed to the continuance of conventional AM because of

*Formerly with Continental Electronics Manufacturing Company.

[1] "Notes on Modulation Theory". John R. Carson, Proc. IRE, Vol. 10, No. 1, June 1925, p. 57.

[2] Inaugural Address of Dr. Lee DeForest upon becoming the President of the I.R.E. Proc. IRE, VOL. 18, No. 7, Jul. 1930, p. 1121.

[3] "Production of Single Side-Band for Trans-Atlantic Radio Telephony". R. A. Heising, Proc. IRE, Vol. 13, No. 3, June 1925, p. 291.

the need for economic public receiver compatibility. Therefore, the broadcast engineers attention naturally turned to improvement of transmission standards such as fidelity, efficiency of transmission, reliability of transmission, and co-channel and adjacent channel interference, etc. Of these, efficiency and reliability of transmission have seen and continue to see the greatest improvement in the views of many transmitter designers and users. Transmission fidelity, always important, reached a plateau in the late 1940's which has not been significantly improved upon even in today's latest transmitter designs; though pre-transmitter program processing equipment and philosophy, which influence perceived reception fidelity, continue to change.

In the early 1980's, stereo transmission on the AM band became a widespread reality on the North American continent. This mode of transmission has become known and is referred to as AM-Stereo. Some AM broadcasters see AM-Stereo as a necessary means to compete with FM-Stereo which continues to gain increased public acceptance for all types of programming. AM-Stereo transmission was approved on a general basis by the FCC in 1981, but unlike FM-Stereo or Color Television, the FCC did not adopt a transmission standard for this potentially revolutionary development in the standard broadcast band. Instead the FCC allowed any commercially developed AM-Stereo transmission system, which will pass FCC type approval requirements regarding mono compatibility and interference, to compete for the broadcasters and consumers favor in order to establish a de-facto standard for AM-Stereo transmission. Some receiver manufacturers have decided to build mass production receivers for only one of the various proposed systems based on their public and private study of the systems merits and market potential. Other receiver manufacturers have decided to build mass production receivers which will automatically and/or manually decode any of the several proposed systems. Which receiver marketing philosophy will prevail and what is in the best long term interest of the listening AM public are yet unknown. Several AM-Stereo systems are presented in the chapter following this one.

AMPLITUDE MODULATION THEORY

Amplitude modulation of a radio signal occurs in at least two basic forms, coded (digital) and linear (analog). The first amplitude modulation processes for long distance communication involved on/off keying of a radio carrier wave. The pattern or "code" of the on/off keying process determined the content of the information being transmitted. Linear or quasi-linear undulations in the amplitude of the carrier wave, however, are normally used to transmit the analog information present in speech and music. The radio carrier wave signal onto which the analog amplitude variations are to be impressed is expressed as

$$e(t) = A \times Ec \times \mathrm{Cos}(\omega ct) \qquad [1]$$

Where:

$e(t)$ = instantaneous amplitude of carrier wave as a function of time (t)
A = a factor of amplitude modulation of the carrier wave
ωc = angular frequency of carrier wave (radians/second)
Ec = peak amplitude of carrier wave

If A is a constant, the peak amplitude of the carrier wave is therefore constant and no modulation exists. Periodic modulation of the carrier wave exists if the magnitude of A is caused to vary with respect to time as, for instance, a sinusoidal wave, i.e.;

$$A = 1 + (Em/Ec) \times \mathrm{Cos}(\omega mt) \qquad [2]$$

Where:

Em/Ec is the ratio of modulation amplitude to carrier amplitude

therefore;

$$e(t) = Ec \times (1 + (Em/Ec) \times \mathrm{Cos}(\omega mt)) \times \mathrm{Cos}(\omega ct) \qquad [3]$$

this is the well known basic equation for periodic amplitude modulation and when all multiplications and a simple trigonometric identity are performed the result is

$$e(t) = Ec \times \mathrm{Cos}(\omega ct) + (M/2) \times \mathrm{Cos}(\omega ct + \omega mt) + (M/2) \times \mathrm{Cos}(\omega ct - \omega mt) \qquad [4]$$

Where:

M = the amplitude modulation factor Em/Ec

Equation [4] can be represented graphically three familiar ways; in the time domain representation as shown in Fig. 1, in the frequency domain as shown in Fig. 2, and as relative vectors as shown in Fig. 3. The graphical representations shown are for a single tone modulation index (M) of 0.7, i.e.; the peak modulating voltage is 70% of the peak carrier wave voltage ($Em/Ec = 0.7$). Fig. 2 shows the occupied bandwidth of an AM signal with single tone modulation. From this Fig.

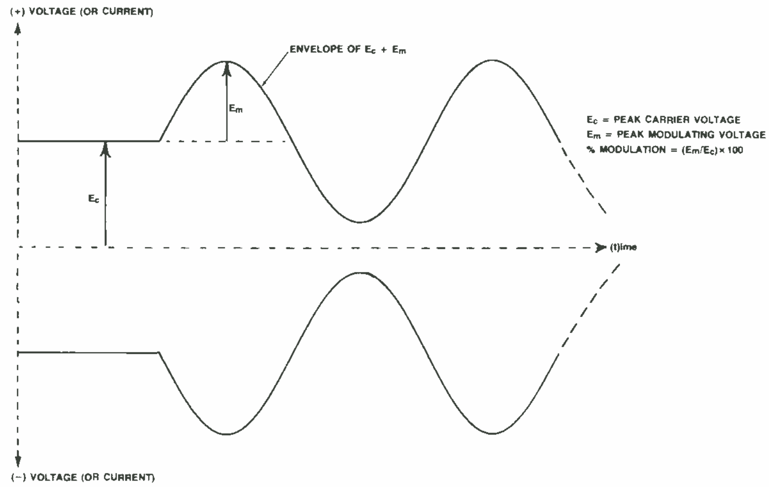


Fig. 1. Time domain representation of a carrier wave signal amplitude modulated by a sinusoidal audio signal to a peak modulation depth of 70%.

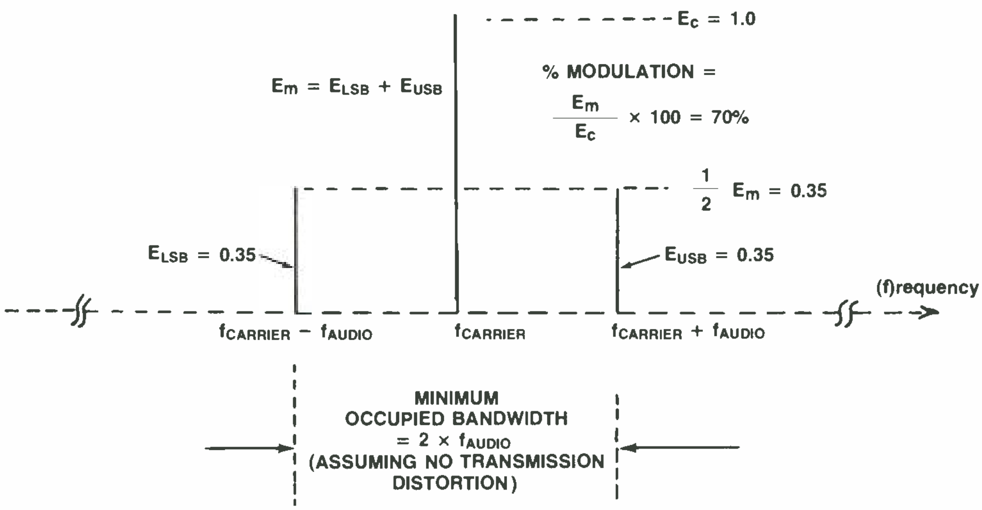


Fig. 2. Frequency domain representation of an amplitude modulated signal showing the carrier wave signal and two resultant modulation sidebands at 70% modulation.

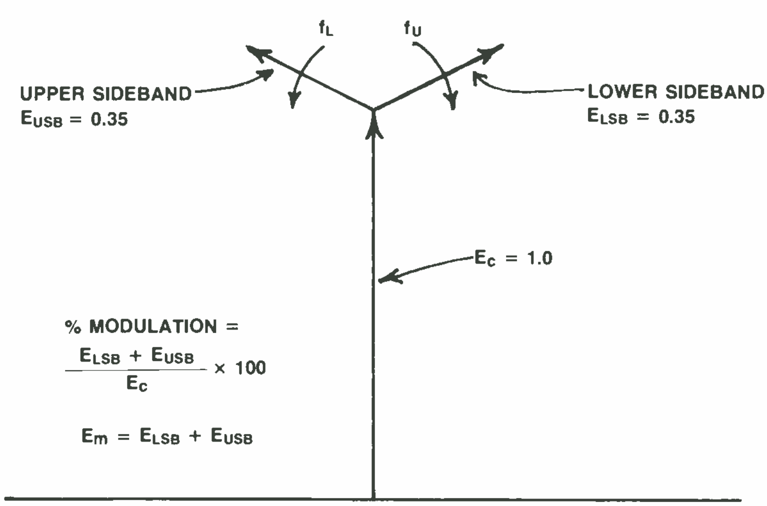


Fig. 3. Vector representation of an amplitude modulation signal showing the carrier wave signal and two resultant modulation sidebands at 70% modulation.

and its defining equation [4], it is clear that the bandwidth of an AM signal is equal to twice the highest modulating frequency if no system distortion is present. High quality music reproductions include frequency components as high as 15 kHz or higher and therefore the required theoretical bandwidth of a DSB-FC-AM signal capable of high quality music reproduction is at least 30 kHz. Harmonic and inter-modulation system distortion have the effect of widening the effective occupied bandwidth of the system. However, most modern transmitters have sufficiently low distortion characteristics that bandwidth stretching is not normally a significant problem. The occupied bandwidth characteristics of an AM broadcast transmitter are discussed in more detail in the section on "Factory Tests."

RADIO FREQUENCY POWER AMPLIFIERS

A common AM transmitter major system component is the power amplifier circuit. High power amplifiers that produce 0.25 to 50 kilowatts (kW) of carrier power are common for AM broadcast transmitters in North America. Carrier power levels up to one megawatt and higher are common in other parts of the world for medium wave broadcasting. Transmitters delivering these high power levels should be designed for highest operating efficiency within the circuit constraints of the particular manufacturer and/or modulation system used. The most common amplifier used to meet the demands of high output power and high efficiency is the vacuum tube class 'C' amplifier. Class 'B' RF power amplifiers, very popular in early transmitters, are still used occasionally at lower power levels and for driver stages of final class 'C' or 'D' stages. Solid state class 'D' amplifiers up to 5 kilowatts are becoming more common as driver stages for final vacuum tube amplifiers and for the final power amplifier modules in transmitters of up to 10 kW carrier power. As stated in the introduction, the most common major concerns for both manufacturers and users of modern AM broadcast transmitters is operating reliability and efficiency, hence operating cost. To achieve high overall operating efficiency, the stages which consume the most power, the modulator and/or the final RF power amplifier stages must be designed for highest possible operating efficiency.

The basic tuned anode vacuum tube amplifier is described in graphical form in Fig. 4. The vacuum tube can be either a triode, tetrode, or pentode. Tetrode final amplifiers are most common in modern high power transmitter designs. The RF excitation voltage is supplied to the grid of the power amplifier tube and the ratio of dc grid bias voltage-to-peak RF excitation voltage, shown sinusoidal in Fig. 4, determines the conduction angle of anode current, given as

$$\theta_c = 2 \times Arc\ \mathrm{Cos}(E_{cc}/(E_g - E_{cc})) \qquad [5]$$

where the exciting grid signal is sinusoidal as shown in Fig. 4a.

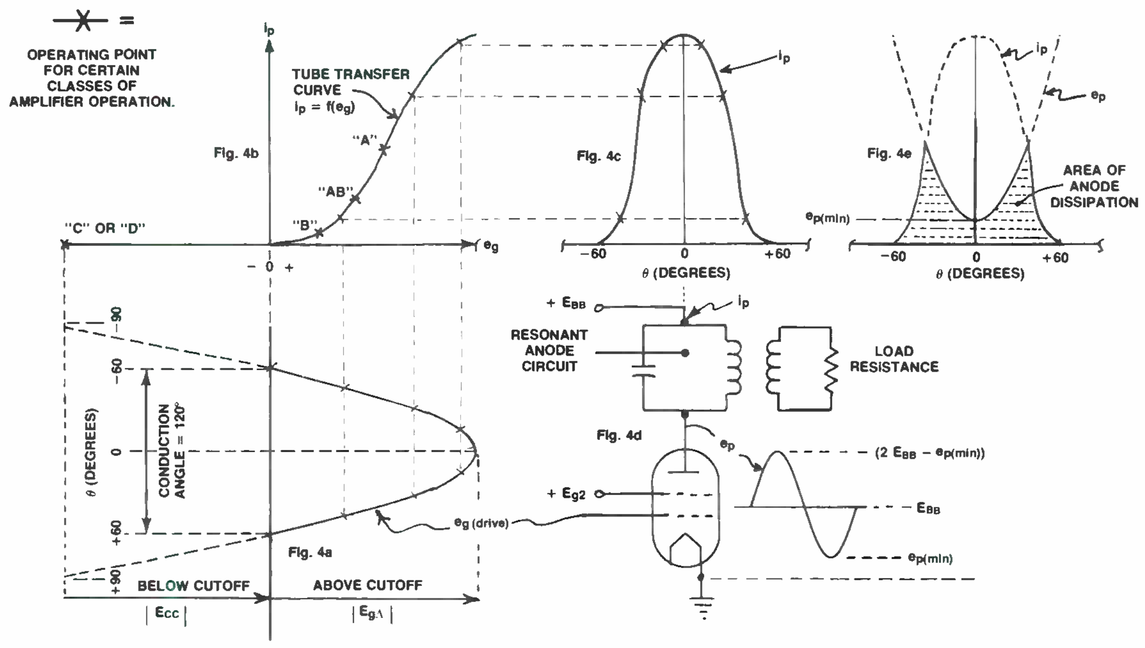


Fig. 4. Classical vacuum tube Class "C" Amplifier with sinusodial grid drive, 120° anode current conduction, and resonant anode load.

The shape of the anode current pulse is determined by the vacuum tube transfer characteristics and input waveshape. The pulse of current thus generated, Fig. 4c, is supplied by the dc power supply, E_{BB}, and passed through the resonant anode tank circuit shown in Fig. 4d. The resonant anode tank circuit is assumed to have sufficient operating Q to force anode voltage, ep, to be essentially sinusoidal and of the same periodic frequency as the RF excitation voltage and resultant anode current pulse. The instantaneous anode dissipation, shown in Fig 4e, is the product of instantaneous tube anode voltage drop and anode current. The tube transfer characteristic is a variable dependent upon many tube factors as well as maximum drive signal, Eg. The exact shape and magnitude of the current waveform is normally obtained from a load-line plot on constant current characteristic tube curves supplied by the tube manufacturer. The resonant anode load impedance is chosen and adjusted to allow ep(min) to be as low as possible without causing excessive screen grid (in the tetrode case) or control grid dissipation. Some manufacturers increase anode efficiency beyond the limits for typical Class 'C' amplifiers by using a circuit employing a third harmonic resonator between the output anode connection and the fundamental resonant circuit. This has the effect of squaring up the anode voltage waveform (ep) thus causing the integral of the ep × ip product, or anode dissipation, to be smaller; resulting in lower anode dissipation for a given RF power output. An amplifier employing the third harmonic anode trap is commonly referred to as class 'C-D'; suggesting an efficiency rating somewhere between conventional class 'C' operation (nominal 120 degree conduction angle) and true class 'D' operation with rectangular anode or collector voltage waveforms. Anode efficiencies can be increased typically to values of 90 percent for transmitters up to approximately 10 kW carrier power and approximately 85 percent for transmitters higher than 10 kW carrier power by using the third harmonic trap technique. Table 1 shows a comparison of anode efficiency for six classes of high power tuned RF amplifiers.

DISCUSSION OF BASIC AM SYSTEMS

High Level Anode Modulation

The first practical method of generating the amplitude modulation signal was Heising constant current modulation[4, 5], a method of applying audio modulation to the anode supply voltage of a class 'C' RF amplifier. This general class of modulation has since been known as high level anode (or plate) modulation. The Heising modulator was used at least as early as 1920 and was usually used to modulate a low power RF amplifier or master oscillator stage which was followed by several linear amplifier stages until the desired power level was attained. In some cases the Heising modulator was used to modulate the final RF amplifier stage of lower power transmitters. The Heising shunt modulator operated in the class 'A' mode and therefore was low in operating efficiency. The early linear amplifiers were tuned class 'B' amplifiers operating with carrier level anode efficiencies of 30% maximum. The Heising and similar systems of audio amplification were also used to modulate the grid bias level of RF amplifier stages in order to obtain the AM signal to be used for further linear amplification. Heising constant current anode modulation was very popular in military and aviation radio sets used through the end of World War II.

Class 'B' High Level Anode Modulation

Historically the most popular method of applying the audio modulating voltage to the anode circuit of a class 'C' RF power amplifier was by a high power push-pull class 'B' audio amplifier. This type of modulation was first used to improve the operating efficiency and to increase the output power of AM broadcast transmitters. Class 'B' push-pull audio amplification was first used

[4] "The Equivalent Circuit of the Vacuum Tube Modulator". John R. Carson, Proc. IRE, Vol. 9, No. 4, Aug. 1921, p. 243.

[5] "Modulation in Radio Telephony". R.A. Heising, Proc. IRE, Vol. 9, No. 3, June 1921, p. 305.

TABLE 1. Comparison of Tuned RF Amplifier Anode Efficiencies

Amplifier Class	Conduction Angle (degrees)	Anode Efficiency (%)	Defined Conditions of Operation
A	360	30	$E_{b(min)} = 0.10 \times E_{BB}$
A-B	240	60	$E_{b(min)} = 0.10 \times E_{BB}$
B	200	67	$E_{b(min)} = 0.10 \times E_{BB}$
C	120	84	$E_{b(min)} = 0.05 \times E_{BB}$
C-D	120	90	$E_{b(min)} = 0.05 \times E_{BB}$
D	120	95	$E_{b(min)} = 0.05 \times E_{BB}$

to improve distortion and output power of telephone transmission amplifiers. The invention was soon recognized by broadcast engineers and applied to high level anode modulation. With the final RF power amplifier operating at approximately 80% anode efficiency and the class 'B' audio modulator total static currents approximately one-tenth that of an equivalent Heising modulator, total anode efficiencies at carrier level rose to approximately 72% compared to 37% for the Heising system and 30% for conventional linear amplification. A simplified drawing of a typical high level class 'B' anode modulation system is shown in Fig. 5. The vacuum tubes shown in Fig. 5 may be either triodes, tetrodes, or pentodes. The output circuit of the class-B modulator shows the output modulation transformer (MT), an audio coupling capacitor (C), and a dc shunt feed inductor (L).

This arrangement was used in all high level class-B high power broadcast transmitters until about 1960 because of a transformer design constraint that would not economically allow unbalanced direct currents to magnetize the transformer core material without poor low frequency distortion. Advanced technology core materials and careful magnetic transformer design allowed elimination of the coupling capacitor and the dc feed shunt inductor, first in some 100 kW European transmitter designs in the early 1960's, and in an American shortwave transmitter design in the later 1960's. Many of the more advanced modern AM broadcast transmitter designs still using high level class-B anode modulation have eliminated the extra C and L components from the modulator output circuitry, and the dc current to the modulated RF amplifier anode flows directly through the secondary of the output modulation transformer. Elimination of these C and L components is necessary for optimum operation of modern AM stations. With the extra C and L components, the modulator output is effectively a three pole high-pass filter which causes low frequency transient distortion to be generated when driven with the complex waveforms that are produced by many modern and popular program audio processors. Eliminating the C and second L component, causes the output modulator circuit to become a single pole high-pass filter, greatly reducing low frequency transient distortion.

Another problem with transformer coupled high level class-B anode modulation is with high frequency audio transient distortion. Stray internal winding capacitances and leakage inductances form multi-pole low pass filtering at the high frequency end of the audio spectrum. This equivalent multi-pole low pass filter generates transient overshoot distortion when driven by the same type of processed complex program waveforms mentioned above. Transient overshoot up to 12% is typical for squarewave modulator input signals and results in a required modulation level reduction of the same 12% in order to maintain peak modulation levels within FCC allowed limits. This high frequency transient overshoot distortion can be effectively minimized by filtering the audio input to the transmitter with linear phase filters, resulting in somewhat lower high frequency audio response, and/or by careful control of the

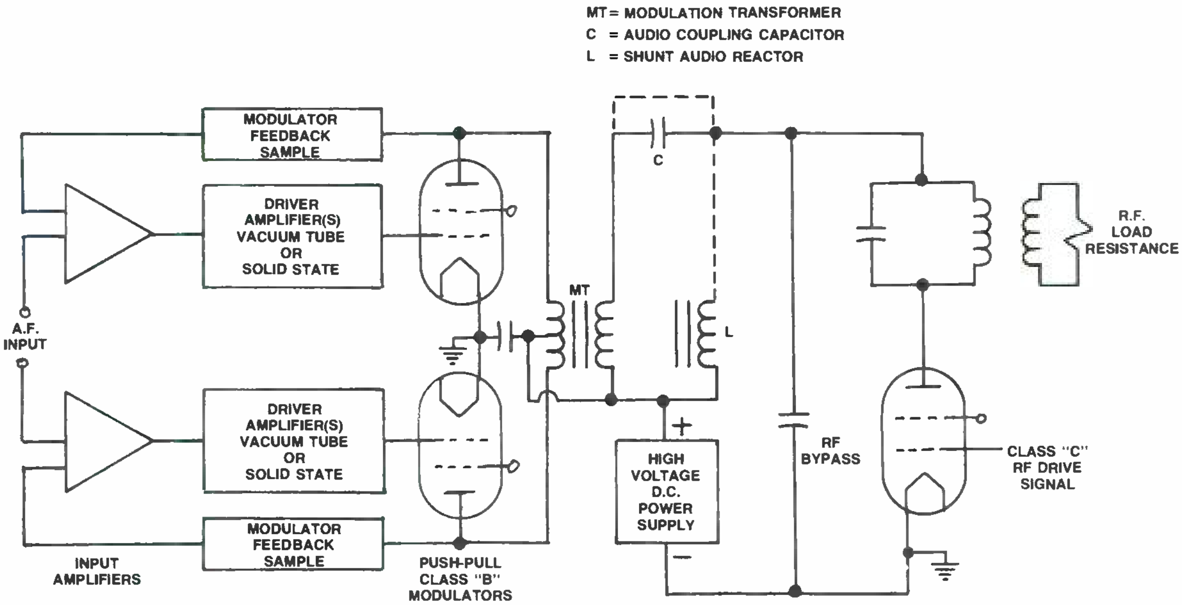


Fig. 5. High level anode modulation employing transformer coupled push-pull Class-B modulators.

modulation transformer equivalent circuit yielding more linear audio phase characteristics for the entire modulator circuitry. Balanced modulator negative feedback is used to reduce modulator non-linear distortion and noise. Negative feedback, however, usually worsens high audio frequency transient distortion characteristics.

Pulse Width High Level Anode Modulation

Pulse width modulation (PWM) of the dc anode voltage of a class-C RF amplifier was first used in commercial high power broadcast transmitters in Europe in the early 1960's. It was the first commercially successful attempt to significantly improve upon the efficiency of the popular high level class-B modulation system by applying and improving basic PWM concepts that were described decades earlier.[6] Since this first success, pulse width modulation has become a preferred method of high level anode modulation by many broadcast engineers, and is employed in several broadcast transmitter designs by several manufacturers world wide. The basic pulse width modulation system and two ingenious improvements to the basic system are shown in Fig. 6. The circuit in Fig. 6a graphically describes the basic principle of PWM. An inherent practical deficiency in the basic concept, Fig. 6(a), is caused by the relatively high shunt circuit capacitance of the modulator tube filament transformer plus stray capacitances. Though special low-capacitance isolation transformers can be used to supply modulator filament and auxiliary power to minimize capacitive switching losses, typical realizable values of capacitance can cause excessive switching losses and audio distortion in lower power transmitters.

For example, the switching losses at carrier level of a typical PW modulator for a 5 kW transmitter can be higher than the quiescent modulator losses of an equivalent power class-B modulator even when the stray modulator tube capacitances are as low as 100 pF. The power lost per modulator switching cycle is

$$P_{modsw} = (CV^2/2) + P_{td} \qquad [5]$$

Where:

- C is the shunt modulator filament transformer plus stray capacitance to ground and;
- V is the pulse switching voltage to ground at the cathode of the tube.
- P_{td} is the saturated tube and diode losses during the respective on and off conduction states

Besides causing switching losses, the high stray capacitance to ground is also a cause of major modulator distortion at high negative modulation indices.[7] The circuit shown in Fig. 6b is one ingenious way to overcome the stray modulator capacitance problem. It is identical to the circuit in Fig. 6a, in principle, except that the high voltage pulse modulated wave is at a point in the circuit where the shunt capacitances to ground are inherently minimized. The circuit in Fig. 6c is basically the same as in 6b except the system ground has ingeniously been placed at another point in the circuit. PW Modulator anode efficiencies approach 92% in some higher power transmitter designs yielding a combined modulator and carrier tube anode efficiency of approximately 74% at all levels of modulation. An added efficiency advantage over high level class-B anode modulation is that a PWM transmitter may have only two high power vacuum tubes, one modulator and one final RF amplifier, thus eliminating the filament heating power of one large vacuum tube required in push-pull modulator designs. A significant disadvantage of PWM as described in Fig. 6b and 6c is that the cathode and grid circuits of the modulated RF amplifier are operated at high voltage levels off ground, adding complications to the circuitry in these areas that are avoided with the classical PWM circuit of Fig. 6a and conventional high level class-B anode modulation. Another major disadvantage of PWM in any form is that of transient distortion caused by the phase non-linearity of the multi-pole PWM filter, similar to that previously discussed for high level class-B anode modulation. The switching frequency is typically 70 kHz for most transmitters manufactured in North America. This frequency is chosen to ease compliance with FCC regulations which requires all spurious radiation more than 75 kHz removed from the carrier frequency to be 80 dB or more below the carrier level. Present FCC regulations require transmitter spurious output between 30 kHz and 75 kHz removed from the carrier to be only 35 dB below the carrier level. To meet these spurious requirements (most manufactures of PWM transmitters far exceed these requirements), a very steep cutoff lowpass filter is required at the output of the PW modulator. Most manufacturers of PWM transmitters make a valiant attempt to maintain linear phase characteristics in this filter to as high an audio frequency as possible, but the laws of nature prevent required linear phase characteristics to the highest third audio harmonic to be achieved within the major attenuation constraints noted above. As

[6] "Transmission System". R. A. Heising, U.S. Patent No. 1,655,543, Jan. 1928.

[7] "Distortion in Pulse Duration Modulation". Ernest R. Kretzmer, Proc. IRE, Vol. 35, No. 11, Nov. 1947, p. 1230.

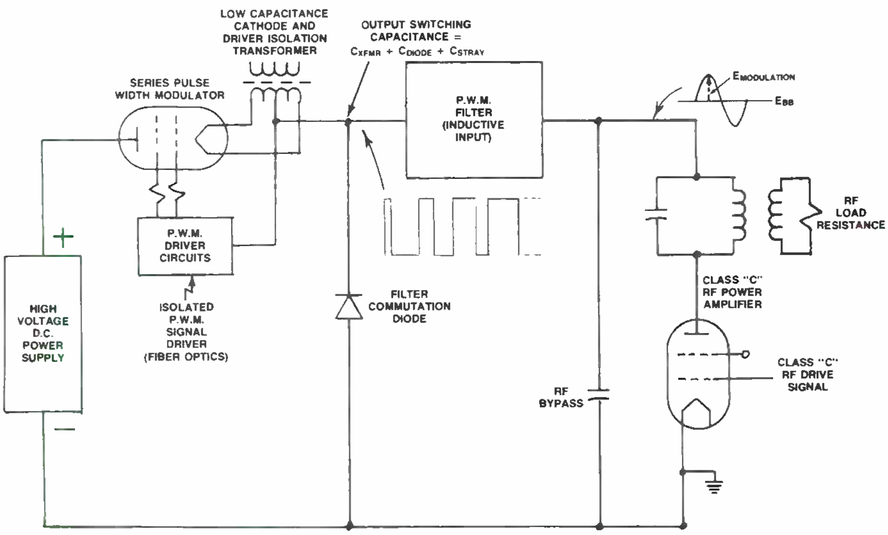


Fig. 6a. Basic classical high level anode pulse width modulation of a vacuum tube Class "C" amplifier.

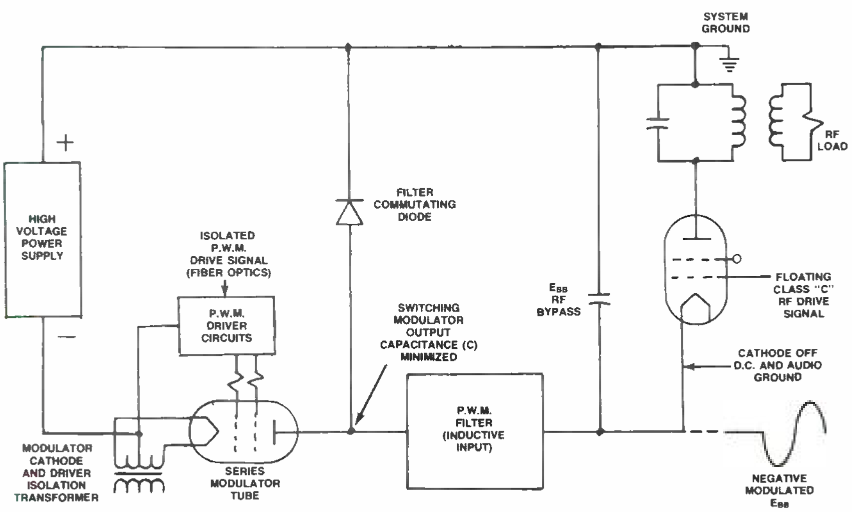


Fig. 6b. Collins Radio/Continental Electronics patented modification to basic high level PWM System to minimize modulator losses by minimizing switching modulator output capacitance.

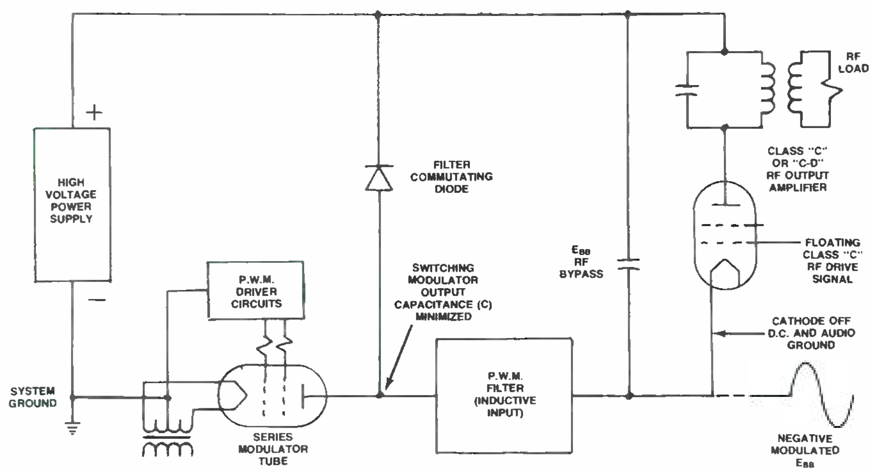


Fig. 6c. Harris Corp. patented modification to basic high level PWM System to minimize modulator losses by minimizing switching modulator output capacitance.

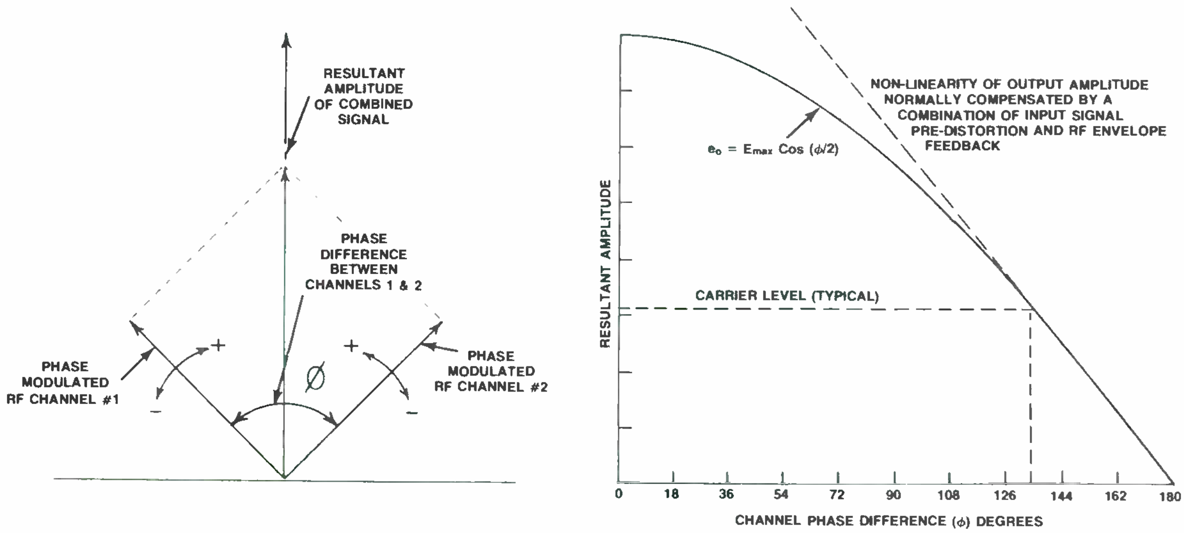


Fig. 7. Principle of Chireix "outphasing modulation".

a result, transient response to a square wave input can result in peak overshoot of approximately 6 %. As with Class "B" modulation, this overshoot can be effectively eliminated by the use of linear phase (Bessel) filtering of the input modulation signal at the expense of a small reduction in modulation frequency response capability.

Chireix "Outphasing" Modulation

Outphasing modulation was originally described in the literature by its inventor Henry Chireix in 1935.[8] It is a unique and ingenious method of obtaining the AM signal by use of counter-phase modulation and vector addition of two separate radio frequency signals. It was marketed for many years by RCA under the trade name "Ampliphase" and many of those transmitters and some of European manufacture are on the air still today at power levels of 100 kW and higher. This system of modulation is described graphically in Fig. 7. Two RF signals are derived from a common excitation source and then split into two separate channels. Each channel is shifted in phase, one positive and the other negative. The two channels are then each phase modulated by the modulating signal, again, in opposing polarity. The two channels are amplified and then recombined in a vector additive network which has the effect of producing the desired amplitude modulation. The main advantage, as with all systems previously discussed, is in operating efficiency. The two independent channels contain only phase modulated RF signals and therefore each can be amplified to the desired power levels in high efficiency Class 'C' or 'D' amplifiers. The actual modulation process takes place both at low level, in the phase modulators, and at high level, in a passive output network combiner.

There are two major disadvantages of this system of modulation. First, the efficiency of the output power amplifiers is not quite as high as the simple description above would imply, due to the fact that at all instantaneous levels of modulation, except one, the anode circuits must work into a reactive load. Secondly, output carrier power setting is sensitive to tuning of any stage anywhere down the chain; operators of early versions of this type of equipment soon learned of the consequences of "trimming" a tuning control of some lower power stage. Later designs used broadband amplifying stages for the lower levels to circumvent the problem of tuning sensitivity of the lower power stages.

Another major disadvantage of the Chireix system in older transmitters was the generation of incidental phase modulation (IPM) in excessive amounts when the two low level phase modulators were not well balanced in opposing PM characteristics. It was common for such an unbalance to exist producing as much as 12 to 18 degrees of peak IPM at either 100% modulation crest. Modern solid state circuitry for the phase modulators using digital phase modulation techniques would completely remove this disadvantage were the system still being commercially produced. The outphasing system of amplitude modulation ceased to be produced commercially when RCA discontinued its AM transmitter manufacturing in the mid-1970's.

Doherty High Efficiency Linear Amplifier

The Doherty high efficiency linear amplifier was first described in the technical literature in

[8] "High Power Outphasing Modulation". Henry Chireix, Proc. IRE, Vol. 23, No. 11 Nov. 1935, p. 1370.

1936 by its inventor, W.H. Doherty.[9] So contrary were the term "linear" and "high efficiency", in the context of amplitude modulated waves, that many engineers in broadcasting refused to accept the concept as workable, similar to the reaction Armstrong received when he proposed that frequency modulation was a practical mode of radio transmission. Nevertheless, the Doherty high efficiency linear amplification system was soon proven to work by 1938[10], and has been used at power levels up to 500 kW carrier power in both the original and in the patented Weldon[11] modified form in many installations throughout the world on the medium wave broadcast and international short-wave broadcast bands, as well as the long wave broadcast band in Europe. Its implementation was the result of true inventive genius, using one or more known basic scientific principles to create a totally new and necessary product. The Doherty linear amplifier is described graphically in Fig. 8. As with conventional linear amplifiers, the AM signal is generated at low levels and applied to the input of the final amplifier stage. The Doherty system employs two output amplifier stages, one defined as the carrier amplifier and the second as the peak amplifier.

The outputs of the two stages are combined in phase at the anode of the peak amplifier tube. At carrier level the carrier tube is operated as a nearly saturated Class 'B' amplifier and thus delivers almost all of the carrier power at Class 'B' efficiencies; i.e., approximately 70% anode efficiency. The peak tube at carrier condition is biased and driven just above cutoff and therefore supplies a small amount (approx. 2 to 6%) of carrier power. The anodes of the two tubes are connected together through a 90 degree impedance inverting RF network. As the modulated signal increases in the positive direction to both peak and carrier tubes, the current supplied to the output load by the peak tube increases. The saturated voltage drop at the anode of the carrier tube remains constant over the entire positive modulation half-cycle, thus causing the current at the output of the inter-anode 90 degree network also to be constant during the same positive modulation half-cycle. The rising current from the peak tube anode has the effect of raising the impedance being presented to the inter-anode network. Since the current from the network is constant, the net effect is an increase in output power from the carrier tube; i.e., $I^2 \times R_1$ increases because R_1 increases. At the 100% positive modulation crest, both tubes are producing exactly twice carrier power to the load, satisfying the requirement that Peak Envelope Power (PEP) = $4 \times P_{\text{carrier}}$. During the negative half-cycle of modulation, the peak tube is cut-off and the carrier tube behaves as a normal linear amplifier, allowing the envelope power output to drop linearly to zero at the 100% negative modulation crest. The anode efficiency of the Doherty high-efficiency linear amplifier at carrier level is more than twice the efficiency of conventional AM class 'B' linear amplifiers.

The Doherty linear amplifier also has two other important advantages for high power broadcast transmitters. First, and most important, the peak anode voltage at either tube is only about one-half that required for an equivalent carrier power high level PWM or class 'B' anode modulated transmitter thus allowing reliability and usable tube life to increase significantly. Secondly, no large modulation transformer or special filtering components are used in the final amplifier stages to cause transient overshoot distortion as previously discussed for Class 'B' anode modulation or pulse-width anode modulation. The main problems of the Doherty linear amplifier are non-linear distortion and an increase in the complexity of tuning. The major sources of non-linear distortion are the non-linearity of the carrier tube at or near the 100% negative modulation crest and the non-linearity of the peak tube at or near carrier level when it is just beginning to conduct. Both sources of distortion are effectively reduced by use of moderate amounts of overall envelope feedback. The tuning complexity problem is usually overcome by experienced and trained operators and simplified tuning procedures aided by built in test equipment used in modern designs since approximately 1948.

High Efficiency Screen/ Impedance Modulation

In 1938, Terman and Woodyard[12] described a modification to the basic Doherty high efficiency linear amplification system previously described. In the new system, the grid bias level of two tubes operating class 'C' is varied at the audio modulation rate thus creating a higher efficiency system of amplitude modulation rather than amplification while still using the impedance inverting

[9] "A New High Efficiency Power Amplifier for Modulated Waves". W. H. Doherty, Proc. IRE, Vol. 24, No. 9, Sept. 1936, p. 1163.

[10] "A 50-Kilowatt Station Utilizing the Doherty Amplifier and Designed for Expansion to 500-Kilowatts". W.H. Doherty and O.W. Towner, Proc. IRE, Vol. 27, No. 9, Sept. 1939, p. 531.

[11] "Amplifiers". J. O. Weldon, U.S. Patent No. 2,836,665, May 1958

[12] "A High Efficiency Grid-Modulated Amplifier". Proc. IRE, Vol. 26, No. 8, Aug. 1938, p. 929. F.E. Terman and John R. Woodyard.

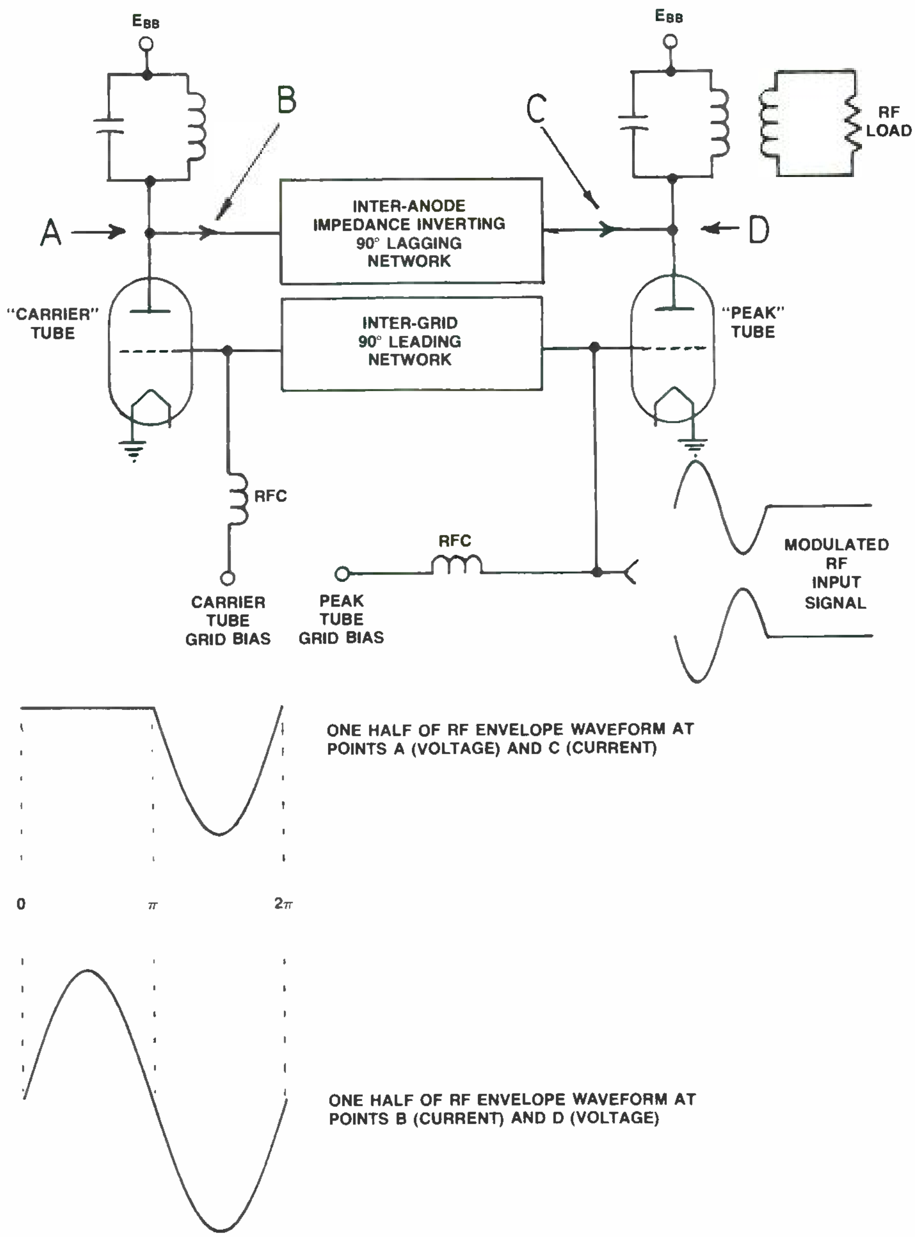


Fig. 8. Doherty high efficiency linear amplifier for modulated waves.

properties of the inter-anode network described by Doherty. The Terman-Woodyard system of modulation, however, was not extensively used in high power commercially successful transmitter designs.

High efficiency screen/impedance modulation[13] is similar to the Terman-Woodyard modulation system except that the audio modulating signal is applied to the screen grids of two tetrode vacuum tubes operating as class 'C' carrier and peak amplifiers. Invented by J.B. Sainton in 1965, the screen/impedance modulation system exhibits significant improvement to the Terman-Woodyard scheme due to the fact that RF excitation voltages and audio modulating voltages are isolated from each other thereby eliminating a troublesome source of tuning vs. modulation interaction. The screen/impedance modulation system is shown and described graphically in Fig. 9. The peak and carrier tubes are biased and driven in quadrature as Class 'C' amplifiers from the continuous wave RF drive source. At carrier level, the screen voltage of the carrier tube is adjusted so that the carrier tube is near anode saturation and delivering approximately 96% of the carrier power; while the screen voltage of the

[13] "High Efficiency Amplifier and Push-pull Modulator". J. B. Sainton, U.S. Patent No. 3,314,024, April 1967

peak tube is adjusted so that the peak tube is just into conduction and supplying the remaining approximate 4% of carrier power. The combined anode efficiency at carrier level is better than 77% as shown in equation [6].

$$n_{at} = 1/\ (pc/n_{ac}) + (pp/n_{ap}) \qquad [6]$$

$$n_{at} = 1/\ (0.96/0.8) + (0.04/0.40) = 0.77$$

Where:

- pc = percent carrier power supplied by carrier tube (as a decimal)
- pp = percent carrier power supplied by peak tube (as a decimal)
- n_{ac} = carrier tube anode efficiency at carrier level (as a decimal)
- n_{ap} = peak tube anode efficiency at carrier level (as a decimal)
- n_{at} = total anode efficiency at carrier level (as a decimal)

The modulation of the RF carrier wave occurs when the screen voltage of the peak tube begins to rise during the positive modulation half-cycle, thus causing the peak tube to supply more RF current to the output load. This increase of current into the output network causes the resistance seen by the inter-anode network to increase and, due to the impedance inverting characteristic of

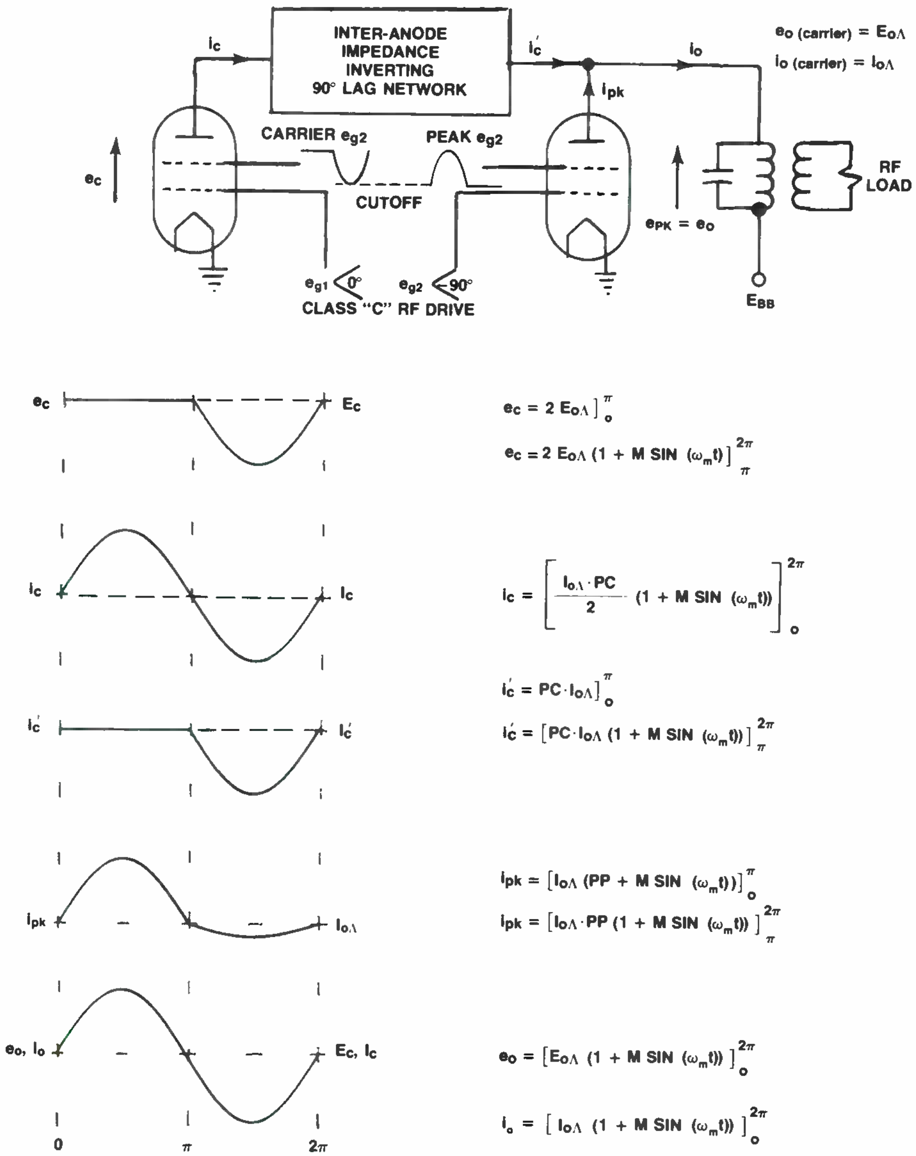


Fig. 9. Principles of Sainton high efficiency screen/impedance modulation.

the 90° inter-anode network, causes a proportional decrease in the load impedance presented to the carrier tube anode. The carrier tube resonant anode voltage drop is fully saturated over the entire positive modulation half-cycle and is therefore effectively a constant voltage source. The power output of the carrier tube thus increases during the positive modulation half-cycle due to the modulated decreasing impedance at its anode until both peak and carrier tubes deliver twice carrier power at the 100% positive modulation crest. During the negative modulation half-cycle, the peak tube is held out of conduction while the carrier tube output voltage decreases linearly to zero output at the 100% negative modulation crest. Simple alterations to the inter-anode 90 degree network characteristic impedance gives the screen/impedance modulation system the capability of full FCC allowed positive modulation up to 125% peak without added distortion. The advantages of screen/impedance modulation are the same as mentioned for the Doherty linear amplifier except that screen/impedance modulation has higher efficiency at all depths of modulation and, is less critical to misadjustment of RF amplifier tuning. Screen/impedance modulation has been successfully used in medium wave transmitters throughout the world at the two megawatt carrier power level and up to 250 kilowatts in automatically tuned transmitters used for international shortwave broadcasting.

In 1980, the British Broadcasting Corporation completed a four year study of known amplitude modulation systems, and selected the screen/impedance modulation system for their upgrade of 50 kW AM broadcast transmitters used in the BBC domestic and worldwide network.

SOLID STATE BROADCAST TRANSMITTERS

Developments in solid state technology in the last few years has allowed solid state devices to be used in broadcast transmitters at high power RF levels. While many engineers maintain that vacuum tubes will continue to have definite operating cost and reliability advantages over solid state devices at most higher RF power levels for sometime because of potentially destructive energy in lightning strikes, power line surges, and other surprises of nature; the facts are clear that present solid state device technology and circuit developments in some military and commercial RF applications up to two megawatts and beyond have already established high power solid state RF technology as a force in future high power transmitter development. Fully solid-state and proven reliable AM broadcast transmitters up to the 10 kilowatt carrier power level are currently commercially available. It is generally believed that higher power levels are not far in the future. The debate will probably continue for some time regarding solid state vs. tube reliability in regard to lightning strikes, power surges, ability to withstand abuse, maintenance cost, etc., but ultimately, most design engineers believe, totally solid state designs will be the standard designs of the future. The reason for this belief is the same reason that has led the AM transmitter industry from the Heising modulator of the 1920's to the best of the modern designs, the continuous attempt of broadcasters and equipment design engineers to improve efficiency, reliability, and total equipment operating cost.

Solid State Designs

Most solid state designs of AM broadcast transmitters currently available on the commercial market employ Class 'D' final RF power amplifiers yielding dc to RF efficiencies of approximately 95%. Pulse width modulation (PWM) of the final, and sometimes driver, RF amplifier supply voltage is employed yielding modulator efficiencies also of approximately 95%. The total dc to RF carrier efficiency is therefore approximately 90%, somewhat higher than normally achieved with tube type class 'C' or quasi-Class 'D' power amplifiers. Furthermore, there is no additional filament heating power required in solid state transmitters and therefore the transmitter auxiliary power consumption is further minimized yielding overall transmitter efficiencies greater than 70% at the 10 kW carrier power level, approximately 10% higher than most tube type transmitters of the same power level. Most solid state designs employ modularization of major system components to ease corrective maintenance procedures as well as providing manufacturing economies.

TRANSMITTER CIRCUITRY

Detailed transmitter circuit design is as varied as the individual designers. In a later section of this chapter will be brief and general discussions of some current production transmitter models which are representative of a particular power level, modulation system, or other unique feature. There is, however, some basic circuitry which is common to all transmitter types and models which will be briefly discussed in this section.

Carrier Frequency Generator/Exciter

The stable frequency source for all transmitters manufactured since about 1930 has been the quartz crystal oscillator. The quartz crystals used in older model transmitters were large cuts of natural quartz, vacuum sealed in glass envelopes, similar to small power vacuum tube envelopes, and occasionally mounted in temperature controlled ovens to obtain the required FCC carrier frequency stability. It is more common in modern designs for the quartz crystals to be enclosed in small hermetic sealed metal cans, made popular and proven reliable by the military and commercial communications equipment industry, and without special temperature controlled circuitry. Modern quartz crystal manufacturing technology and the use of proven solid state crystal oscillator designs allows these types of small metal sealed units to adequately maintain the FCC's current requirement of ±20 Hz carrier frequency tolerance. Stability of the quartz oscillator circuits is normally adequate over the full lifetime of the equipment. Should frequency adjustment ever be required to bring a unit back inside the FCC limits, mechanically stable adjustment components, usually a glass or ceramic piston type of capacitor, are provided for use by a qualified station engineer using certifiably calibrated frequency measuring equipment. (Ref: FCC Rules and Regulations Volume III, October 1982, Part 73.1540 and 73.1545). Exciters for all proposed AM-Stereo systems provide the carrier frequency excitation for the transmitter as well as the stereo generating circuitry. Because of manufacturing advantages, some manufacturers of AM-Stereo exciters generate the desired carrier signal with frequency synthesizer techniques. This method of generating the basic carrier frequency is generally equal to or better than the discrete quartz oscillator method with regard to frequency tolerance but can produce higher phase modulation noise if improperly designed or adjusted.

RF Power Amplifier

Practically all modern AM broadcast transmitters employ Class 'C' amplification in the final RF power amplifier stage(s). Some employ designs using third harmonic trap circuitry yielding quasi-class 'D' operation for improved efficiency. Most solid state transmitters currently available in the North American market use Class 'D' RF final power amplification. Refer to paragraph 3 for a discussion of power amplifier theory.

RF Output Networks

The purpose of the RF output network is to match the impedance of the load; i.e., the common-point impedance of one or more antenna matching and combining networks, to the impedance required by the final RF power amplifier tube(s) or transistor(s) to produce the desired carrier and sideband power. The output network circuit also is designed to provide the attenuation characteristics necessary to meet the FCC's requirements for spurious and harmonic output. There are many techniques to accomplish these basic tasks. One simple and effective method is shown in Fig. 10 as an illustrative and typical example. The terminating load impedance for the network shown in Fig. 10 is defined by the Smith chart representation of Fig. 11, which is a typical common point impedance characteristic of a multi-tower AM broadcast directional antenna array.

Fig. 12 shows the impedance vs. frequency characteristics at the input of the network; i.e., at the anode of the final RF amplifier tube.

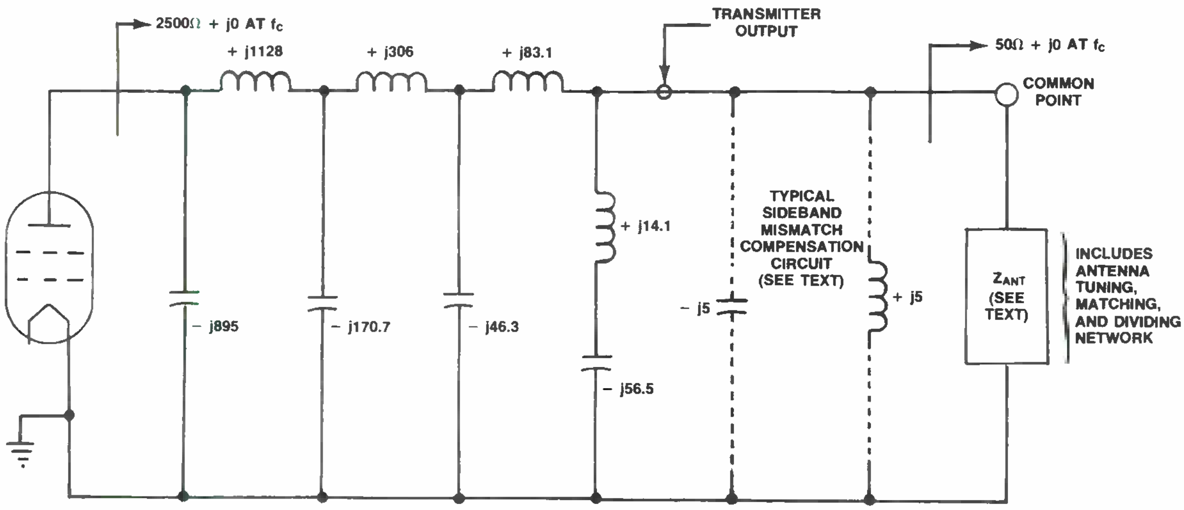


Fig. 10. Typical output matching network yielding 80dB or better harmonic attenuation and a compensated symmetrical sideband load at the anode(s) of the RF output amplifier.

IMPEDANCE COORDINATES—50-OHM CHARACTERISTIC IMPEDANCE

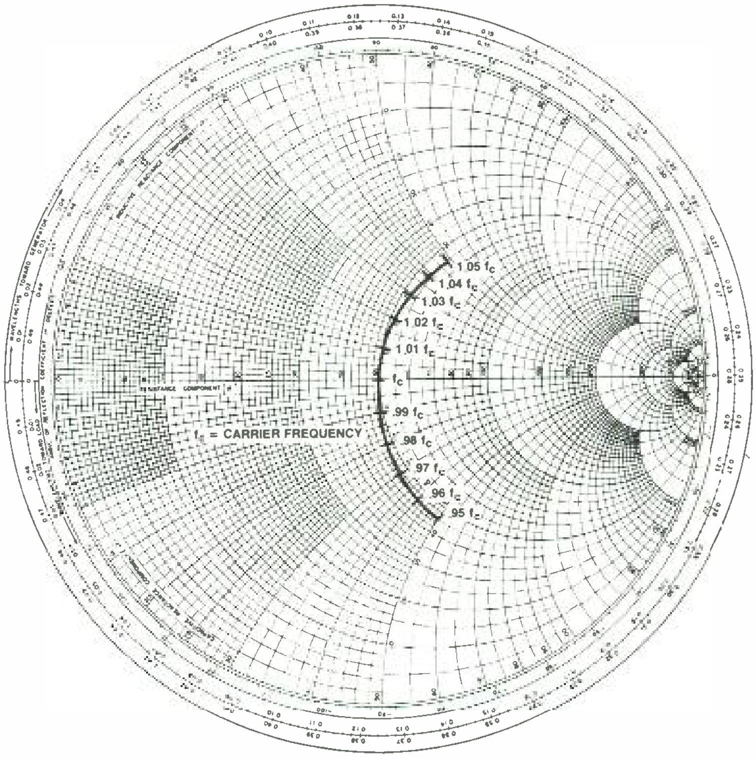


Fig. 11. Typical impedance vs. frequency characteristic of antenna common point.

IMPEDANCE OR ADMITTANCE COORDINATES

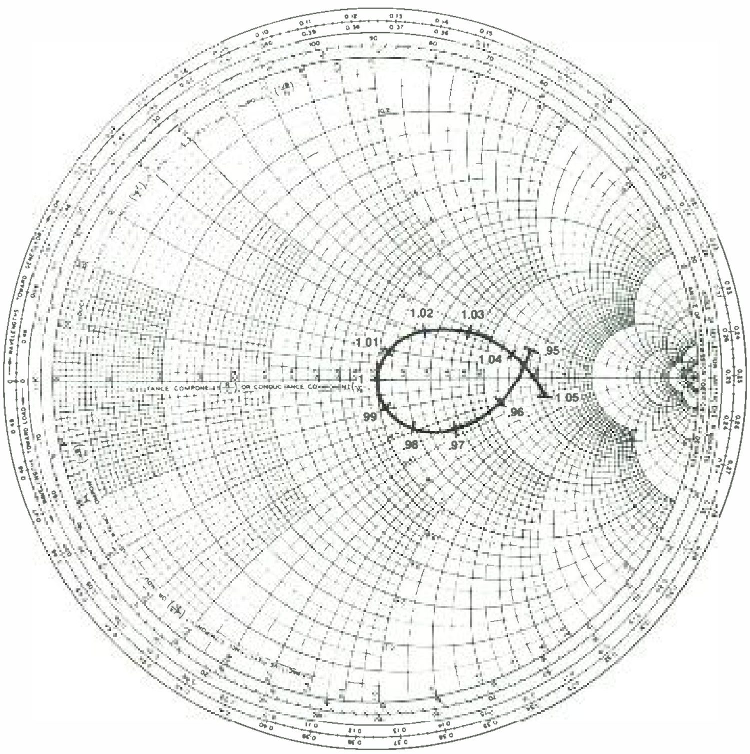


Fig. 12. Impedance vs. frequency characteristic at anode of final RF power amplifier.

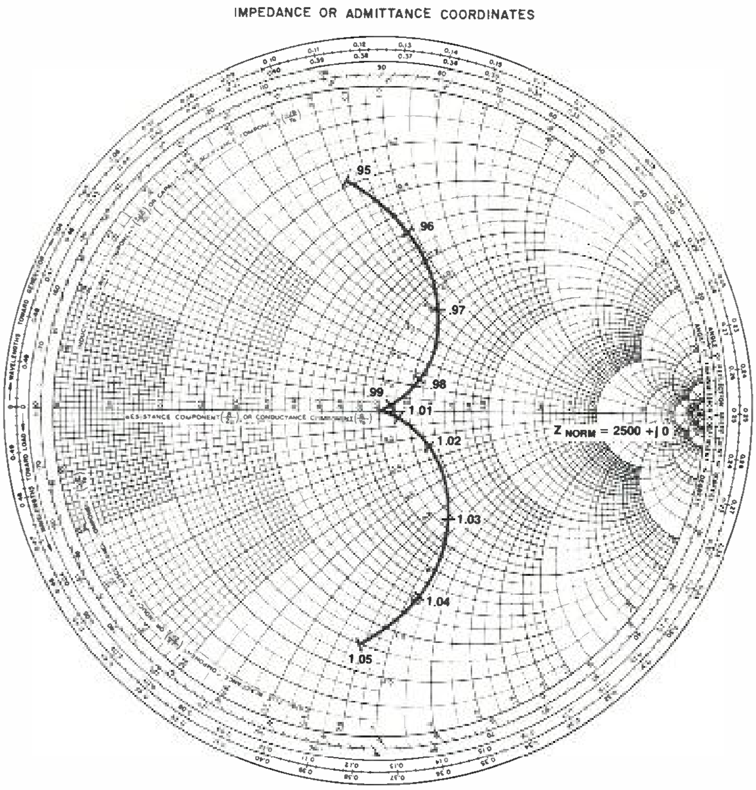


Fig. 13. Impedance vs. frequency characteristic at output anode with sideband mismatch corrective network added.

The shape of the impedance curve in Fig. 12 differs significantly from the shape of the terminating impedance curve shown in Fig. 11 due to the narrowing of bandwidth caused by the output matching network. The VSWR (normalised to the resistive carrier impedance) at ± 10 kHz is about 1.2 in both Figs. 11 and 12 but at ±50 kHz the VSWR at the input to the network has increased from about 2.4 to approximately 3.0. The shape of the impedance vs frequency curve at the anode(s) of the RF output amplifier tube(s) [or transistor collector(s)] has an important role to play in the high audiofrequency performance characteristics of the transmitter, such as frequency response and harmonic distortion.[14] The shape of the impedance vs. frequency curve should be symmetrical about the resistive axis of the Smith Chart and should yield lowest practical VSWR values at the highest expected fundamental sideband frequencies. Fig. 13 shows the result of an attempt to lessen the magnitude of the mismatch at the input to the matching network by the addition of the dotted components shown in Fig. 10. The result of adding the dotted components in Fig. 10 is a significant reduction in tube anode load mismatch at sideband frequencies of ±10 kHz; i.e., from about 1.2 VSWR in Fig. 12 to approximately 1.1 VSWR in Fig. 13. The reduction of mismatch at ±10 kHz, however, comes at the expense of a worsening of mismatch at sideband frequencies greater than ±30 kHz. The worsening mismatch at higher sideband frequencies is normally not a problem and often tends to improve transmitter harmonic distortion performance, acting as an output bandpass filter to higher order harmonics. There is theoretically no limit to the amount of sideband mismatch correction that can be achieved with more complex network circuitry. The most important criteria is symmetry about the Smith Chart resistive axis. Station engineers or their engineering consultants should work with the transmitter manufacturers and antenna system designers to, first determine, then achieve the desired impedance vs. frequency curve at the output RF amplifier tube(s) anode connection. The example given above for load sideband mismatch correction is for illustrative purposes only. It is quite common for the complex impedance at the sideband frequencies to present a more severe mismatch at the antenna common point than shown in Fig. 11. More severe mismatch and im-

[14]"Operation of AM Broadcast Transmitters into Sharply Tuned Antenna Systems". W. H. Doherty, Proc. IRE, Vol. 37, No. 7, July 1949, p. 729.

pedance dissymmetry require more complex sideband mismatch corrective networking.

Transmitter Control and Monitoring

AM broadcast transmitter control circuitry is normally very basic and uncomplicated. It is common to find the task of transmitter control performed with discrete digital IC logic circuitry in modern designs that used to be accomplished with simple relay control logic. Some manufacturers are incorporating microprocessor technology in their latest equipment designs to replace discrete digital logic circuitry. The aim of manufacturers is normally to provide the operating engineers and technicians with the most basic, reliable, and easy to maintain and troubleshoot transmitter possible. Experience has shown that well designed relay control logic, discrete digital IC logic, and microprocessor-based logic are all about equal in terms of reliability and ability to perform the required basic transmitter control functions. Future microprocessor-based transmitter control logic offers the promise to provide self and remote assisted diagnostics of transmitter problems and remote interrogation of transmitter operating parameters. Future microprocessor based logic offers the added promise of altering the basic characteristics of a transmitter control system through software control, allowing the basic transmitter design to be more easily "customized" to individual users' operating requirements. Because of the high voltage and high current faults that can exist in any high power transmitter component, extra care must be taken by designers and manufacturers of high power AM broadcast transmitters to prevent the potential destructive energy in these faults from affecting the performance and operation of the relatively delicate solid-state control logic circuitry. High speed vacuum contactors and solid state regulator/controllers are being used in many modern designs to control the high voltages and currents encountered in all levels of high power AM transmitters, which previously had been controlled by slower though equally reliable air-magnetic contactors and relays.

Remote control systems are commercially available that will allow remote control of almost any transmitter, old or new. These remote control systems, like the basic transmitter control system, use relay, discrete digital IC, microprocessor-based logic circuitry, or combinations of these, depending on the manufacturer of the remote control equipment and the complexity of the remote control functions desired. Many transmitter equipment manufacturers provide limited built-in remote control functions and circuitry.

The FCC requires manufacturers of AM broadcast transmitters to provide certain basic control and monitoring functions for operator safety and convenience and proof of certain performance parameters. These requirements are contained in the FCC Rules and Regulations Volume III (October, 1982) Parts 73.49 and 73.1215.

High Voltage Power Supplies

High voltage power supplies in AM broadcast transmitters must be designed to provide minimum acceptable performance in two basic areas: power supply ripple, which affects transmitter hum and noise output, and dynamic regulation, which affects low frequency modulation transient response. It is typical for transmitters of five kilowatts carrier power and lower to operate from a single phase ac power sources, usually 240 volts. Transmitters with carrier power ratings of ten kilowatts operate with single phase or three phase power source, depending upon the manufacturer of the transmitter. Transmitters with carrier power ratings of 50 kW or higher operate only from three phase power source, usually 480 volts for power levels up to 100 kW and 4160 volts in North America for higher carrier power levels. Three phase power has the advantage of being easier to filter and usually providing better dynamic regulation of critical modulator voltages than single phase supplies. Single phase power, on the other hand, is more readily available, which is the only reason it is used at the lower transmitter power levels, because initial installation cost would be disproportionately increased were three phase power required. Single phase rectifier power supply systems generally require L/C filtering to provide the necessary low ripple output for low transmitter hum and noise specifications. L/C filtering also creates, however, power supply resonances in the audible to sub-audible range of modulating frequencies, and therefore is a source of poor dynamic power supply regulation when the modulator circuitry of the transmitter is excited by vowel sounds or musical percussion sounds. It has been common since about 1970 for higher power transmitters to use special high voltage supply transformers to generate a six-phase ac supply from the basic three phase power source. The six-phase supply, when full wave rectified, yields a 12 pulse rectified dc waveform that has both lower ripple content and a higher ripple frequency than conventional three-phase full wave rectification. As a result, the output of the rectifier can be sufficiently filtered with no additional filter inductors, thus improving low audio modulating frequency dynamic power supply regulation and hence, low frequency transient distortion.

TRANSMITTER PERFORMANCE MEASUREMENTS

Certain basic AM transmitter performance parameters should be measured on a periodic basis in order to assure that certain minimum broadcast quality standards are provided the listening public. Excluding the performance standards for AM-Stereo, these are listed below. (From Volume III of the October 1982 edition of the FCC Rules and Regulations). Some are specifically required by the FCC, while others are a matter of good engineering practice.

a) Operating power
[Part 73.51]

b) Carrier output power delivered to the antenna system
[Part 73.54]

c) Modulation capability

d) Total audio frequency distortion

e) System frequency response

f) Carrier-amplitude regulation (Carrier Shift)

g) Hum and noise output level

h) Carrier frequency tolerance
[Part 73.1545(a)]

All audio measurements are to made from a demodulated signal at the antenna system common point.

Regarding item b above, the measurement of transmitter power output by the direct method described in Parts 73.51 and 73.1215 is subject to more than 13% error if allowed FCC measurement inaccuracies are taken to the limit. For example, assume that a common point impedance of 50 ohms resistive can be measured to within 2% accuracy with a radio frequency impedance bridge, a realistic tolerance. Further assume a direct reading RF ammeter having a full scale reading of 100 amperes and an FCC allowed tolerance of ±2% of full scale is used for this measurement and indicates 33.33 amperes of common point current; i.e., just meeting the minimum FCC accuracy and indication requirements in Part 73.1215(b)(2) and (3). Under these conditions, the actual power delivered to the antenna is between the limits of (35.33 × 35.33 × 51) 63.65 kW, as a maximum, and (31.33 × 31.33 × 49) 48.10 kW, as a minimum; yielding a total measurement error of approximately +14/−13%. Using an RF ammeter with a 50 ampere full scale reading, the same FCC allowed inaccuracies, and the same meter indications as above, would result in power output measurement errors of approximately ±8%.

FACTORY TESTS

When buying or looking for an AM broadcast transmitter to buy for replacement or new equipment, it is advisable to ask manufactures for specific and detailed test and performance data at the start of the buy investigation. Before a final decision is made, specific and detailed tests should be made at the manufacturers factory under strict control of an experienced engineer or engineering consultant. Most manufacturers of AM broadcast equipment welcome this kind of intelligent approach to the decision making process. Some hints on what kind of tests and a discussion on the details of each test is given below.

Demodulated Characteristics (Audio Performance)

As stated earlier in the chapter, most audio performance characteristics reached a plateau in the late 1940's to early 1950's which has not seen significant improvement in even the most modern AM broadcast transmitter designs.

Audio Frequency Response

At one time the term "broadcast quality audio" meant a standard to which all other system equipment, by comparison, was inferior. Today the audio quality of consumer high fidelity and stereo equipment surpasses, with one exception, the signal quality which can be broadcast by any AM transmitter manufactured today or likely to be made in the future. The one exception is audio frequency response. Practically all mass production AM receivers made in any country of the world have IF and audio amplifier bandpass characteristics that limits receiver -3 dB high end audio frequency response to between approximately 2500 Hz and 5000 Hz, with 2500 Hz the more common of the two figures. Typical low end -3 dB frequency response of consumer AM radios is between 100 Hz and 300 Hz, with approximately 200 Hz a common value. AM transmitter frequency response characteristics, referred to a reference frequency of either 400 Hz or 1000 Hz, typically is well within ±1 dB from 50 to 10000 Hz at any depth of modulation.

Audio Harmonic and Inter-Modulation Distortion

Current production AM broadcast transmitters typically produce less than 2% total harmonic distortion (THD) up to 90% modulation at any frequency of modulation between 50 and 10000 Hz for monophonic transmission.

Intermodulation distortion (IMD) has been known for years, and has been documented in technical audio journals, to be a more disturbing kind of distortion than harmonic distortion; though both are important. The CCIR method of IMD measurement is the preferred method for radio broadcast transmitters. With this method, two equal audio tones separated by 170 Hz are input to the transmitter and the peak modulation level adjusted to between 85 and 95% modulation. The level of odd and even order products are measured using an audio wave or spectrum analyzer connected to the test output terminals of a high quality modulation monitor. Two IMD measurements should be taken, one with the two tones near mid audio band, for example: 400 and 570 Hz, and one test with the two tones near the upper audio end, for example: 7000 and 7170 Hz. High quality broadcast transmitters should produce IM distortion products more than 30 dB below the level of either of the two modulating tones. The RSS value of all IM products, relative to the level of either modulating tone, should also be less than 7% at 90% peak modulation levels.

Residual AM Hum and Noise

The FCC requirement for AM noise of about 60 dB below the 400 Hz/100% modulation level can be achieved by most current production AM broadcast transmitters. The bandwidth of noise measuring equipment should be 20 kHz. Typical modulation monitor demodulated audio bandwidth is approximately 25 kHz to accommodate both of these FCC bandwidth requirements.

Residual PM Hum and Noise

Residual Phase Modulation (PM) noise is normally not a problem for modern AM broadcast transmitter designs for monophonic broadcasting. The quartz crystal oscillator circuits and even moderately careful RF component mechanical designs produce quite acceptable phase noise characteristics. However, for AM-Stereo applications, where a frequency synthesizer may be used for the RF signal generation, is it wise to perform a test to determine the purity of the synthesizer circuitry. Very excessive residual PM noise can convert to AM noise over certain nighttime propagation paths and appear to distant listeners as objectionable AM hum and noise. An acceptable value of PM noise is -25 dB rms relative to one radian peak, measured in a 15 kHz bandwidth, for monophonic medium wave band AM transmission. Transmitters with quartz crystal oscillator exciters typically exceed this recomendation by 25 dB or more. Transmitters used for international short wave broadcasting (4 to 26 MHz) require PM noise levels of approximately -45 dB relative to a one radian peak because of more severe skywave PM to AM conversion at higher frequencies.

Incidental Phase Modulation (IPM)

Like residual PM, IPM is more important in stations using or planning to use AM-Stereo transmission. IPM can be defined as the peak phase deviation *of the carrier frequency* (in radians) resulting from the process of amplitude modulation. IPM values of several radians were common in the very early days of broadcasting. Typical values of IPM for modern transmitters that have not been specifically designed or adjusted to minimize IPM range from about 0.1 to 0.5 radians peak (approx. 6 to 30 degrees). A maximum acceptable value of IPM required for present or future AM-Stereo operation is generally considered to be approximately 0.05 peak radians with a desired value of better than 0.02 peak radians. State-of-the-art modulation meters such as the Hewlett-Packard model HP-8901 and other similar instruments are preferred for accurate PM and IPM measurements.

Carrier Amplitude Regulation (Carrier Shift)

The amount of carrier level shift in a given transmitter has more importance to overall transmitter performance than many broadcast engineers realize. Large values of negative carrier shift can have as much effect on effective transmitted sideband power as poor transient overshoot distortion. The term "carrier shift" may be somewhat confusing, especially to newcomers in radio broadcasting who often equate the terminology with frequency shift instead of level or amplitude shift, hence, the new terminology of "carrier amplitude regulation". The older terminology, however, is still used quite extensively by older engineers as well as new engineers. The CCIR refers to the same characteristic as "carrier level shift", which appeals to many engineers because of its closer adherence to the original terminology, but without the ambiguities.

"Carrier level shift" is the effective shift in apparent carrier level due to the amplitude modulation process. Carrier level shift can be caused by either poor power supply regulation, or modulation even order harmonic distortion, which generates a dc offset component in the modulated RF envelope, or both. Carrier level shift can be either positive or negative, though is usually negative because power supply regulation is most often the major source of carrier level shift and power supply regulation is most generally negative in sign i.e., lower voltage output at higher current loads. Poor power supply regulation is not always caused by the transmitter power circuitry, it can also be, and often is, caused by poor supply line voltage regulation or, more generally a combination of the two.

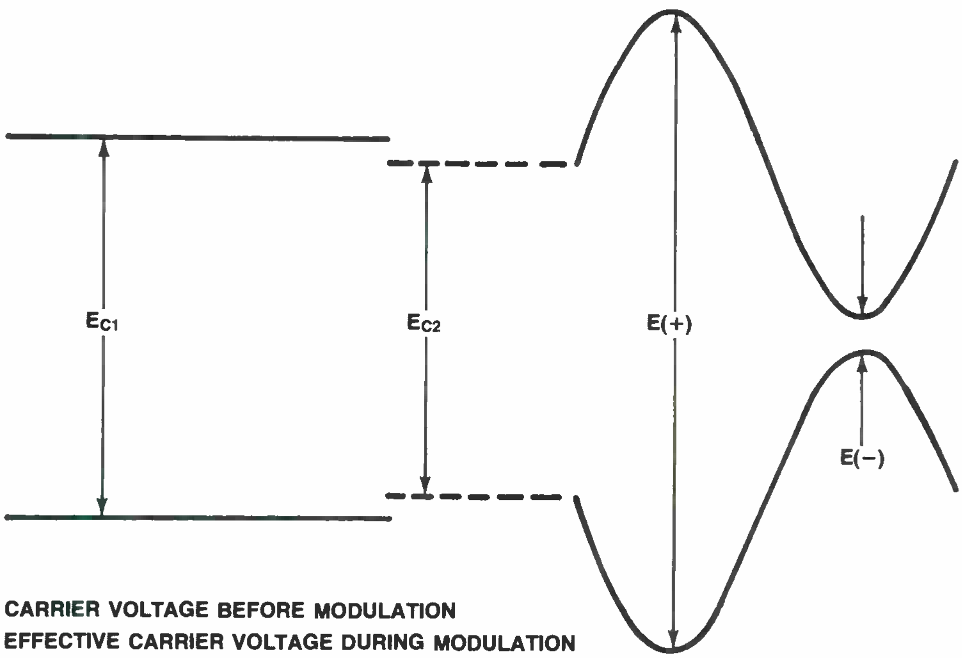
Another common misconception regarding carrier level shift is that the defined level shift is direct shift in carrier power. Actually, carrier level shift is defined as the shift in effective carrier voltage or current due to the process of modulation. This means that a carrier level shift of minus 5% is equivalent to a carrier power shift of approximately minus 10% i.e.,

$$\begin{aligned} P_{\text{carrier (mod)}} &= \text{Pcarrier} \times (1-0.05)^2 \\ &= 0.9025 \times \text{Pcarrier} \end{aligned}$$

Transmitters having no carrier level shift produce an average output power of 1.5 times the carrier power level with 100% sinusoidal tone modulation. Transmitters that exhibit -5% carrier level shift produce an average output power only 1.35 times the carrier level at the same conditions of sinusoidal tone modulation. A broadcast station engineer or engineering consultant needs to have complete understanding of these and other equally important transmitter characteristics in order to be able to make an intelligent buying decision based upon measurable and proven technical merit. Fig. 14 gives a graphical representation of carrier level shift.

Audio Phase Linearity

Proper attention is given to phase linearity by most station engineers and engineering consultants. Station managers and engineers are usually properly concerned about the "sound" of their stations; and loudness, or perceived loudness, is a common criteria of "quality" in many stations with diverse programming formats. However, it is not uncommon to learn that many



E_{C1} = CARRIER VOLTAGE BEFORE MODULATION
E_{C2} = EFFECTIVE CARRIER VOLTAGE DURING MODULATION

$$\text{CARRIER LEVEL SHIFT (CS)} = \frac{\dfrac{E(+) + E(-)}{2}}{E_{C1}}$$

$$\text{\% MODULATION} = \frac{E(+) - E(-)}{E(+) + E(-)} \times 100$$

Fig. 14. Graphical representation of carrier level shift and formulas for calculation of carrier level shift and % modulation.

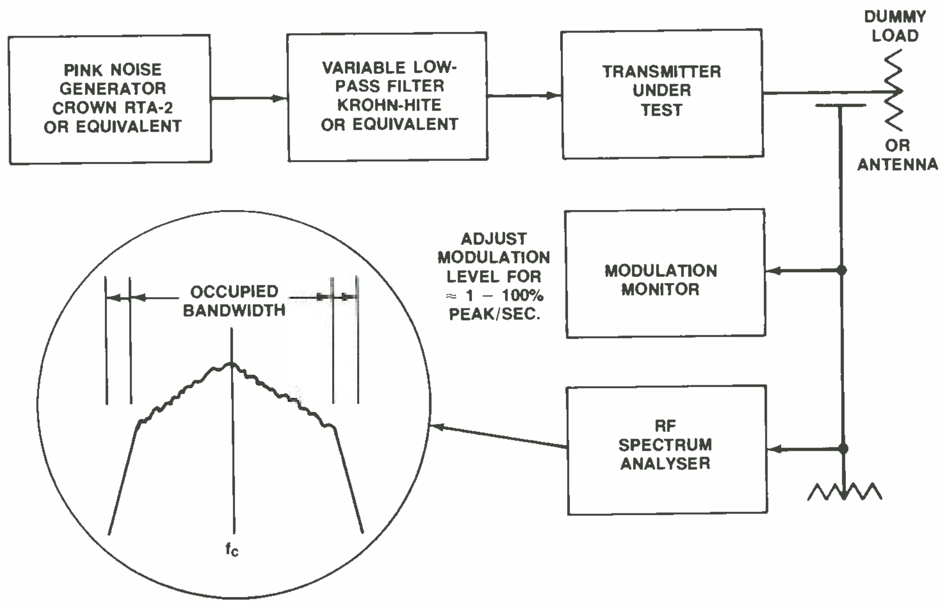


Fig. 15. Block diagram test procedure for determining transmitter occupied bandwidth.

station engineers and program directors spend more time researching the various objective and subjective merits of program processing equipment than the one characteristic in their potentially more expensive transmitter purchase that could partially neutralize the potential advantage from a new or different program processor; that one characteristic is audio phase linearity.

Audio phase linearity, or more accurately audio phase non-linearity, and its major detrimental result, transient overshoot, has been discussed in previous sections of this chapter. Before the popularity of modern broadcast audio processing and some modern broadcast programming format, audio phase linearity of a transmitter or any component in the program audio chain, was not as significant as it is today. (Modern programming philosophy, and the program processors that philosophy has caused to be manufactured, have made transmitter audio phase linearity characteristics a more important performance criteria.) In the late 1960's, at the long and persistent insistence of the AM broadcast industry, the FCC reluctantly authorized 125% positive program modulation, allowing AM transmissions to accomodate certain naturally occuring asymmetry in voice and music thus achieving a gain of 2 dB of real loudness or actual program sideband power; i.e., 20 log(1.25) = 2 dB. As stated earlier in paragraph 4.1.1, some AM broadcast transmitters in present use exhibit as much as 12% overshoot of a squarewave input, due to phase non-linearity, which has the effect of taking away one of those 2 dB the industry tried so hard to get; i.e., 20 log(1/1.12) = -1 dB.

A simple way to determine the effects of phase non-linearity, is to require the prospective transmitter manufacturer to demonstrate rectangular wave modulation characteristics of the transmitter under investigation. When such a test is performed, the overshoot produced by the transmitter will be directly visible and measurable on the oscillographic display of the transmitter output envelope. The station engineer or his engineering consultant should also investigate the effects that antenna system phase non-linearity may have on the total system transmission transient characteristics.

Occupied Bandwidth

Occupied bandwidth of an AM broadcast transmitter can best be measured with the use of a band limited colored gaussian noise source, similar to pink noise used in certain acoustical tests, to provide a continuous wideband modulating signal to the transmitter. Pink noise has equal energy per bandwidth ratio; i.e., equal energy per octave, per third octave, per tenth octave, etc. White noise has equal energy per bandwidth; i.e., equal energy per Hz. Both noise signals can have a gaussian or pseudo-gaussian probability density function, a probability density function that closely resembles that of typical voice and music. Fig. 15 shows a block diagram of the test procedure used to measure occupied bandwidth. The reader is also referred to CCIR recommendation 559-1 (Vol X Part 1 Geneva, 1982). The measurement of occupied bandwidth is a dynamic measurement which effectively summarizes two important transmitter static parameters; audio non-linear distortions, which are the source of

IMD, and THD, and incidental phase modulation (IPM).

Harmonic and Spurious Output

Harmonic and spurious output of an AM broadcast transmitter or transmission system can only be effectively measured two ways. One way is to use a sample of the transmitter RF output signal when operating into a dummy load with a known or accurately measurable impedance characteristic out to approximately the tenth RF harmonic. Then, with a calibrated measuring system, measurements and calculations are made to compute the output power of each harmonic according to the formula:

$$P_N = V_N^2 \times R_{pN}$$

Where:
P_N = the power at the Nth harmonic
V_N = the corrected measured voltage of the Nth harmonic at the calibrated impedance point
R_{pN} = The parallel resistive component of the load impedance at the Nth harmonic

A second way to effectively determine the harmonic and spurious output of an AM broadcast transmitter is to measure the actual power radiated from the antenna system at each harmonic frequency or suspected spurious output frequency by standard field intensity measurement techniques. The field intensity method is the most meaningful of the two techniques because it allows the measurement to be made under actual conditions of operation with all systems interconnected. It is normally the new transmitter manufacturer and users joint responsibility to correct any actual interference problems to other broadcast or non-broadcast communication services even when the particular interfering signal meets the standard FCC requirements. (FCC Rules Part 73.44.c.)

Carrier Output Power

The most accurate method of measuring the RF output power of a transmitter is by the calorimetric method, a method which uses the very accurately known and measurable physical and thermal characteristics of water or other similarly well defined liquids. This measurement is usually only done in a transmitter manufacturer's factory because the capital investment required to purchase and maintain calibration of this kind of infrequently used equipment is usually not justified for AM broadcasting operations.

Water is known to have a thermal capacity very close to 4.186 Joules per degree C per gram weight at a mean temperature of 60 degrees centigrade. A Joule is equal to one watt-second. Therefore, the capacity of water to absorb power is 69.8 watts per degree C per liter of water flow per minute, or

Power (kW) = Flow (lpm) × $DT°C$ × 0.0698 for water flow measured in liters per minute

or, Power (kW) = Flow (gpm) × $DT°C$ × 0.2641 for water flow measured in U.S. gallons per minute (gpm)

The flow of water can be measured with an accuracy of approximately ±1%, by even the most common methods. Differential temperature measurement accuracy of approximately ±0.1°C is commonly practical, which, for temperature differentials of 20°C, is equivalent to ±0.5% accuracy. Therefore, using calorimetric measurement techniques, the output power of AM transmitters can be measured with total accuracies better than ±2%. The RF output amplifier efficiency factor F, referred to in FCC Rules Part 73.51.e.1, can then be determined for future operating and proof of performance reference. However, even with this method of determining the factor F, an accuracy of less than ±2% cannot be maintained over the life of the equipment. The required FCC transmitter voltage and current meters have a ±2% accuracy, in addition to the multiplied power accuracy of approximately ±4%. When accuracy levels are combined, this fields a total uncertainty of ±6% for the factor F. Still, this is considerably better than the accuracies obtainable by the direct measurement technique.

Operating Efficiency and Input Power

The determination of transmitter input power should be done under actual or simulated operating conditions. This requires program or simulated program signals for the transmitter modulating source during the period of power input measurement is by use of a standard rotating disk watt-hour meter. The measurement accuracy of these familiar watt-hour meters is typically better than 0.5%, better than four times the accuracy of any other conventional direct ac power measurement technique. The watt-hour metering system should be connected in the main power feed line to the transmitter under test. Sinusoidal test signals, useful for other tests of transmitter performance, are not recommended for power consumption tests because of the distinctly opposite statistical characteristics of sinusoidal signals and typical voice and music program

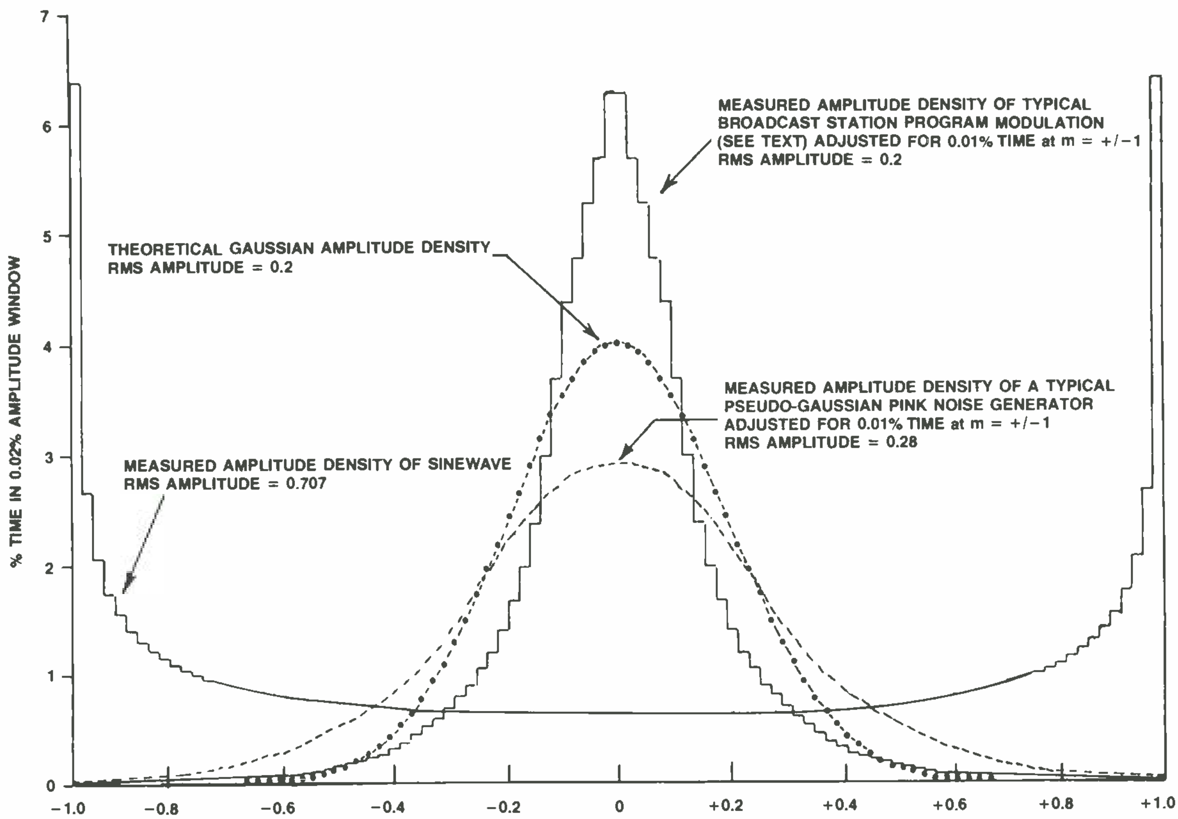


Fig. 16. Amplitude density of sinusoidal test signals and typical program or noise test signals. Program or noise test signals recommended for transmitter efficiency measurements (see text).

material and the effect this difference has on actual operating power consumption measurements. This difference between periodic sinusoidal signals and mathematically random types of signals such as human voice and/or music has been known for decades. The effect this difference produces in AM broadcast transmitter power consumption and operating efficiency measurements, however, was first documented in 1980, by investigators in Europe[15] and was later verified and further explained by investigators in the US[16 & 18] and other countries.

The critical difference between sinusoidal signals and typical program types of signals is explained in Fig. 16 which shows the amplitude density characteristics of a sinusoidal waveform (U-shaped curve) and an amplitude density characteristic of a typical broadcast program modulation (pyramid curve). The totally different shapes of the two curves in Fig. 16 cause the measured transmitter efficiency at identical RMS modulation levels also to be quite different. The program measurement of five differently formatted FM radio stations located in Dallas, TX. Data was taken for continuous ON-AIR hours for each station and averaged for presentation as shown in Fig. 16. FM stations were used in the collection of the data because of their consistent day/night signal levels, symmetrical modulation characteristics for easier comparison to the sine wave, and omni-directional emission patter. Some transmitters, or more correctly modulation systems, are more efficient with program modulation than with equivalent RMS sinusoidal test signals while others have poorer efficiency with program modulation techniques that yield higher carrier efficiencies also yield higher program efficiencies; i.e., lower program power consumption than with an equivalent RMS modulation sinusoidal test signal. It is obvious that the program input power consumption test is the one most meaningful to broadcasters whose purpose it is to broadcast programs, rather than sinusoidal tones. In this discussion, the terms "program efficiency" and "program input power consumption" have generally been given equal weighted

[15]"Efficiency of High Power Broadcasting Transmitters in Regular Programme Service". M. Sempert and W. Tschol, BBC Review, July 1980.

[16]"Efficiency Comparison of AM Broadcast Transmitters". G. W. Woodard, IREE Journal (Australia), Vol.2 ,No. 2, June 1982.

[18]"Simulating Typical Programme Modulation for Measurements of Operating Efficiency and Modulation Capability of AM Broadcast Transmitters". G. W. Woodard, Radio and Electronic Engineer (UK),Vol. 53,No. 9, p. 325.

value. It is correct to equate the weighted value of these two parameters for transmitter to transmitter comparisons if it is assumed or defined that average transmitter-to-transmitter output power is constant for a defined program input. Such an assumption is correct except that transmitters with excessive transient overshoot will have correspondingly less average modulated RF power output for given peak levels of processed program modulation. Experimental methods have been devised to measure average long term transmitter output power; i.e., RF kilowatt-hour output.[17] With such equipment it would be convenient to accurately compute actual program efficiency knowing both kilowatt-hours input and kilowatt-hours output. Until such equipment becomes commercially available, however, only the transmitter input power can be accurately measured with the familiar single phase and multi-phase rotating disk kilowatt-hour instruments.

Fig. 16 also shows a third curve (dashed curve) which is the amplitude density function of a Crown model RTA-2 pink noise generator. It is recommended that either a pink noise signal with amplitude density characteristics similar to the curve in Fig. 16 or a recording of actual program modulation be used as a program source for factory input power consumption tests. Program processing equipment, if used, and peak modulation levels should be adjusted to satisfy normal station operating procedures and transmitter input power measurements then taken during a 30 minute (minimum) segment of the program source material. Refer to FCC Rules Part 73.1570(a) and (b) for modulation setup procedures. The average input power determined by the method described will be very close to the transmitter power consumption during its total operating life and therefore can be used to accurately predict actual operating energy costs. This is the only method of transmitter power consumption tests that will provide accurate energy consumption forecasting data.

IMPORTANT AM-STEREO TRANSMITTER CHARACTERISTICS

Incidental Phase Modulation

Several of the proposed AM-Stereo systems utilize a form of phase modulation for encoding the stereo signal onto the carrier of the AM signal. It is for this simple reason that the most important transmitter characteristic affecting AM-Stereo operation is incidental phase modulation (IPM) which has been previously discussed in section 7.1.5. Excessive IPM can affect stereo separation, single channel distortion, and the occupied bandwidth of a stereo transmission; with the most significant of the three being single channel distortion.

Recognizing the various causes of IPM is the first step in correcting the trouble. There are many potential sources of IPM but the most common source is insufficient amplifier neutralization, either of a final modulated RF amplifier or of a lower power driver stage. The solution is, of course, to perform better neutralization of the offending amplifier stages. This is easier said than done in most cases. Since the inception of AM-Stereo broadcasting, manufacturers of the transmitters have paid more attention to the problems of IPM and in most cases have reduced production levels of IPM in current model transmitters to acceptable levels for AM-Stereo operation. In some cases, for especially difficult problems, field engineers from the manufacturers factory must make special on site adjustment of individual transmitters which are particularly susceptible to excessive IPM. Several engineering consulting firms have developed special knowledge in the AM-Stereo field, having collected data on many transmitters in current use, some of which have been out of production for several years. It is suggested that such a firm be contacted to solve particularly difficult neutralization procedures on old or new equipment which may not have been satisfactorily performed by the original manufacturer.

Phase Noise

Residual phase modulated noise cannot normally be detected by a standard AM broadcast receiver employing envelope detection, while stereophonic AM receivers are sensitive to PM noise. PM to AM conversion can occur on the medium wave band over multi-ionospheric "hops" (night-time skywave propagation) which can then be detected by receivers employing standard envelope detection. Very early transmitters sometimes produced more residual phase noise sidebands than those produced by the desired program amplitude modulation. Night-time reception of distant stations sometimes was accompanied by a 120-360 Hz "roar" which was caused by phase modulation from the master oscillator filament supply. A somewhat subdued "roar" can be heard even today from many shortwave stations using older transmitting equipment, especially on the higher shortwave bands where multi-hop propagation is more prevalent.

Significant phase modulated noise in modern transmitters is virtually nonexistent due to the use of high quality quartz crystal oscillator and syn-

[17] "Measurement of the efficiency of high-power AM transmitters with regular broadcast programmes". A. W. Jaussi and W. Tschol, EBU Review, No. 194, August 1982, p. 169.

thesizer circuitry. Phase noise modulation of 0.6 degrees (0.01 radians) avg. is fully acceptable for monophonic AM broadcasting. Phase noise modulation of approximately 0.2 degrees (0.0032 radians) avg. is usually considered acceptable for AM stereophonic broadcasting. As with measurements of Incidental Phase Modulation (IPM), an HP-8901, or equivalent, Modulation Analyzer is recommended for phase noise modulation measurements.

Stereophonic Phase/Gain Equalization

Standard production exciters for AM-Stereo systems may incorporate built-in circuitry designed to approximately match the phase and gain characteristics of the normal monophonic transmitter transmission path to the transmission path for the encoded stereo signal. Transmitters which have excessive in band non-linear phase characteristics sometimes require special "out-boarded" phase/gain equalization networks to achieve optimum stereo performance.

INTERNATIONAL SHORTWAVE BROADCAST TRANSMITTERS

Introduction

International shortwave broadcasting in the United States and North America began at about the same time as medium-wave broadcasting, about 1920, but did not have any substantial growth until pre-World War II propaganda activities by the influential nations. Until this growth in the mid-1930's, shortwave broadcasting was done primarily by amateur and special industry groups using the higher frequency bands for experimentation and hobby interests. Almost immediately after the invasion of Pearl Harbor, accelerated shortwave broadcast activity in the United States began by the Office of War Information (OWI) which later assumed the wartime responsibility of the then inexperienced Voice of America (VOA) in June of 1942. Later, in mid-1943 the Armed Forces Radio Service began shortwave relay broadcasts to France and other parts of Europe from bases in England. After WWII, the Voice of America resumed control of their wartime expanded facilities and, together with virtually all nations, began continuous operations of native and foreign language broadcasts for information exchange and general entertainment purposes. Other private interest in shortwave broadcasting also began to expand after the war, primarily from various religious and politically based organizations seeking means to advance their cause or provide information and service not otherwise available by government operations. Commercial shortwave broadcasting in the United States has generally not been profitable primarily because of the limited audiences in North America given the popularity of AM, FM, and TV broadcasting. The quality of signal in shortwave listening is not of the same quality as in either of the three other more popular modes of broadcasting, due to the long distances covered and atmosphere-generated fading and noise. Therefore, shortwave broadcasting in North America attracts a different kind of listener; one who is more dedicated to the "hobby" of shortwave listening and therefore more willing to tolerate the disturbances associated with short wave listening. Shortwave listening in the world outside North America is very popular, and necessary, among large segments of the population as a source of news, socio-political or religious view-point, and entertainment.

Shortwave Transmitters

Shortwave broadcast transmitters are similar in many respects to medium wave transmitters and very different in others. The similarities are in methods of modulation, control, and monitoring. The differences are, generally, that shortwave transmitters are higher power, more complex in tuning, and more difficult to operate and maintain than medium wave "standard" broadcast transmitters. Although there are numerous exceptions, the general rule, written or unwritten, is that the minimum usable carrier power level is 50 kW and the maximum economical carrier power level from a single transmitter is 250 kW to 500 kW.

It is not unusual for a shortwave transmitter to operate on five to ten separate frequencies every broadcast day. Modern broadcasting schedules of the prestigious broadcasting organizations are very tight, thus necessitating built-in automatic frequency changing circuitry which allows the transmitter to tune to several programmed frequencies in approximately 10 to 30 seconds, with minimum or no operator intervention. The trend in shortwave transmitter operation is toward unattended or minimum attended sites, with program and frequency changes done by remote and/or computer control.

Single Sideband Operation on the International Shortwave Bands

Since about 1964, the International Telecommunication Union (ITU) and its radio broadcasting special committee, the International Radio Consultative Committee (CCIR) have encouraged the adoption of a form of Single Sideband as the standard modulation system for international shortwave broadcasting. This recommendation has been prompted due to the ever increasing congestion in the shortwave broadcast bands, the competitive increases in transmitter power levels, which contributes to the congestion; and the realization of the increased efficiency of the SSB mode of transmission. A World Administrative Radio Conference Committee, meeting in Geneva in January and February of 1984, adopted a specific and detailed 20 year plan for conversion to SSB on the international shortwave broadcast bands. The committee report addressed necessary changes in both transmitter and receiver technologies, thus creating the beginning and emphasizing the importance of the changes which can double the available channel space and improve reception quality in the current international broadcast bands.

Figs. 35 and 36 show a block diagram and photograph of a 250 kW international broadcast band transmitter with ten channel semi-automatic tuning capability in use since 1980 by Radio Free Europe/Radio Liberty.

REPRESENTATIVE COMMERCIAL PRODUCTION TRANSMITTERS

This section shows examples and a brief description of some current model production transmitters which are representative of some of the basic transmitter types discussed in previous sections.

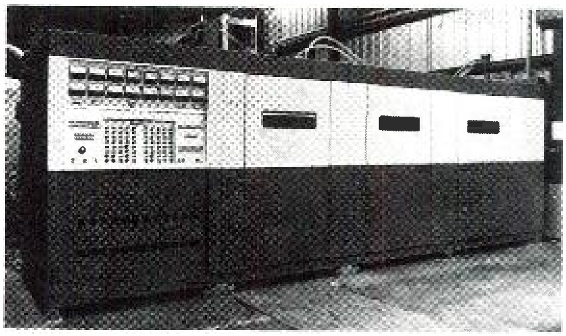


Fig. 36. Continental Electronics Model 419F semi-automatically tuned 250kW shortwave transmitter.

LPB Model AM-150

The LPB Model AM-150 is the highest power model of a line of low power transmitters used for various special broadcast applications such as Pre-Sunrise operation of daytime only stations, travelers information services, standby transmitters for emergency use, Armed Forces radio services, etc. None of these family of transmitters carries FCC type acceptance for broadcast use since there are no standard catagories for the use of these power levels in broadcasting. Some of the family line is approved or pending approval by the Department of Communications of Canada for continuous broadcast service.

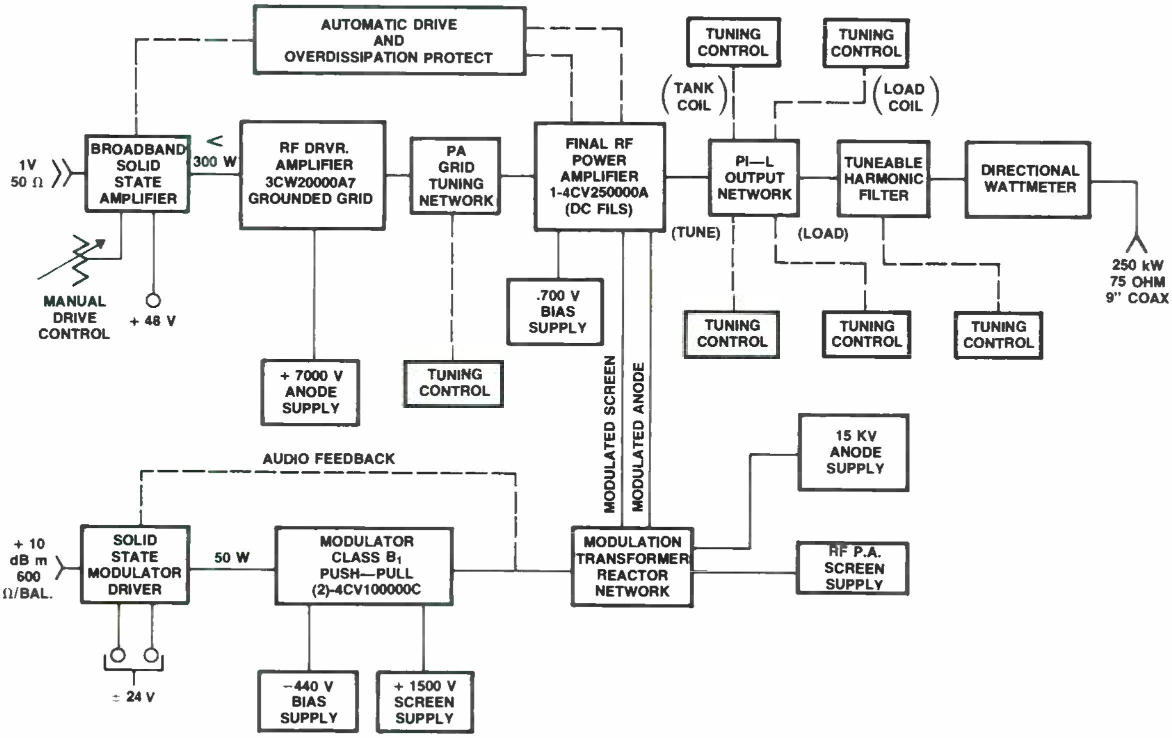


Fig. 35. Block diagram, Continental Electronics Model 419F 250 kW shortwave transmitter.

The transmitters use low level modulation followed by one or more conventional linear amplifier stages until the desired power output level is attained. The entire family of transmitters are completely solid-state and many have built in modulation limiting and modulation level metering. The Model AM-150 is shown in the photograph of Fig. 17.

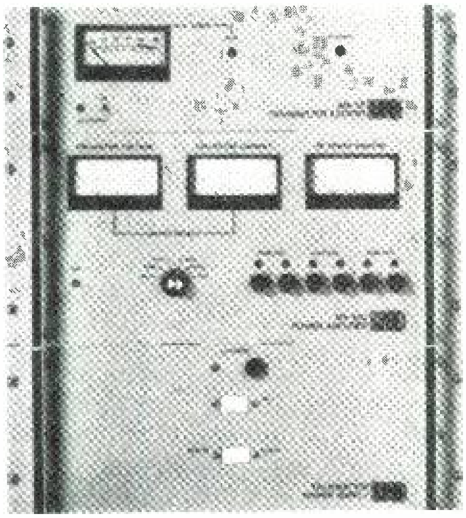


Fig. 17. LPB Model AM-150 pre-sunrise post-sunset low power AM transmitter.

McMartin Model PS-1K

The McMartin Model PS-1K is not a transmitter but is a unique product which is designed specifically for use in post-sunset pre-sunrise operation of daytime only stations in the United States. It is a passive device that reduces the power of any 250, 500, or 1000 watt transmitter to an authorized pre-sunrise level through the use of resistive attenuation networks. All controls for automatic or manual power change are built into the unit with facilities for proper transmitter interface during the switchover period. The unit requires no tuning or special external ancilliary items. A photograph of the Model PS-1K is shown in Fig. 18. For information regarding pre-sunrise authorization, the reader is referred to the FCC Rules and Regulations at Section 73.99 (may be found in Title 47 of the Code of Federal Regulations).

McMartin Model BA-1K 250-1000 Watt Transmitter

The McMartin Model BA-1K is a conventional Class "B" high level anode (plate) modulated transmitter. The BA-1K uses only one tube type, a 4-500A or 4-500B power tetrode,—two in the push-pull class "B" modulator final amplifier and

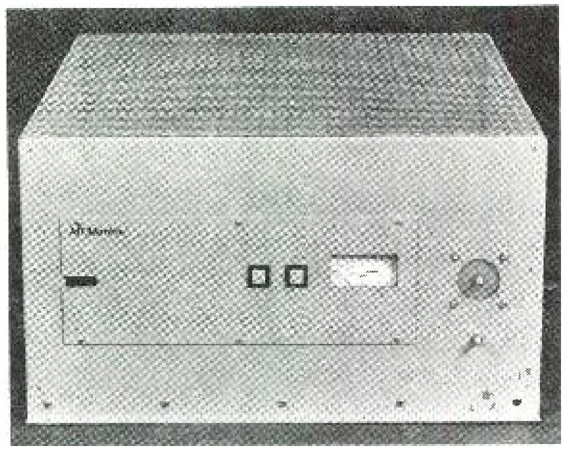


Fig. 18. McMartin Model PS-1K pre-sunrise post-sunset transmitter power reduction unit.

two in the final modulated RF power amplifier. All other stages are completely solid state. The BA-1K transmitter, employing an oil filled modulation transformer, was first manufactured in 1974, and has proven reliability in nearly 250 stations; approximately one-half in the United States. A block diagram and photograph of the unit is shown in Fig. 19 and 20 respectively.

Harris Models SX-1, -2.5, -5

The Harris family of all solid state transmitters employ high efficiency Class "D" final RF stages and high efficiency polyphase-pulse width modulation for high overall transmitter efficiency, typically greater than 72%, at all conditions of modulation. The transmitters are of modular construction to ease maintenance procedures and are the first in the industry to use a microprocessor control system. A block diagram (block diagrams of all models are identical) and a photograph of the Model SX-5 are shown in Fig. 21 and 22 respectively.

Nautel Models AMPFET 1, 5, 10

The Nautel family of all solid state broadcast transmitters first entered the market in 1982 and have established a good record of reliability and performance. The transmitters employ modular construction of the RF and Modulator units. The family of transmitters all employ high efficiency MOS-FET Class "D" final RF output amplifiers which are collector modulated by a high efficiency MOS-FET series pulse width modulation system. The manufacturer lists overall transmitter efficiency at better than 72% at all conditions of modulation. Approximately 20 transmitters of each power level are presently in operation or planned operation. A block diagram and photograph of the Nautel model AMPFET 10 is shown in Fig. 23 and 24 respectively.

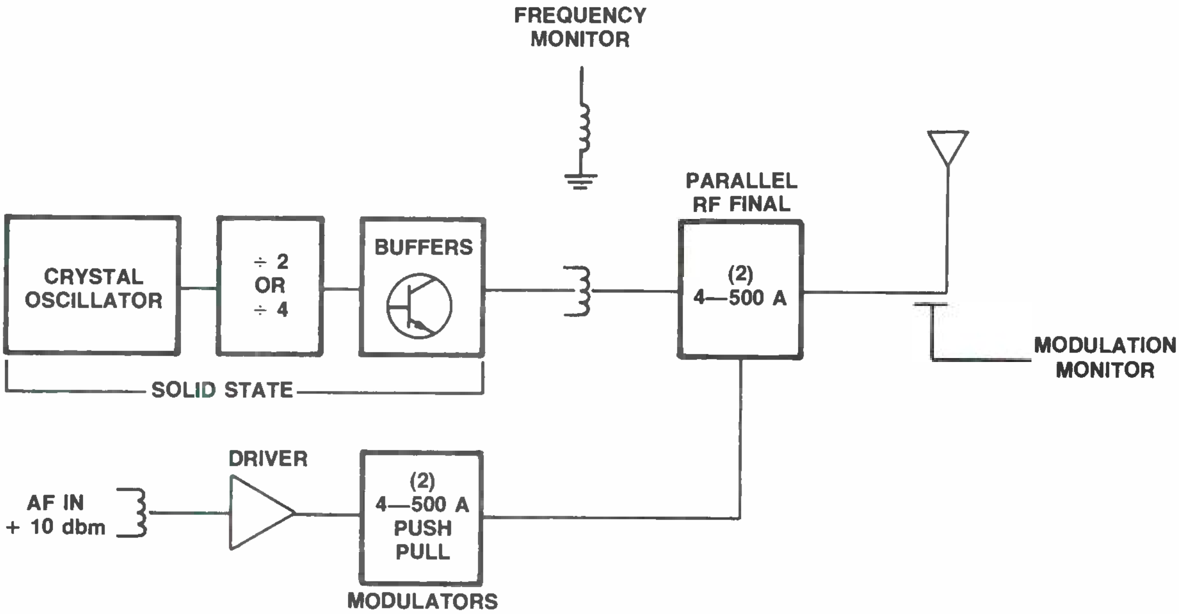


Fig. 19. Block diagram, McMartin Model BA-1K, 1 kW transmitter.

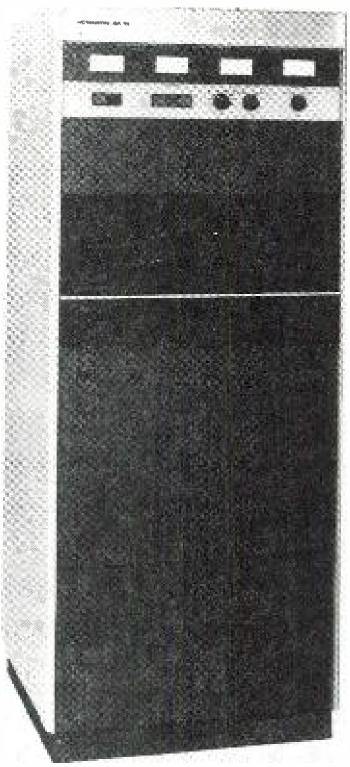


Fig. 20. McMartin Model BA-1K.

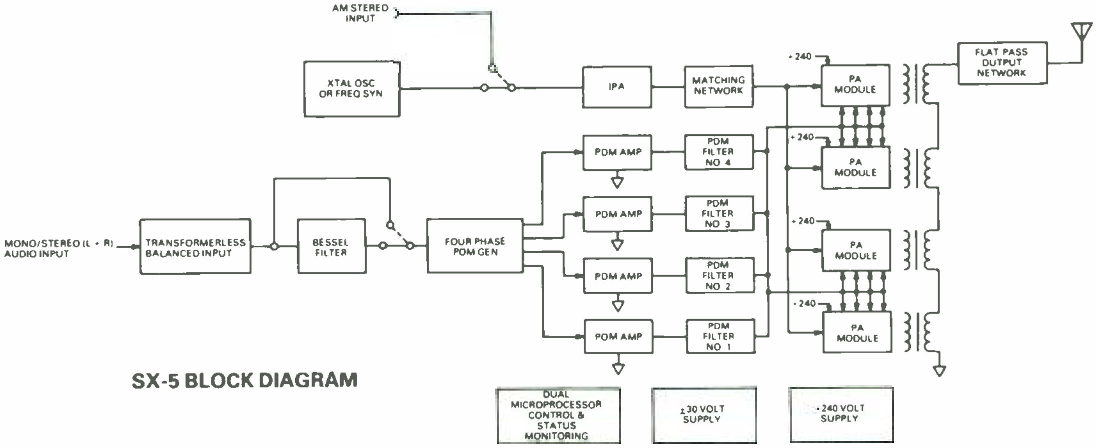


Fig. 21. Block diagram, Harris Model SX-5, 5kW transmitter.

Fig. 22. Harris Model SX-5.

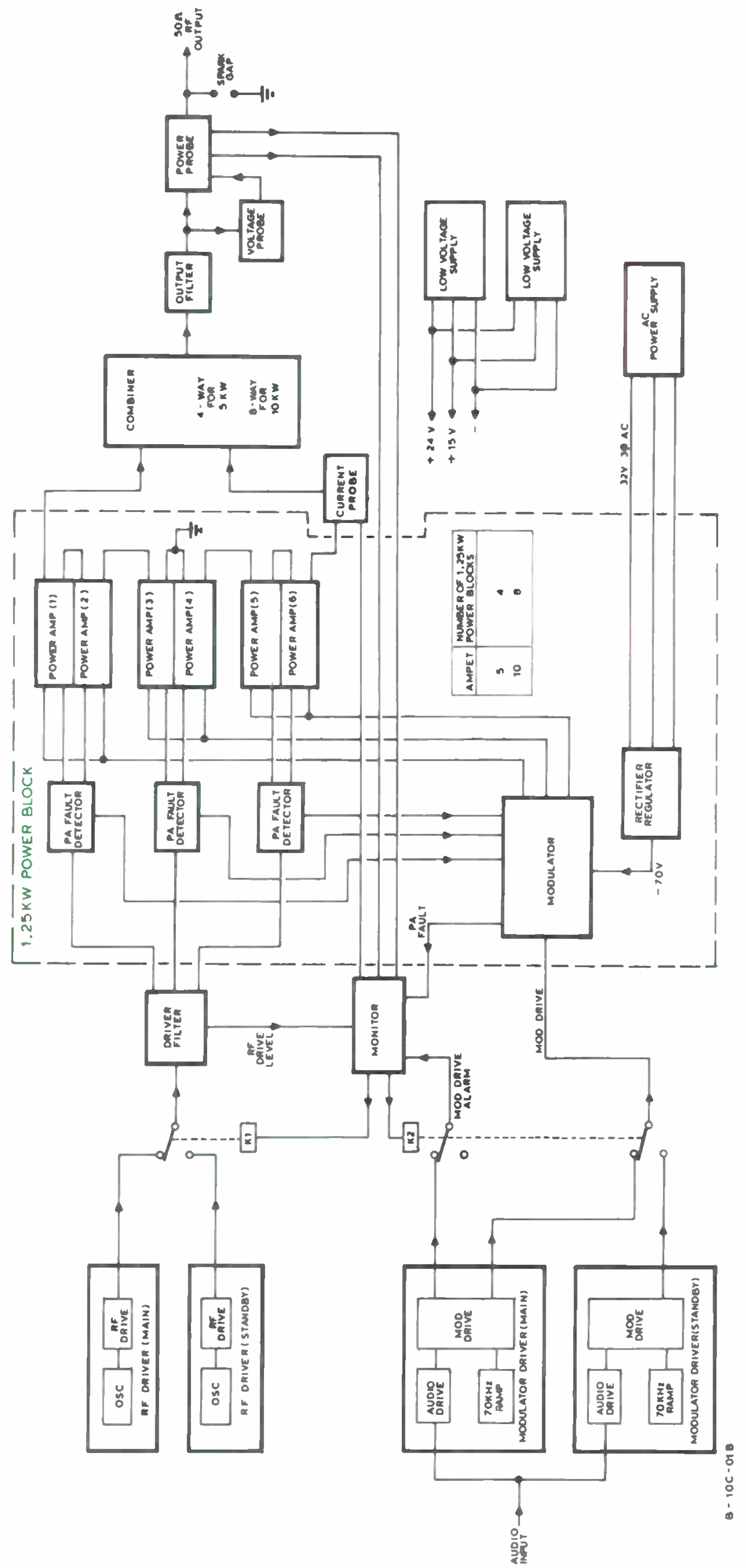


Fig. 23. Block diagram, Nautel Model AMPFET 5 and AMPFET 10, 5 & 10kW AM broadcast transmitters.

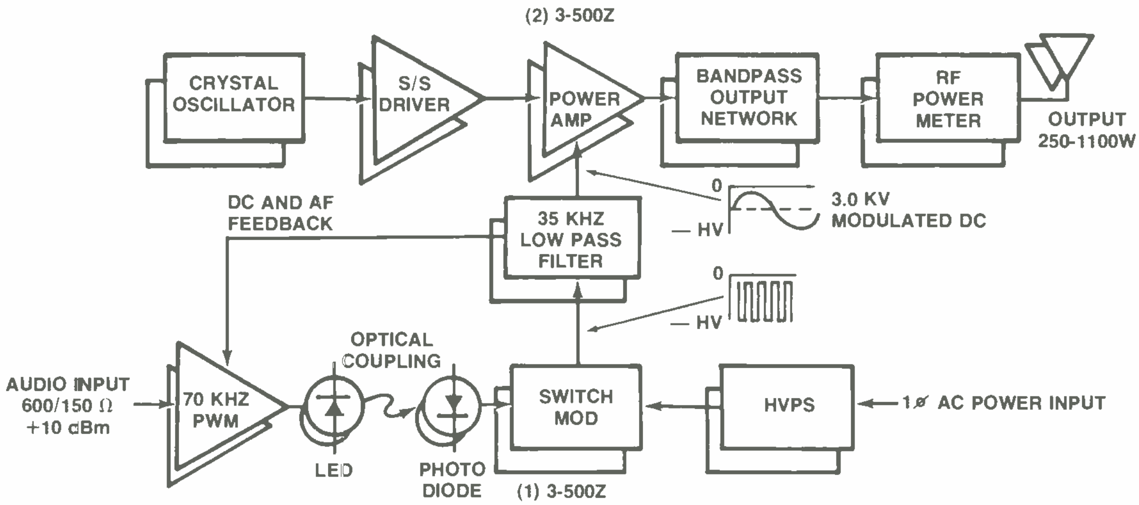


Fig. 25. Block diagram, Continental Electronics Model 314R-1 1kW AM broadcast transmitter.

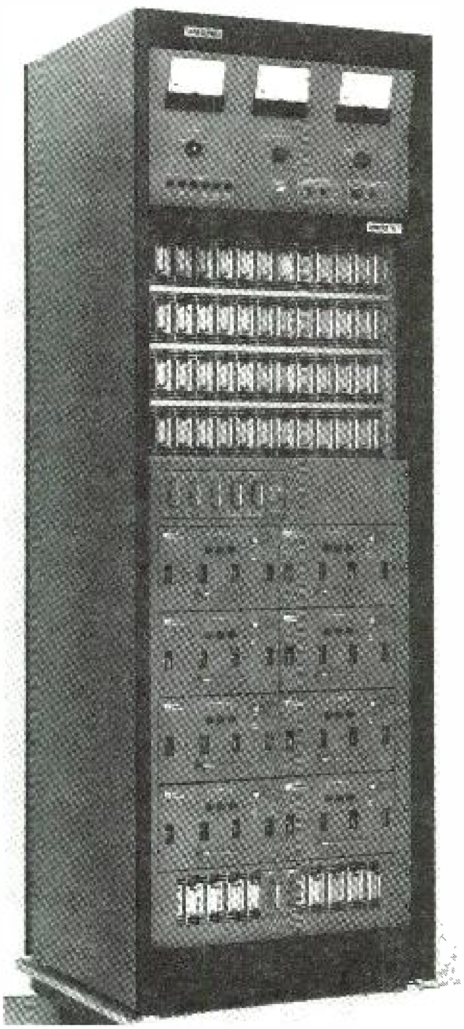


Fig. 24. NAUTEL Ampfet-10 transmitter.

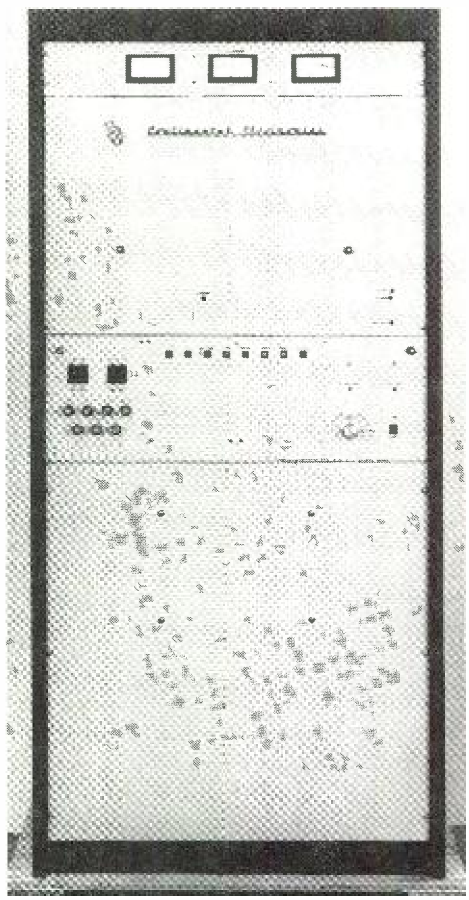


Fig. 26. Continental Electronics Model 314R-1 1kW transmitter.

Continental Electronics Model 314R-1, 1 kW Transmitter

Formerly the Collins Radio Co. Model 828C-1 before the sale of the Collins Radio Broadcast Division to Continental in 1981, the 314R-1 1kW transmitter has been in continuous production since 1977. A single tube type (3-500Z) is used three times in the transmitter, two in the high efficiency quasi-Class "D" final RF amplifier and one in the patented high efficiency series pulse width modulator (see Fig. 6(c)). All other circuitry is solid state. The transmitter has an overall efficiency of greater than 43% at any condition of modulation and a proven performance record in approximately 128 stations throughout the world. A block diagram and photograph of the model 314R-1 is shown in Fig. 25 and 26 respectively.

Continental Electronics Model 315R-1, 5kW Transmitter

Formerly the Collins Radio Co. Model 828E-1, the 315R-1 5 kW transmitter has been in continuous production since 1978. One tube type (3CX3000F7) is used in the transmitter, one in the high efficiency quasi class "D" (Class C-D) final RF amplifier and one in the patented high efficiency series pulse width modulator (See Fig. 6(c). All other circuitry is solid state. The transmitter has an overall efficiency of greater than 57% at any condition of modulation and a proven performance record in more than 300 stations throughout the world. A block diagram and photograph of Continental Electronics model 315R-1 are shown respectively in Fig. 27 and 28.

McMartin Model BA-50K

The McMartin model BA-50K employs conventional high level anode (plate) push-pull class "B" modulation. Four tubes are used, EIMAC 4CX20000B, two in the push-pull final modulator and two in the modulated final RF power amplifier. All other circuitry is solid state.The manufacturer lists an overall efficiency of better than 60% at average depths of modulation. The transmitter was first manufactured in 1982 and four are in use at this writing. A photograph of the McMartin model BA-50K is shown in Fig. 29.

Harris MW-50C

The Harris model MW-50C 50 kW transmitter is a four tube high level anode (plate) modulated transmitter employing pulse width modulation. It uses two 4CX35000C power tetrodes, one as a quasi class "D" anode modulated final RF amplifier, and one as the series pulse width modulator described above, see Fig. 6(c). Each 4CX35000C is driven by a 4CX1500A power tetrode vacuum tube. All other circuitry is solid state. The power consumption at 30% sinusoidal modulation is rated at 87 kW yielding an overall efficiency of 66% at that condition of modulation. The MW-50B has a switchable audio input linear phase (Bessel) filter to reduce transient overshoot distortion. The MW-50B is an improved version of the older MW-50A model which was first introduced in 1973. Approximately 100 MW-50 models are in service throughout the world. A block diagram and photograph of the Harris MW-50B are shown in Fig. 30 and 31 respectively.

Continental Electronics Model 317C-2

The Continental Electronics model 317C-2 employs the patented high efficiency and reliable screen/impedance system of modulation. Having almost linear audio phase response to 30 kHz, the 317C-2 can produce high modulation density levels without transient overshoot distortion and with only approximately one-half the high peak anode voltages encountered with high level anode systems of modulation. The overall efficiency of the 317C-2 is better than 60% at any condition of modulation. First manufactured in 1968, more than 250 units are in operation throughout the world at this writing. Fig. 32 and 33 respectively show the block diagram, and a photograph of the Continental Electronics model 317C-2 transmitter.

Continental Electronics Model 419F 250 Kilowatt Shortwave Transmitter

The model 419F transmitter uses conventional class "B" high level anode (plate) modulation. The final RF amplifier uses a single EIMAC vapour cooled 4CV250000B power tetrode and the modulators employ two EIMAC vapour cooled 4CV100000C power tetrodes in a special push-pull arrangement that allows very low modulator static currents for highest possible program efficiency. The modulation transformer is an oil-filled unit designed to allow the RFPA dc anode current to flow directly through its secondary without saturation, thus eliminating the need for high pass coupling of the audio signal from the modulator output to the RFPA anode. The final RFPA is driven by an EIMAC 3CW20000A7 power triode. All other circuitry is solid state. Overall efficiency of the 419F is better than 60% at any condition of program modulation. The 419F is capable of ten channel preset tuning by local or remote control without operator intervention. First manufactured in 1980, there are twelve units in operation in Europe and the Middle East at this writing. The 419F is the largest of a family of similarly designed 50 and 100 kW transmitters, Continental models 417C and 418D.

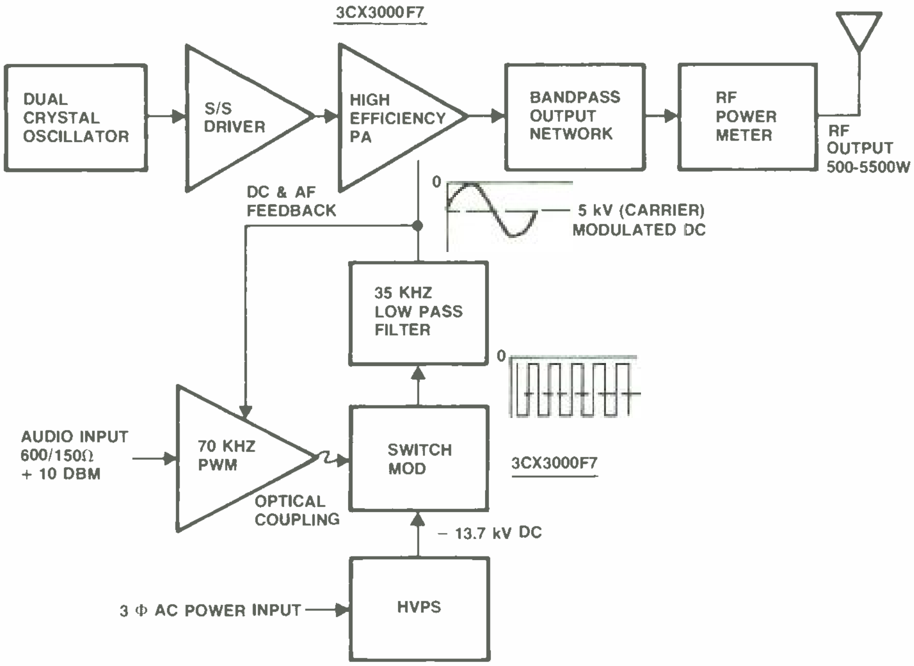


Fig. 27. Block diagram, Continental Electronics Model 315R-1 5kW AM broadcast transmitter.

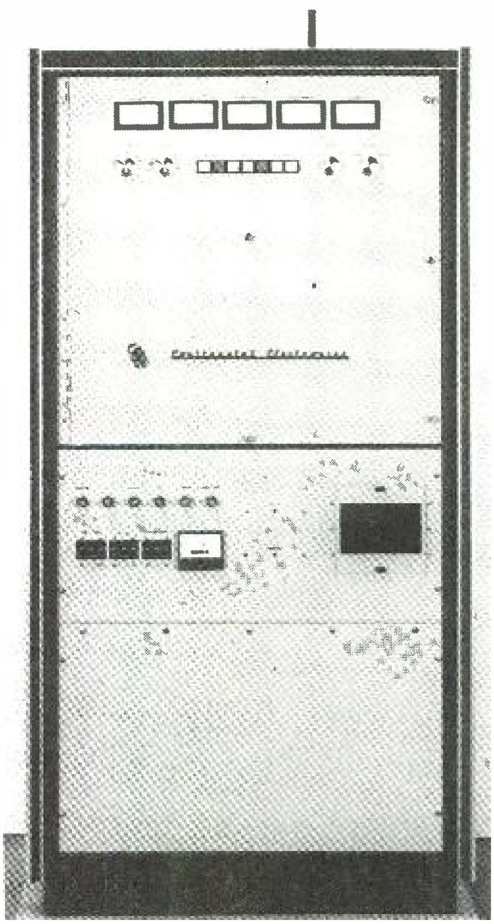


Fig. 28. Continental Electronics Model 315R-1.

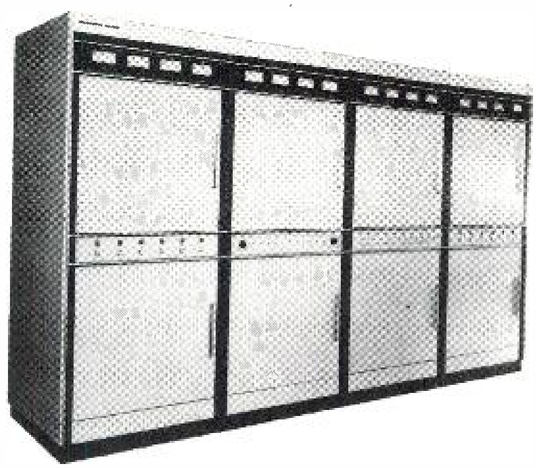


Fig. 29. McMartin Model BA-50k, 50kW AM broadcast transmitter.

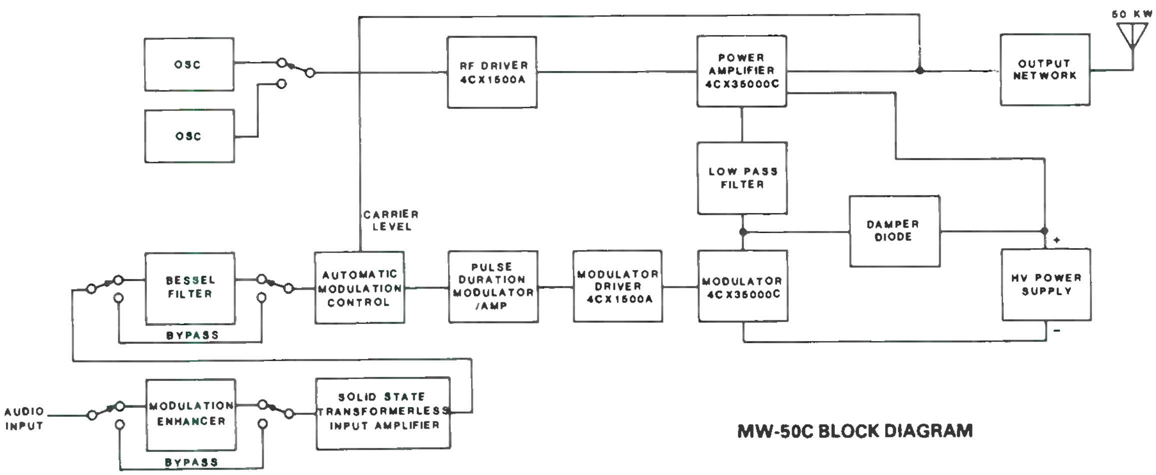


Fig. 30. Block diagram, Harris Model MW50C, 50kW broadcast transmitter.

Fig. 31. Harris Model MW 50C transmitter.

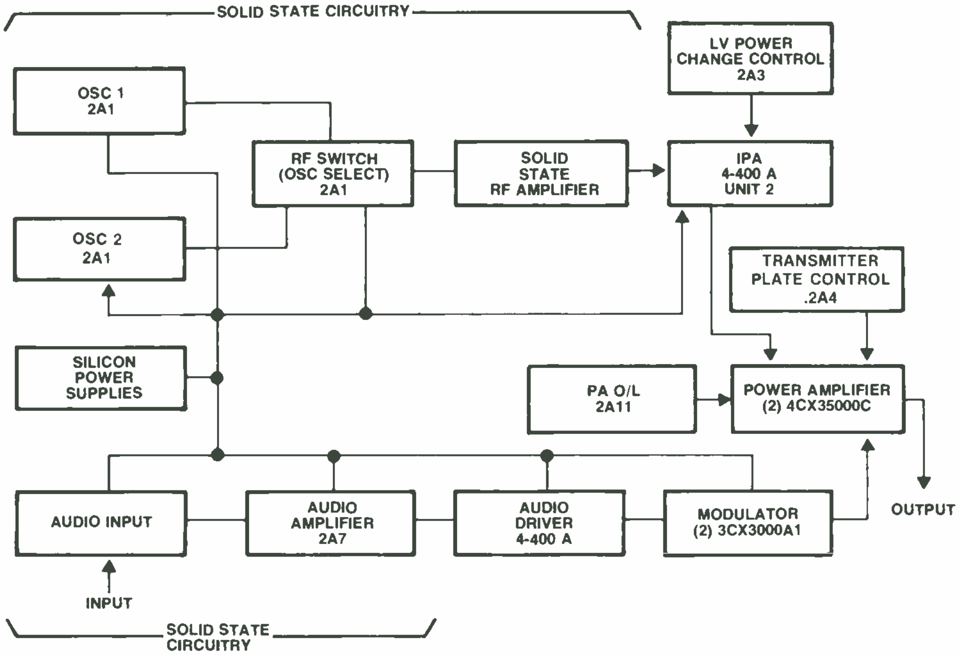


Fig. 32. Block diagram, Continental Electronics Model 317C-2 50kW AM broadcast transmitter.

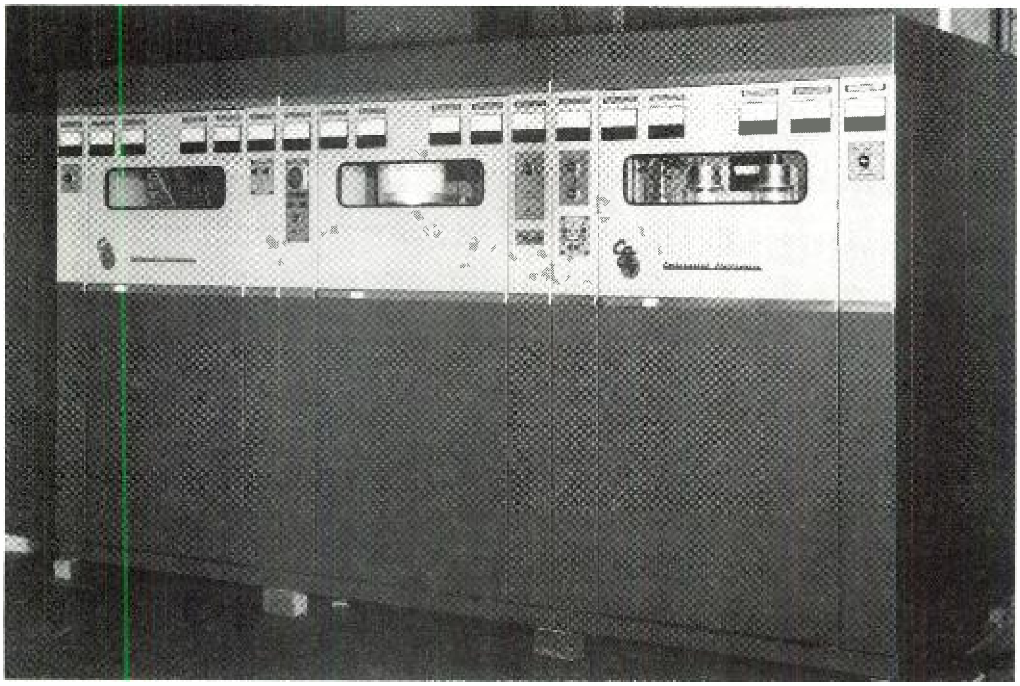


Fig. 33. Continental Electronics Model 317C-2.

3.2

AM Stereo Systems

An Overview

Edmund A. Williams
Staff Engineer
Department of Science and Technology
National Association of Broadcasters
Washington, DC

INTRODUCTION

The idea of stereophonic transmission on an AM broadcast station is not new or remarkable. While AM is an efficient mode of transmission providing the maximum amount of information in a minimum of spectrum, additional spectrum economies can be achieved which can be used for stereo information.

AM stereo was approved by the Federal Communications Commission in 1981 in a Report and Order which introduced a new concept in the process of selection of new broadcast transmission standards—selection by the "marketplace". Because no single system of the several offered by various manufacturers was approved by the Commission it was left to broadcasters to decide which system to implement. As a result neither AM broadcaster nor receiver manufacturer were willing make decisions on their own but instead preferred to wait for the other to act. This has caused the implementation of AM stereo to be delayed while each group carefully tests the stereo waters. Proponents of AM stereo systems developed two distinctly different means of transmitting two audio channels on a single AM broadcast carrier. Both are capable of producing good high quality stereo sound and cause no significant increase in interference levels. Both are compatible with monophonic receivers. However, the two systems are not compatible with each other and require entirely different means to detect the signals. One approach is to independently modulate the two sidebands with the left (L) channel on one sideband and the right (R) channel on the other sideband. The second is to modulate the main AM carrier with the sum (L + R) information and to use any one of several variations of quadrature or phase modulation for the difference (L-R) audio channel. Best receiving performance will be achieved by designing the detector to a specific modulation format. However, a single compromise detector may be used to produce adequate stereo performance from any of the quadrature or phase modulation techniques. As a result receiver designs will evolve around one or two detection methods while broadcasters will select between four and five transmission systems.

The two forms of modulation for AM stereo each have distinct advantages and disadvantages as described by the proponents in the following chapter. Each AM station, in addition to deciding to make the transition to stereophonic transmission, must also decide upon which of the several transmission systems to employ. This decision will be based on economic as well as technical factors. Obviously the cost of modifying existing transmission facilities for stereo will weigh heavily for the lowest cost system. However, the technical aspects of selecting an AM stereo system must necessarily include the range of the station, the nature of the programming, proliferation of receivers able to detect the transmissions, the ability of the manufacturer to support the equipment and, of course, the technical quality of the signal itself. Other factors, such as sensitivity to transmitter and antenna system tuning tolerances, ease of conducting a "proof", compatibility with directional antenna systems and general maintenance must also be considered when selecting the transmission system.

It is not the intent of this chapter to add confusion to the selection process by interjecting too many selection criteria into that process. Rather, it is to point out the fact that broadcasting is no longer simple AM and FM. Technological advances in communication during the past two decades have proven that there may not be a single "best" system for anything in the future. Instead there will be more emphasis on permitting the marketplace to have a hand in the process in addition to the need for broadcasters to adopt new transmission modes according to their business instincts. The marketplace will react by adapting to the new transmission systems either by accommodating several with automatic receiver switching or evolving toward the lower cost or higher quality system. The broadcaster will influence the process by selecting and advocating, at the earliest opportunity, the approach which will provide the greatest benefit to the audience.

The marketplace approach to the selection of AM stereo is a clear indication of trends broadcasting will take in the future. The extent to which broadcasters lead the way will determine the direction of the trend of that marketplace.

3.2

AM Stereo Systems

Motorola C-Quam AM Stereo System

Chris Payne
Motorola Corporation
Schaumburg, Illinois

Many attempts have been made to develop an AM stereo system using quadrature modulation. However, the basic problem with quadrature systems has always been monaural compatibility; that is, high levels of distortion in typical AM receivers. Previous attempts to solve the incompatibility problem have resulted in poor coverage, reduced stereophonic information, poor separation, or a combination of these ill effects. The Motorola C-Quam (Compatible Quadrature) system solves the usual incompatibility problem without incurring losses in range, separation, or distortion.

AUDIO PERFORMANCE

The Motorola system is, in concept, without flaws. This means that the theory tells us that the system is capable of zero distortion, infinite frequency response, and infinite separation. In practice, the performance is limited primarily by the fidelity of the broadcast transmitter and receiver. In the laboratory, frequency response of 20 Hz to 15 kHz, separation of more than 40 dB and distortions of less than 0.3 percent, have been readily attained. In the field, performance has exceeded FM stereo proofs and won the support of major receiver manufacturers.

AUDIO PROCESSING AND PREEMPHASIS

Actual broadcasting experience with the Motorola system has proven that high degrees of audio processing and preemphasis may be used without difficulty. Stations, in fact, report that the received monophonic sound is as good or better when operating with Motorola AM Stereo as with previous monophonic operation.

SPECTRUM

The Motorola AM Stereo system, when operating with typical state-of-the-art audio processing equipment with preemphasis, occupies essentially the same spectrum as does a monophonic station with the same processing. The Motorola exciter has been FCC-type accepted for broadcast use, without any special stereo filter restrictions.

THE C-QUAM TRANSMITTER

The C-Quam transmitter is shown in Figure 1. Note that pure quadrature is generated by taking $L+R$ and $L-R$ and modulating two balanced

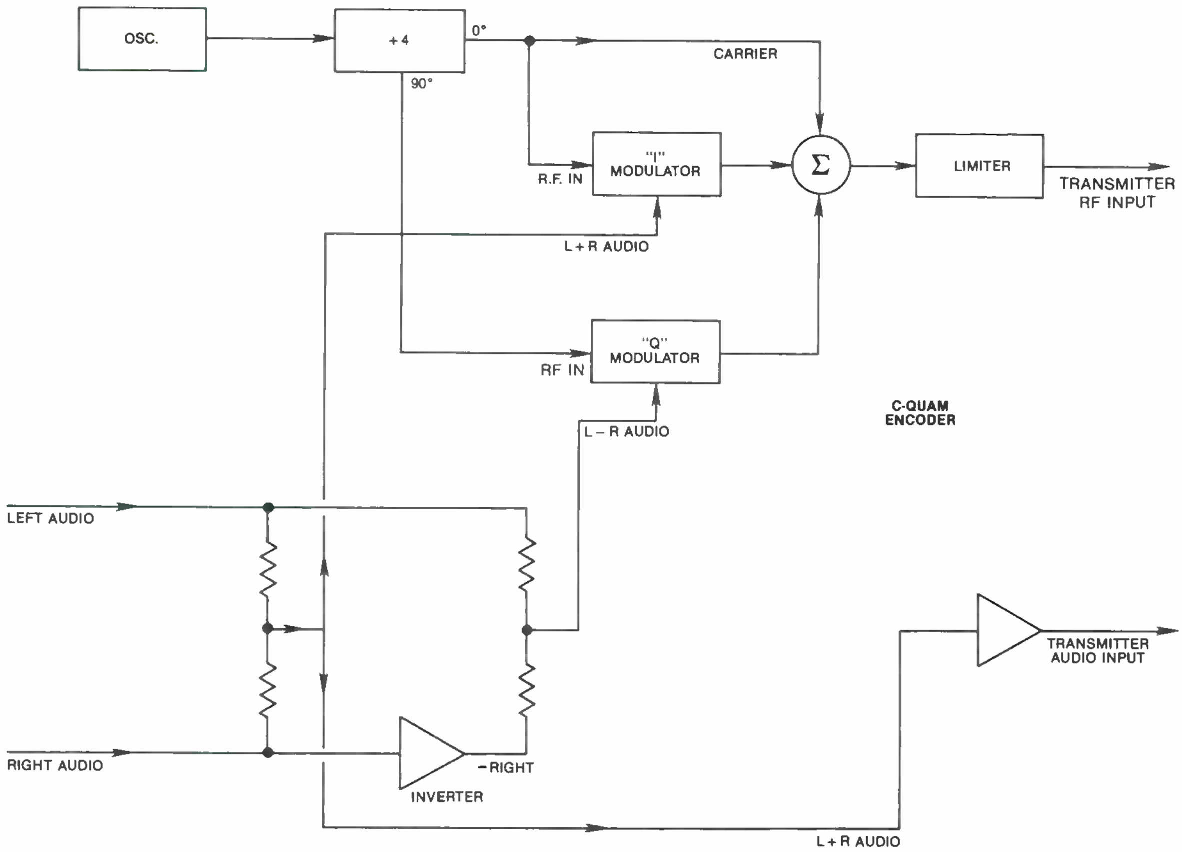


Fig. 1. C-Quam transmitter.

modulators fed with RF signals out of phase by 90 degrees. In this case the 90 degrees phase shift is derived by using a Johnson counter which divides an input frequency (four times station carrier frequency) by four and automatically provides digital signals precisely 90 degrees out of phase for the balanced modulators. The carrier is inserted directly from the Johnson counter. At the output of the summing network, the result is a pure quadrature AM stereo signal. From there it is passed through a limiter which strips the incompatible AM components from the signal. The output of the limiter is amplified and sent to the broadcast transmitter in place of the crystal oscillator.

The left and right audio signals are precisely added and sent as normal and compatible $L+R$ to the audio input terminals of the broadcast transmitter. That's the Motorola C-Quam encoder.

DECODING C-QUAM

C-Quam is decoded by simply converting the broadcast signal (which is already close to a quadrature signal) to pure quadrature and then using a quadrature detector to extract and $L-R$. Refer to Figure 2. Note that the demodulator contains a section which is a pure quadrature demodulator. In order to prepare the received signal for the quadrature demodulator, it has to be converted from the envelope detector compatible signal that is broadcast to the original quadrature signal that was not envelope detector compatible. This is done by demodulating the broadcast signal two ways; with an envelope detector, and with an "I" detector. The two signals are compared and the resultant error signal is used to gain modulate the input of the "I" and "Q" demodulators.

When the transmitted signal is $L+R$ (monaural, no stereo) the transmitted signal is pure AM or only "I" sidebands. In this case the envelope detector and the "I" demodulator see the same thing. There is no error signal, the input modulator does nothing and the signal passes through without change. However, when a left or right only signal is transmitted, both AM and PM is transmitted and the input signal is shifted in phase to the "I" demodulator and loses some of its "I" amplitude. The envelope detector sees no difference in the AM because of the phase modulation, and when the envelope detector and the "I" demodulator are compared, there is an error signal. The error signal pushes up the input level to the detector. This makes the input signal to the "I" and "Q" demodulators look like a pure quadrature signal and the audio output gives a perfect and $L-R$ signal. The demodulator output is combined with the envelope detector output in a matrix to give left and right audio out.

SYSTEM PERFORMANCE UNDER HEAVY MODULATION

There are many advantages of the Motorola system. One is its performance under 100 percent negative amplitude modulation conditions. When the carrier momontarily goes to zero, as in 100 percent negative modulation, the output of the envelope detector becomes zero. Because of the action of the comparator and inverse modulator, the output of the "I" demodulator goes to zero. Simultaneously, therefore, the output of the "Q" demodulator also is forced to go to zero. This means that there will be no severe noise or interference popping from the stereo channel under 100 percent negative amplitude modulation.

THEORETICAL AUDIO PERFORMANCE

There is no theoretical limit to the audio frequency response, to the lower limit of distortion or to the separation of the Motorola system. All of these characteristics are primarily limited by the performance of the encoder, AM transmitter, and modulation monitor or receiver. The only other limit would be due to the bandwidth of emissions under stereo modulation conditions with single tone modulations above 7500 Hz. It is theoretically possible to force the system with tone modulations, to exceed the FCC bandwidth limits permitted for AM stations by a few dB. However, under the most heavily processed and preemphasized stereo program conditions, the emissions are well below FCC requirements for bandwidth.

ACTUAL STEREO PERFORMANCE

The actual measured performance of the Motorola system connected to a typical plate modulated AM broadcast transmitter is excellent, nearly that of FM stereo. The frequency response of left or right only transmissions is basically the

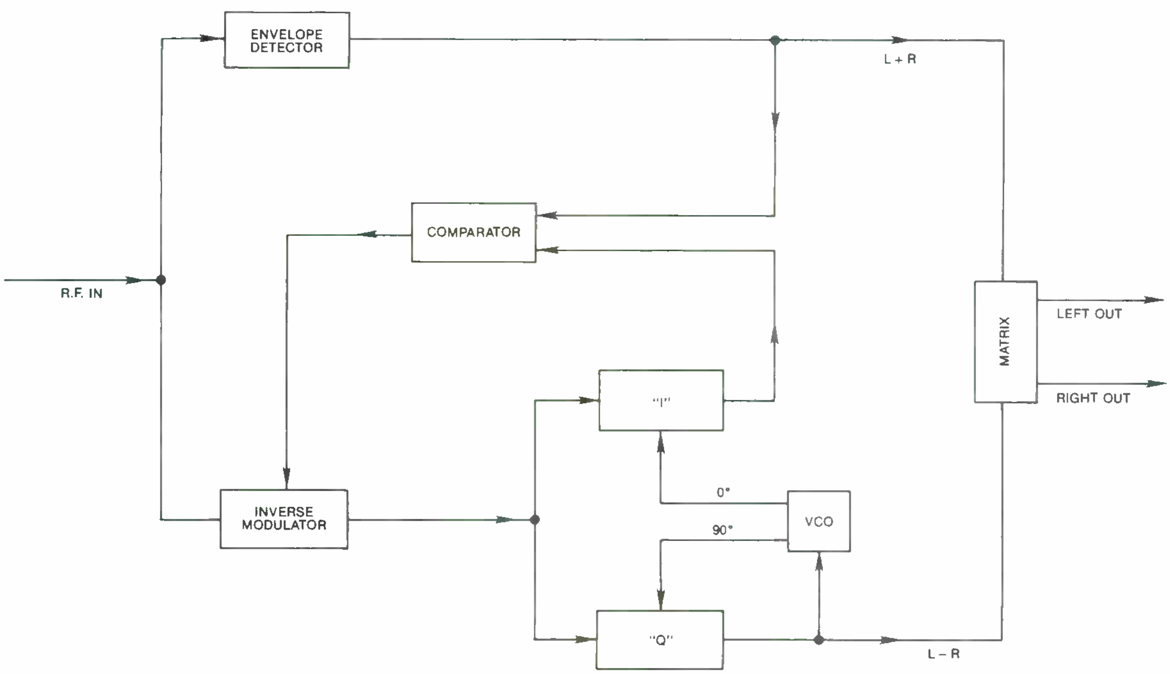


Fig. 2. C-Quam decoder.

same as the AM response of the transmitter. (In the case of tests made at WIRE in Indianapolis, about 1 dB down from 30 Hz to 12 kHz). The separation was held to better than 30 dB from 100 Hz to 7.5 kHz and within 20 dB from 50 to 10 kHz. Distortion of the left or right only transmissions were below 1.5 percent from about 80 Hz to 7.5 kHz. This data is taken from actual measurements made from an AM stereo modulations monitor.

RECEIVER AUDIO AND COVERAGE

Because the Motorola AM Stereo system acts like a full quadrature system under normal stereo conditions, its signal to noise characteristics are excellent. Under typical stereo programming, the increase in noise is estimated to be about 1.5 dB which is nearly imperceptible. In other words, the stereo coverage of the Motorola system is essentially the same as the monaural AM coverage.

The audio fidelity that can be expected from AM stereo radios depends upon the receiver design. In the past, receiver manufacturers were generally not willing to make AM radios with better fidelity because they felt that there was no demand for such equipment. With AM stereo, the ability to offer "stereo" means a much more attractive product which will allow a high price. Therefore, the AM stereo radio designer will have a much wider latitude in designing the receiver selectively and will be more able to achieve higher fidelity while holding down the susceptibility to adjacent channel interference. With slightly more expensive IF selectivity, the receiver designer can obtain audio response well beyond 5 kHz while still rejecting the adjacent channel interference. Two bandwith receivers, variable bandwidth receivers, and fixed bandwidth receivers to 7 kHz may be excepted for AM stereo.

CONCLUSION

The Motorola AM Stereo system is the system designed to meet the needs of AM radio. In other words, designed to perform well yet to be received inexpensively so that millions of radio listeners can enjoy AM stereo. While the design is adaptable for cost effective radios, it is also the best overall technical performer for the broadcaster because it is capable of FM-like audio performance with full amplitude modulation capability and stereo coverage essentially the same as monaural coverage.

3.2

AM Stereo Systems

The Harris Linear AM Stereo System

David Hershberger
Harris Corporation
Quincy, Illinois

INTRODUCTION

Harris promotes linear quadrature modulation and synchronous detection for AM stereo. Although envelope detectors produce some distortion when receiving linear quadrature modulation, the distortion is almost entirely second harmonic, which is seldom audible and not objectionable. The Harris AM stereo sytsem incorporates special patent pending audio processing techniques which ensure that envelope detector distortion will not be objectionable to the listener. Major broadcasters using the Harris linear system are very pleased with both stereo performance and monophonic compatibility.

AM stereo can appear deceptively simple to the casual investigator. Although amplitude modulation is a "low-technology" mode, AM stereo, and especially the selection of an AM stereo standard, is a complex issue requiring special insight. First impressions are usually wrong in AM stereo.

A thorough investigation of AM stereo, including a fundamental evaluation of compatibility, will show that linear systems are best suited for AM stereo.

The primary reason for AM stereo is to allow AM stations to compete more effectively with FM by improving the technical quality of the AM service. Although AM does have a higher noise level than FM, the audio quality limitation is primarily due to poor radios. Most AM stations transmit audio to beyond 10 kHz, but most radios have a frequency response of only 2 to 3 kHz. Virtually all radios use the antiquated envelope (diode) detector to recover modulation. Envelope detectors are inherently prone to the production of various types of distortion and other undesireable effects. AM radios can be vastly improved by extending the bandwidth for better frequency response, and by replacing the envelope detector with a synchronous detector for reduced distortion.

NONLINEAR SYSTEMS

A linear system is one in which superposition applies; that is, the system response to several inputs applied simultaneously is the same as the sum of the system responses when the inputs are applied individually. A nonlinear system is one in which superposition does not apply.

Proponents of nonlinear modulation use it in an attempt to obtain compatibility with envelope detectors. If receiver bandwidth were infinite, there would be perfect compatibility.

One way of analyzing nonlinear modulation is as a predistorted system. Because the envelope detector is also a nonlinear device, nonlinear systems must predistort the upper and lower sidebands of an AM stereo signal in order to force the envelope to be $L+R$.

This is equivalent to transmitting distortion 180 degrees out of phase with the distortion the envelope detector generates. Although such a signal is perfectly compatible with envelope detectors, it is unfortunately incompatible with another circuit found in all AM radios: the bandpass filter.

It can be shown that nonlinear systems have an infinite number of sidebands, and in practice three or four orders of these sidebands are significant. The high order or predistortion sidebands must all be present at the receiver's detector for proper operation and compatibility. But by the time a nonlinear AM stereo signal passes through the IF bandpass filter of a typical narrow bandwidth radio, the relative amplitude and phase relationships of the required predistortion sidebands may be altered such that distortion is unacceptable, even though the radio may be correctly center-tuned. Truncation of the spectrum by the bandpass filter's amplitude characteristic, and phase distortion caused by the filter's group delay characteristic, will cause the predistortion sidebands to arrive at the wrong amplitude and at the wrong time to "cancel" the envelope detector distortion.

There are two significant types of compatibility distortion with nonlinear systems: harmonic distortion and difference tone intermodulation distortion. Of the two, the difference tone distortion is more objectionable to listeners.

It may be argued that these types of distortion are of little significance since program material generally contains little high frequency energy. That argument, however, is invalid because of the general use of preemphasis on AM. To partially compensate for the limited frequency response of AM radios, most AM broadcasters boost the higher modulating frequencies. Since the improvement in received signal quality can be quite dramatic, broadcasters will be unwilling to sacrifice the use of preemphasis.

Preemphasis, which frequently boosts the higher frequencies as much as 20 to 30 dB, severely aggravates the filtering-related compatibility problem of nonlinear signals.

LINEAR SYSTEMS

Linear systems are those for which superposition applies. Two important characteristics of linear systems are:

1. Only one set of sidebands is produced—bandwidth is the same as mono.
2. Linear product detection is both $L+R$ and $L-R$ channels means that there is no "noise burst" problem whatever.

Linear quadrature modulation is simply expressed as:

$$S(t) = [1 + L(t) + R(t)]\sin(wt) + [L(t) - R(t)]\cos(wt) \quad [1]$$

Where:

$S(t)$ = stereo AM signal
$L(t)$ = left channel modulating signal
$R(t)$ = right channel modulating signal
w = carrier (radian) frequency
t = time

The stereo signal consists of a monophonic sum signal (the sine term) with the linear addition of a double sideband suppressed carrier left minus right signal in quadrature with the carrier (the cosine term). Since both signals (sine and cosine terms) have only one set of sidebands, and since addition is a linear process, the resulting stereo signal also has only one set of sidebands—no new frequencies are created by adding the two signal components.

Linear quadrature modulation is highly desirable for AM stereo broadcast application because of its many advantages over nonlinear systems.

1. Same RF bandwidth as mono
2. No distortion from spectrum truncation
3. Compatible with preemphasis
4. Compatible with synchronous detection
5. Compatible with envelope detection (with audio processing)

Synchronous detectors, in turn offer the following advantages over envelope detectors:

1. No distortion due to receiver mistuning
2. No distortion due to selection fading or AM multipath
3. No distortion from directional transmitting antennas (far-field overmodulation)
4. No distortion from modulation overshoots
5. No distortion due to co-channel interference
6. No inherent cross modulation
7. Some reduction of impulse noise.

COMPATIBILITY WITH HUMAN HEARING

Compatibility is without doubt the most complex AM stereo requirement. To determine the extent to which a system is compatible, it is necessary to evaluate the end result, which is the listenability of a system on a narrowband envelope detector radio, transmitting real program material to real human beings. The final test of compatibility comes when the radio station transmitting AM stereo goes into its ratings period.

Other tests, such as transmitting tones through wideband envelope detectors to instruments, while valuable in the quantitative sense, are not necessarily measurements of compatibility with human hearing.

It is in the determination of compatibility that errors in system evaluation are most likely to be made. If compatibility is not thoroughly investigated, it is very likely that one could conclude that a compatible system should be rejected in favor of another system that is actually incompatible with existing radios.

When defining compatibility, one's first impulse would be to state that a compatible system should have low distortion. Distortion usually is taken to mean total harmonic distortion or THD. This is the pitfall to be avoided—a simplistic view of distortion.

The point is this: the criteria for the evaluation of distortion should accurately represent the sensitivities of the human ear. For example, second harmonic distortion is only objectionable when very large. On the other hand, difference tone intermodulation distortion is quite objectionable to the ear and should weigh heavily in any system evaluation.

Linear systems have excellent potential in every area except one: there is an apparent lack of compatibility with envelope detectors. The envelope of eq. [1] is:

$$e(t) = \left([1 + L(t) + R(t)]^2 + [L(t) - R(t)]^2 \right)^{1/2}$$

Which is a distorted version of the desired signal, $1 + L(t) + R(t)$. If a simple total harmonic distortion (THD) calculation or measurement were made, the value would appear to be too high. For example, envelope THD values for several single channel modulation levels are given in Table 1.

TABLE 1. Envelope distortion vs. single channel modulation.

Left Channel Modulation	THD
75%	28.9%
50	16.1
25	6.7

But what simple THD values do not convey is that the distortion is primarily second harmonic, which makes the distortion relatively innocuous.

Second harmonic is the least objectionable of the harmonics. In general, even-order harmonics, such as second, fourth, sixth, etc., are less objectionable than odd-order harmonics such as third, fifth, seventh, etc. Furthermore, the lower the order of the harmonic, the less objectionable than is, say eleventh harmonic.

These general characteristics of human hearing are in turn due to the physiology of the ear/brain combination, and how we hear sounds. The ear, for instance, has a large amount of second harmonic built in. If two flutes are played at moderate volume, the ear will create a third tone due to the difference in frequencies between the two flutes. In fact, some pieces of music take advantage of the phantom difference tone created by the ear. Several Telemann flute sonatas created a difference tone part which harmonizes with the "real" parts. A more familiar example of music with a third part created by the ear is "Yankee Doodle."

In fact, there is some indication that second harmonic distortion is in some instances actually desirable. One example is a popular piece of studio equipment which is essentially a second harmonic generator. This unit is widely used by recording studios to process music for increased appeal.

The addition of second harmonic, instead of sounding "distorted," sounds like increased treble response. While conventional "equalization" or high frequency boost also increases the noise, artifical highs from a second harmonic generator will not boost the noise. In summary, adding second harmonic sounds basically like treble boost, but without a background hiss boost.

CONTROLLED—COMPATIBILITY LINEAR QUADRATURE

It has been the finding of Harris and other investigators that linear quadrature modulation is "almost compatible." If something minor could be done to obtain compatibility without sacrificing the linearity of the system, then a truly superior AM stereo system would be feasible.

Harris has developed a special audio processing technique which ensures compatibility with envelope detectors, while not significantly affecting the stereo signal. The audio processing (compatibility control) works by determining the distortion which would occur in an envelope detector; when the envelope distortion exceeds an acceptable level, single channel negative modulation constraints are reduced. Normally, single channel left or right signals are allowed to modulate up to +80% and −80%. When envelope detectors produce excessive distortion, the negative modulation constraint for single channel signals is reduced from 80% to something less; typically −70% or −60%. The negative constraint is reduced until the envelope detector distortion is within acceptable limits.

A block diagram of the AM stereo signal generator is shown in Fig. 1.

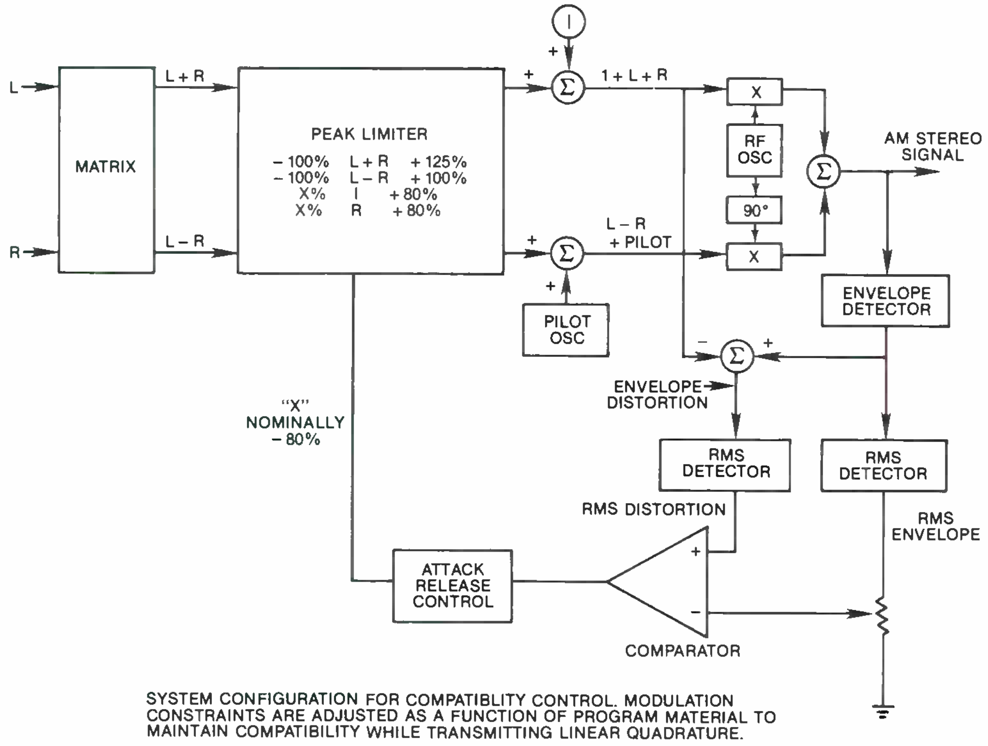


Fig. 1. Block diagram of the AM stereo signal generator.

The time constant associated with the compatibility control processor is somewhat slow, in the hundreds of milliseconds. The deliberately slow time constant reflects the design philosophy of having a system which is designed for the human ear rather than for the sake of instrumentation. In order for the ear to detect low orders of distortion (such as second order), the distortion must be present for some time. Unlike crossover distortion, for example, which can be detected immediately, quadrature envelope distortion must be present for several hundred milliseconds before it is detectable. Accordingly, the audio processing, whose purpose it is to model the ear, is made to have a similarly slow time constant.

Broadcasters using the Harris linear quadrature system for AM stereo are unanimous in finding both excellent stereo performance and good quality compatible reception on monophonic radios. The "bottom-line" proof of the system's compatibility lies in the ability to satisfy program directors, and in station market share ratings.

CONCLUSION

Linear quadrature modulation is very appealing for application in AM stereo. Many early attempts at using quadrature modulation compromised the benefits of the system. Some systems simply reduced the amplitude of the $L-R$ channel, thereby incurring a signal-to-noise ratio penalty. Other systems sacrificed to be $1+L+R$. These nonlinear systems occupy excessive bandwidth, and in spite of the fact that the transmitted envelope is $1+L+R$, radio IF bandpass filtering causes the nonlinear system to generate distortion at the radio's envelope detector. Nonlinear systems, in the final analysis, have serious compatibility problems.

The supposed "incompatibility" of linear quadrature modulation is not a real problem. When human beings are used to determine compatibility, minor audio processing makes the use of linear quadrature modulation an ideal solution to the AM stereo problem.

Broadcasters using the Harris linear system are pleased with both stereo and mono performance, while broadcasters using nonlinear systems often voice dissatisfaction with monophonic compatibility, especially when preemphasis is used.

The use of linear quadrature modulation will encourage the development and marketing of long-overdue improvements to the AM sections of radios: synchronous detection and wider IF bandwidths, for less distortion and wider frequency response. These improvements and the use of linear quadrature modulation are most consistent with the goals of the AM broadcaster and the consumer: to make AM stereo sound competitive with FM stereo.

AM Stereo Systems

AM Stereo System Transmission and Reception

Leonard R. Kahn
Kahn Communications, Inc.
Garden City, New York

The AM Stereo system proposed by Kahn Communications and the Hazeltine Corporation is an Independent Sideband System wherein the lower sideband represents the right stereo channel. Thus, this system is essentially a frequency separation system as distinguished from the other major systems which are essentially phase separation systems in that the stereo separation is achieved by detecting the phase difference of stereo components.

It is the author's belief that this distinction; i.e., frequency separation vs phase separation, is the basis for the differences in performance of the systems especially under adverse conditions such as co-channel and adjacent channel interference, multipath conditions such as skywave/ground wave interference, sensitivity to transmitting and receiving antennas, and capability of using advanced receiver technology such as Asymmetrical Sideband Selectivity.

While the means for generating an ISB wave is, for the block shown below, phase sensitive, nevertheless, once the independent sideband wave is generated stereo separation is insensitive to phase as the wave is passed through the transmitting antenna, the propagation path, the receiving antenna, the receiver's IF and RF circuits, and the input circuit feeding the stereo demodulator.

TRANSMITTING EQUIPMENT

Please see Fig. 1 which is a simplified block diagram of the type of AM Stereo generator used in the STR-77 unit. The left and right inputs are summed and fed through a constant amplitude, phase difference network. The output of this network feeds an audio frequency amplifier which produces an output suitable for modulating conventional AM transmitters.

The difference between the L and R components is fed to a companion phase difference network which assures a quadrature relationship between the output of the sum circuit and the difference circuit. This output is then fed to a phase modulator. The resulting phase modulated wave is frequency translated to the carrier frequency of the station. An RF amplifier is used to raise the level to 3 watts, sufficient for exciting most AM transmitters. This phase modulated wave is then envelope modulated in the station's transmitter to produce an independent sideband wave wherein the lower sideband carries the left information and the upper sideband carries the right information.

In a practical configuration, circuitry is provided for insuring proper time delay so that the AM and PM components arrive at the transmitter's modulated stage coincidentally. Additional

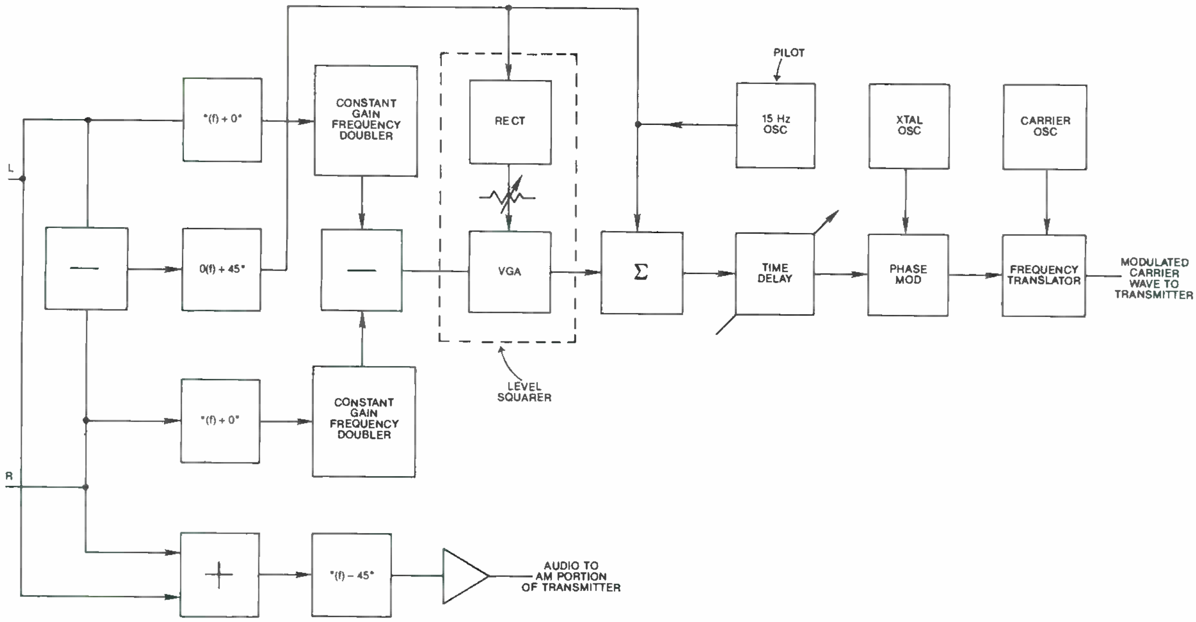


Fig. 1. Single side band AM stereo system.

circuitry is shown in Fig. 1 for producing a second order $L-R$ wave component which is added to the fundamental $L-R$ audio wave fed to the phase modulator. This second order component minimizes bandwidth and allows low distortion reception.

Fig. 2 shows a simplified phasor sketch showing how the first order sideband components line up so that the lower sideband represents the left stereo information and the upper sideband represents the right stereo information. The following equation is a descriptive equation of the modulated wave:

$$
\begin{aligned}
e_s = E_c\ [\ 1 + m_L\ L(t)\ e^{J}\ \pi/4 + m_R\ R(t)\ e^{J}\ \pi/4\ ] \\
\times \cos\ [w_c t + m_L\ L(t)\ e^{-J}\ \pi/4 - m_R R(t)\ e^{-J}\pi/4 \\
- .532\ [\ m_L(L(t)\ )^2 - m_R\ (R(t)\ )^2\ ] \\
\times\ k/T\ {}_0\!\int^T |m_L\ L(t) - m_R\ R(t)|\ dt + \lambda \sin 2\pi\ 15t\ ]
\end{aligned}
\qquad [1]
$$

RECEIVER

Fig. 3 is a simplified block diagram of a receiver suitable for demodulating an independent sideband stereo wave. Details concerning the carrier track circuitry and the inverse modulation circuit is contained in the referenced patents and Hazeltine's report.*

ASYMMETRICAL SIDEBAND SELECTIVITY

The ideal receiver for the instant stereo system provides enhanced selectivity for adjacent channel interference by taking advantage of the fact that adjacent channel interference is naturally separated so that lower channel interference falls solely in the left stereo channel and higher channel interference falls in the right stereo channel. This natural characteristic is the basis for the "cocktail party effect" which causes interference to fall full to the right or full to the left edge of the stereo stage according to whether the interference is above or below the desired channel.

In contrast, phase separation systems produce interference in both the left and right channels, causing the interference to fall "on stage." The natural separation of the interference on one channel or the other in the present system also allows the use of a new form of selectivity which is called Asymmetrical Sideband Selectivity.

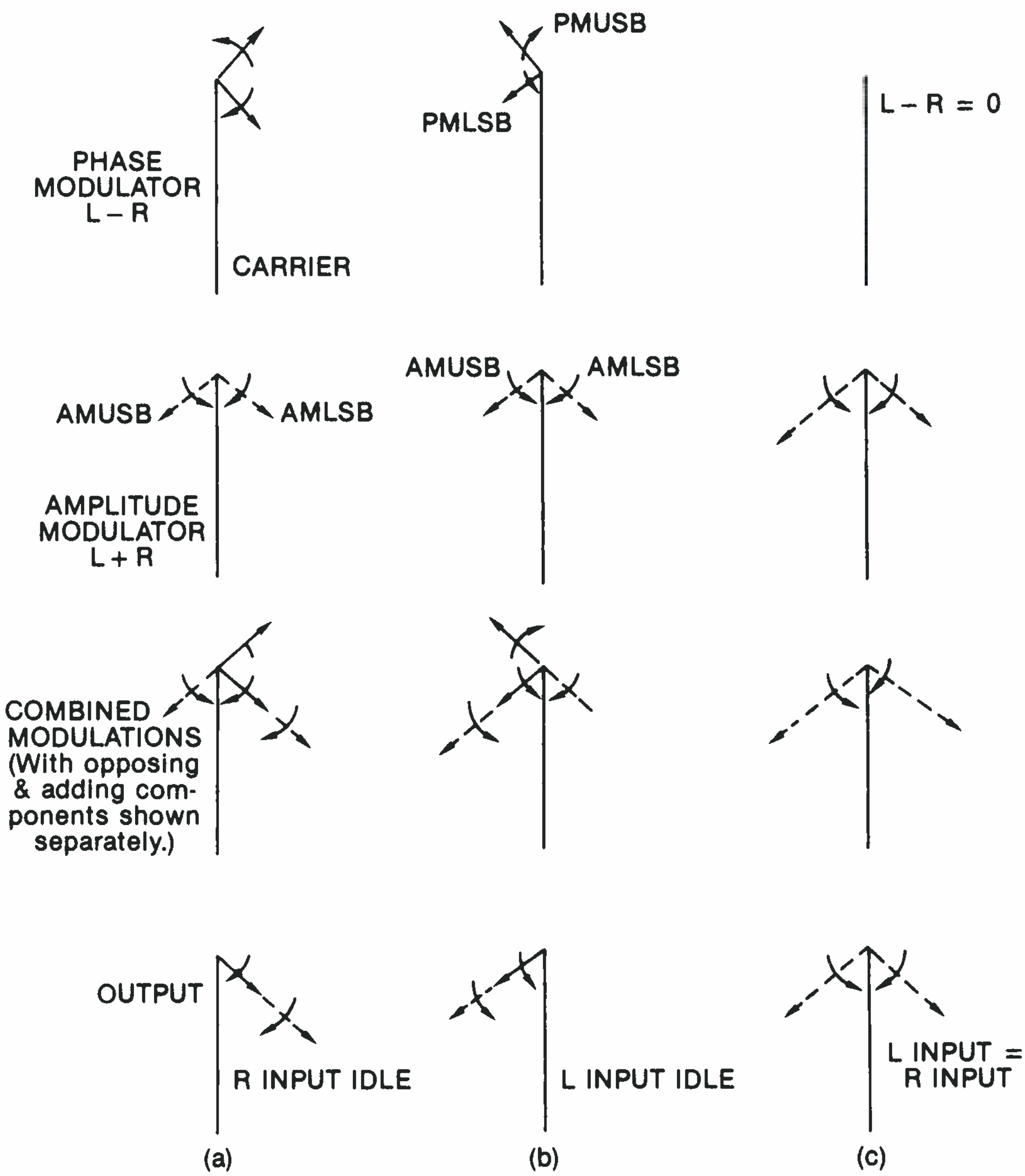


Fig. 2. Simplified phasor sketch.

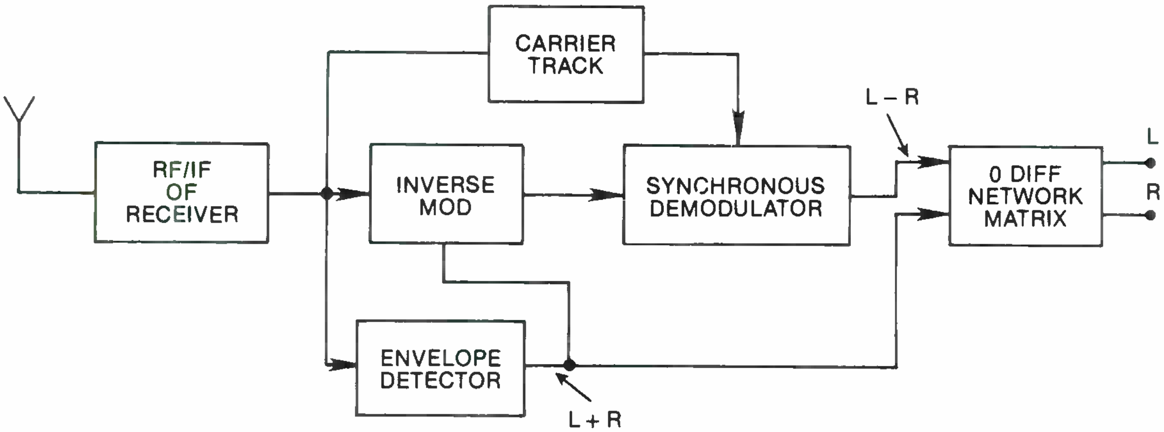


Fig. 3. Simplified block diagram of a receiver suitable for demodulating an independent sideband stereo wave.

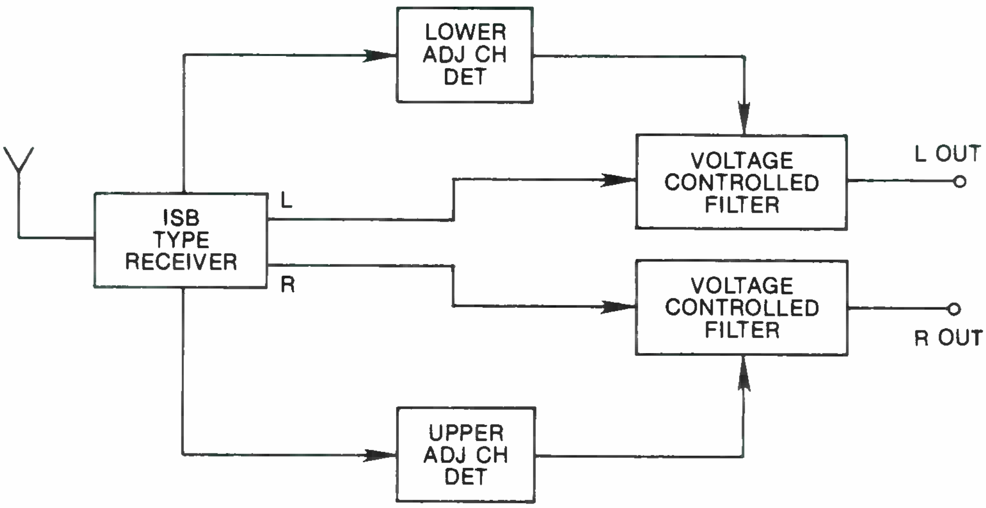


Fig. 4. Simplified block diagram of asymmetrical sideband selectivity.

(Please see Fig. 4 which is a simplified drawing of asymmetrical sideband selectivity.)†

It is noted that the interference above and below the desired channel is constantly monitored. Under low interference conditions, as would be expected when listening to the local stations or during daytime reception, the receiver's bandpass is wideband, providing high fidelity performance. When significant interference is detected, the channel with interference is automatically reduced in bandwidth, attenuating the offending adjacent channel interference.

Since under normal conditions the interference is not symmetrical the frequency response of the overall system is restricted by stereo channel (*L* or *R*) having the higher response because it has the least interference. However, there is a 6 dB step in the overall response because only one channel would have a 10 kHz signal level, for example. Thus, by use of the "mixed highs" concept the frequency response is degraded by much less drastic loss; i.e., 6 dB rather than say, 20 dB or more. This, of course, is substantially better response than would be the case if both channels were reduced in bandwidth symmetrically. It is important to note that this type of adaptive selectivity is also usable for monophonic reception of mono signals.

It is the author's opinion that this improvement in effective selectivity, since it can be implemented by use of newly developed** IC technology at very low cost, may well be the main advantage in implementing the ISB AM Stereo system and may have a major effect on the future growth of AM broadcasting.

---

*L.R. Kahn, U.S. Patent 4,018,994 dated April 19, 1977.
L.R. Kahn, U.S. Patent 3,973,203 dated April 3, 1976.
Hazeltine Research Inc., Note 1, April 5, 1982, "ISB AM Stereo Receiver Practices."

†L. R. Kahn, U.S. Patents 4,192,970 and 4,206,317.

**R.W. Broderson, P.R. Gray and D.A. Hodges, "MOS Switched-Capacitor Filters," PIEEE, vol. 67, pp. 61-75, January 1979.

3.2

AM Stereo Systems

BELAR AM Stereo System

Arno M. Meyer
BELAR Electronics Laboratory, Inc.
Devon, Pennsylvania

The Belar AM-FM System is a matrixed system which uses left and right channel audio material to form a sum signal $(L+R)$ and a difference signal $(L-R)$. The sum signal is applied to the amplitude modulation circuitry of a conventional AM transmitter, as currently done in monophonic broadcasting. Stereo $(L-R)$ information is used to angularly modulate the AM carrier. The angular modulation has FM characteristics for low audio frequencies, and PM (phase modulation) characteristics for mid audio frequencies. The $L-R$ signal from the audio matrix circuit is applied to a pre-emphasis circuit. This FM-to-PM changeover is accomplished with a controlled preemphasis network. Preemphasis, combined with deemphasis in the receiver, serves to reduce the detected noise in the stereo signal; primarily for the low audio frequency range. The Belar system has a significant advantage over competing systems in this regard. Peak FM deviation of the lowest audio frequencies transmitted is 312.5 Hz or 6.25 radians at 50 Hz. Similarly, the phase modulation index of the highest audio frequencies is 0.85 radians/sec. The preemphasis network has a controlled time constant to limit deviation at high audio frequencies, which will limit the occupied bandwidth of the stereo signal. A 10 Hz tone with a deviation of 2 radians is used as an indicator for receivers of the transmission of a stereo signal.

The receiver is essentially a complementary system to the transmitter. A conventional diode detector detects the L + R signal in the envelope. A limited IF signal (free of AM modulation) is applied to an FM discriminator. An audio high-pass filter rejects the pilot tone. The deemphasis network is the complement of the preemphasis network in the transmitter. The L + R and L − R signals are applied to an audio matrix which produces the original left and right audio signals. The output of the discriminator is applied to a band-pass filter centered at 10 Hz so that most program material and noise is removed from around the pilot tone. The presence of a signal at the output of this filter triggers a stereo indication mechanism.

The use of transmitted preemphasis and a deemphasis receiver (as is done in the Belar system) will reduce the low frequency microphonic and tuning noise which occurs naturally in non-synthesized (and some synthesized) receivers. This is one of the significant advantages of the Belar systems, and is one of the reasons the Belar system received a #2 rating in the FCC original analysis (FCC "Further Notice").

3.2

AM Stereo Systems

AM Stereo and the PMX AM Stereo System

Robert D. Streeter, P.E.
Consultant for
N.A.P. Consumer Electronics Corp.
Knoxville, Tennessee

INTRODUCTION

The PMX AM STEREO broadcast system is based on simple fundamental technical concepts, yet it offers a high level of technical and operating performance. The broadcast signal uses frequency modulation for the stereo identification tone, linear phase modulation for the stereophonic content (the vector difference of the left and right audio channels), and conventional envelope modulation for the monaural content (the vector sum of the left and right audio channels). The system was initially selected as the USA standard, and achieved top scores in both technical analysis conducted by the Federal Communications Commission.

MATHEMATICS

The broadcast signal equation for the PMX AM Stereo system is:

It should be noted that the pilot tone is actually a frequency modulated signal, although the single tone phase equivalent equation is given for simplicity. The generalized pilot tone phase equivalent equation would contain the integral of the frequency modulation input signal.

SYSTEM IMPLEMENTATION

A block diagram of the encoder and transmitting system is shown in Fig. 1. The audio processor is shown to emphasize the importance of correct processing to obtain the full capabilities of am stereo. Standard FM Stereo audio processing may allow as much as a 6 dB modulation and loudness loss in conventional monaural AM receivers. The processed audio is applied to the AM stereo exciter unit.

The processed left and right channels are matrixed in the stereo exciter unit to form the

$$[1 + (L(t)+R(t))] \sin [w_c t + (L(t)-R(t)) + B + P\sin(10\pi t + A)]$$

Where: $L(t)$ is the left channel audio signal
$R(t)$ is the right channel audio signal
$-1 < (L(t)+R(t)) < +1.25$ [am limits: -100% to $+125\%$]
$-1 < (L(t)-R(t)) < +1$ [pm limits: ± 1 radian]
w_c is the angular carrier frequency
B is an arbitrary phase constant
P is the pilot tone phase deviation [$3.0 < P > 3.6$]
A is an arbitrary phase constant
t is time

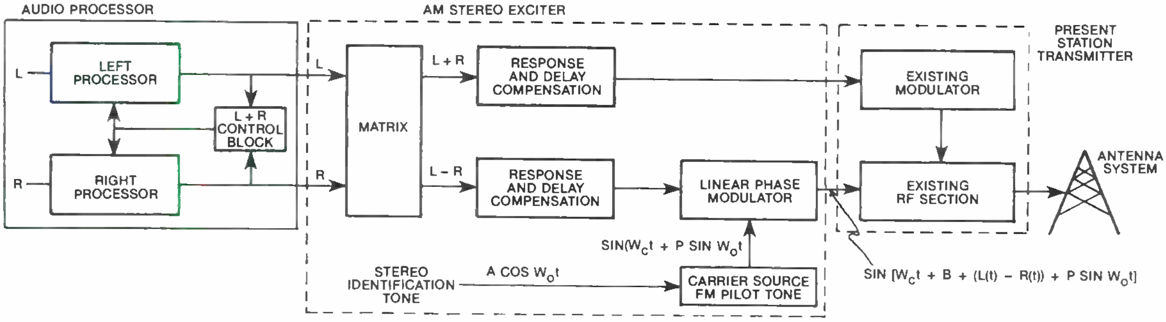


Fig. 1. AM/FM system block diagram.

L + R and L − R audio signals. The L + R and L − R signals are compensated for amplitude and time delay effects in the transmitter. The purpose of the compensation is to match the response of the amplitude and angular modulation channels, and to insure that the amplitude and angular modulation audio components will be broadcast at exactly the same instant from the antenna. This optimizes the receiver performance. The L − R audio signal is applied to a linear phase modulator, which varies the instantaneous phase of the transmitted RF signal at an audio rate. A peak phase deviation of 1 radian is used. The carrier signal is also frequency modulated by a sub-audio signal to generate a stereo identification tone. The stereo identification tone (pilot tone) deviates the carrier by about 17 Hz peak, at a 5 Hz rate. The pilot tone may also be a digital data signal (SCA), proving that the receiver will recognize it as a pilot signal. This AM SCA is provided now.

The stereo exciter supplies the RF signal to be transmitted. The master oscillator in the transmitter is disabled, and the exciter RF signal is substituted in its place. This RF signal contains the angular modulation of the pilot tone and the L − R audio signals. The RF signal is amplified in the transmitter to full power, and is amplitude modulated with the L + R audio signal from the exciter. For best stereo performance, the transmitter should have optimum monaural performance and have any extraneous angular modulation signals reduced or eliminated. No other changes to the transmitter are necessary.

Fig. 2 is a block diagram of the stereo receiver. It is the inverse process of the transmitter function. The received signal is envelope detected in the normal manner to recover the L + R audio signal. The stereo decoder detects the L − R audio signal by first reducing (or removing) the envelope modulation from the IF signal. The resultant signal is then linearly phase detected to obtain the audio L − R component. The pilot tone is also detected to provide an indication of a stereophonic transmission. The receiver operation can be made automatic if the pilot tone is used to control the operation of the receiver matrix circuits. Reception of the pilot tone would add the L − R audio to the matrix and provide stereo, while the absence of the pilot tone would provide just the L + R audio signal to the left and right audio channels.

The transmission and reception shown here is mathematically correct. That is, under ideal conditions there is no limit to the excellent fidelity, separation, and distortion performance of the system. Any performance less than perfect is due to the imperfections of the transmitting or receiving hardware.

It is impossible to detect stereo when the envelope is completely cut off (−100% modulation) for ANY stereo system. While such a signal can be transmitted easily, its detection would require the instantaneous recovery of the angular modulation signal when the signal has completely disappeared, an impossibility. All stereo systems either provide some form of receiver control to mask this impossible condition or restrict the transmitted signal so the condition will not occur (or both). The PMX system imposes a minimum of restrictions, and is capable of employing several types of receiver control.

ACTUAL SYSTEM PERFORMANCE

AM stereo hardware should be expected to operate with near perfection under laboratory conditions. The inability of the stereo exciter to produce high performance data with an ideal envelope modulator operating into a correct AM stereo detector is proof of an unacceptable stereo system. The performance data should always be obtained from a single stereo detecting device, since multiple devices (spectrum analyzer, separate single channel detectors, etc.) could result in adjustments that are not representative of the actual end performance obtainable. Multiple devices may be useful during adjustments, but are not a correct source of performance data.

Even more important than laboratory performance data is the performance that can be obtained with actual operating hardware. The data shown in Figs. 3, 4, 5, 6, 7, and 8 are graphs of the performance of the PMX system in use with a popular current transmitter. The stereo exciter used is a Continental Electronics 302A and the stereo monitor used is a Continental Electronics PMX-SM1. Fig. 9 is a block diagram of the equipment interconnection used for the data.

The 315R-1 used here was a stock transmitter with two new convenience features: the third harmonic resonator and neutralizing controls are made adjustable through the rear of the transmitter. This permits adjustment while the transmitter is in operation. It is necessary to obtain adequate RF drive power to the final amplifier, and this was done by maintaining the RF driver supply voltage above 190Vdc. The transmitter was operated into a dummy load, and was adjusted for maximum efficiency and minimum incidental phase modulation at the plate current minimum (plate dip). There was no special compensation, correction, or filtering used in the transmitter for stereo operation.

The AM stereo equipment was initially adjusted to provide optimum performance under ideal laboratory conditions using a high linearity envelope modulator in place of the AM transmitter. The ideal modulator was removed, and the 315R-1 connected in its place. The only adjustments made to the stereo system were the use of the correct time delay compensation value for the 315R-1 and the correct L + R signal level to the audio input on the 315R-1.

The data shown in these graphs reflects the performance of a current transmitter operating correctly without any additional corrections within the AM stereo system for the specific properties of the transmitter. Better performance can usually be obtained by compensating the stereo system for the envelope modulation properties of the transmitter.

All the data was taken at 85% main channel modulation, with the transmitter operating at 5 kW into a dummy load. Fig. 3 shows the envelope modulation performance (response and total harmonic distortion) as measured with an envelope detector. The noise floor was measured at −63.4 dB relative to 100%, or −62 dB relative to 85% modulation. Fig. 4 presents the same information for the L − R transmission channel, measured with a linear phase modulation detector. The low frequency rolloff allows the measurements to be made with the pilot tone present, as would be the case for an actual broadcast transmission.

Figs. 5 and 6 present the same information for the left and right channels respectively. It is significant to note that these measurements were made at 85% envelope modulation. This corresponds to 170% left (or right) channel modulation by FM stereo standards, and represents an operating stress level that some am stereo systems cannot withstand. Figs. 7 and 8 show the separation as measured with the 85% L + R (envelope)

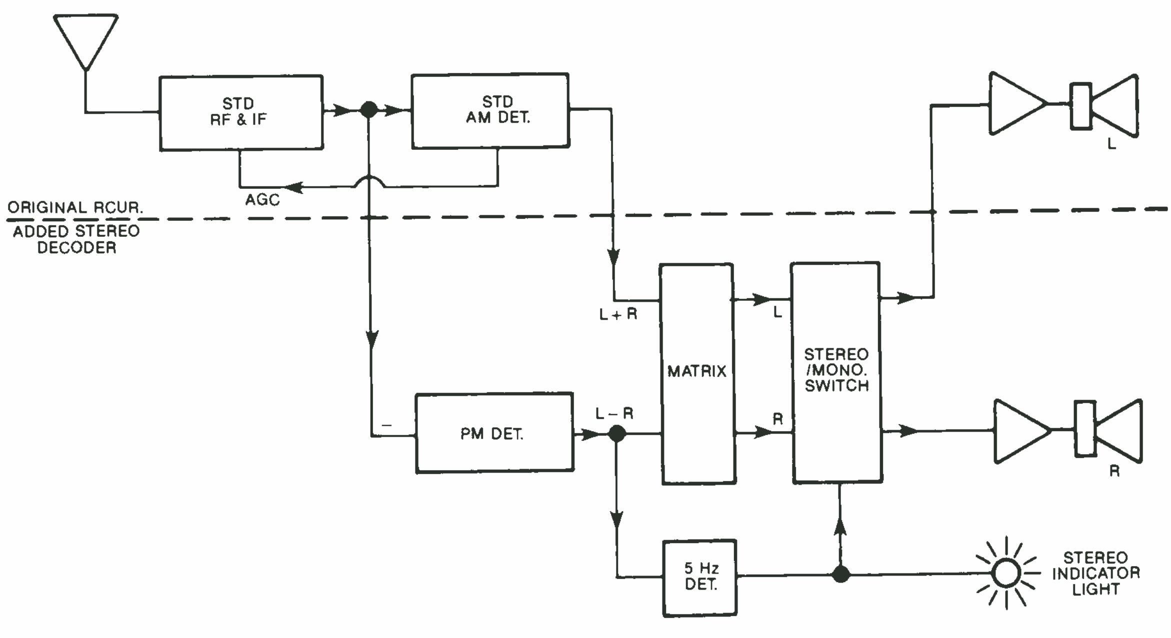


Fig. 2. Block diagram of the stereo receiver.

and L − R (phase) modulation levels used for the other stereo data.

MEASUREMENT METHODS AND TECHNIQUES

The above data presents an excellent indication of the actual performance that can be obtained with this system. The measurements are quite significant, but are not the only measurements significant to AM stereo transmissions.

The response and *THD* of an envelope detector for a left (and right) transmission will provide compatibility information, while an L − R transmission will provide cross-mode conversion (crosstalk) data. A L + R transmission with a L − R detection will measure the opposite cross-mode conversion, and is commonly used to evaluate the operating condition of the broadcast transmitter for AM stereo use. One of the most realistic tests of an AM stereo system is the simultaneous separate modulation of each channel with different test signals, as it simulates actual stereo programming conditions. It is important to evaluate both the left and right channels under the same conditions, as one channel can usually be "adjusted" for good performance at the expense of the performance in the opposite channel.

All AM stereo systems must have a common operating condition for the case where L = R, which is "center channel" stereo (which is identical to conventional monaural broadcasting). The differences between the AM stereo systems occur ONLY with stereo content in the signal, or L different than R.

OPERATING CHARACTERISTICS

Compatibility

The PMX system is perfectly compatible with conventional wideband envelope detectors. All am stereo systems have various degrees of imperfections with stereo using narrowband envelope detectors, as a result of crossmode conversions and RF filtering effects.

Coverage and Loudness

All stereo systems are equal with L = R modulation, and any differences are due to audio processing effects. With stereo program content, the ability to handle single channel modulation becomes important. The PMX system has no restrictions on single channel modulation, while all other active systems do. The PMX system offers the maximum available coverage and loudness because of this.

Occupied Bandwidth, Fidelity, Preemphasis, and Audio Processing

All of these items are interrelated, and must be considered together. Increased transmitted fidelity must result in increased occupied bandwidth for any system. The use of preemphasis must increase the high frequency sideband levels for any system. Audio processing effects will control the sideband density for any system. All systems are identical for the L = R condition, and any differences will be the result of the stereo content of the transmitted signal. The PMX system remains within the FCC occupied bandwidth restrictions without filtering for any reasonable operating condition, while providing full legal fidelity performance.

Synchronous Detection

All stereo systems operate correctly with synchronous detection for the L = R case. As the amount of stereo information increases, only a pure quadrature system retains maximum performance. Experience has shown that the PMX system delivers acceptable stereo performance with a synchronous detector, even though the system is not intended for such operation.

Soundfield Disturbances

Interference from co-channel and adjacent channel signals will influence the received signal characteristics, as will self-interference from reradiation. All stereo systems are affected by reradiation and the received audio signal will "move about" in a fashion that depends on each system. Adjacent channel interference will increase the effective modulation seen by the stereo detector, but the result will depend on the details of the detector operation as much as on the stereo system. There are differences in the co-channel interference performance, and the PMX system pilot tone tends to scramble the interfering phase structure and eliminate the undesired effects on the desired received stereo signal. The PMX system is thus relatively immune to co-channel interference.

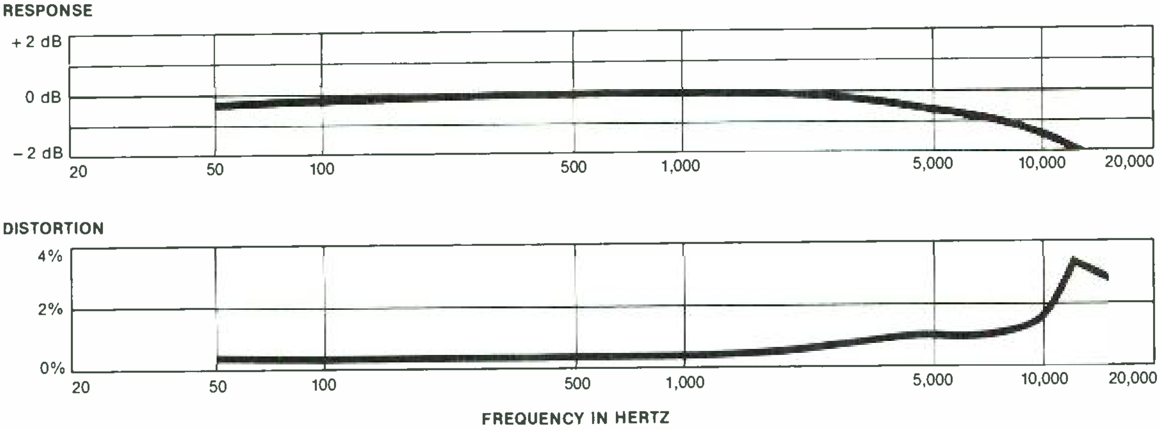


Fig. 3. L + R 85% envelope modulation. −62 dB noise floor/ref 85%. THD measured at constant envelope modulation 5 KW 315R-1 (828E-1) 1440 KHz.

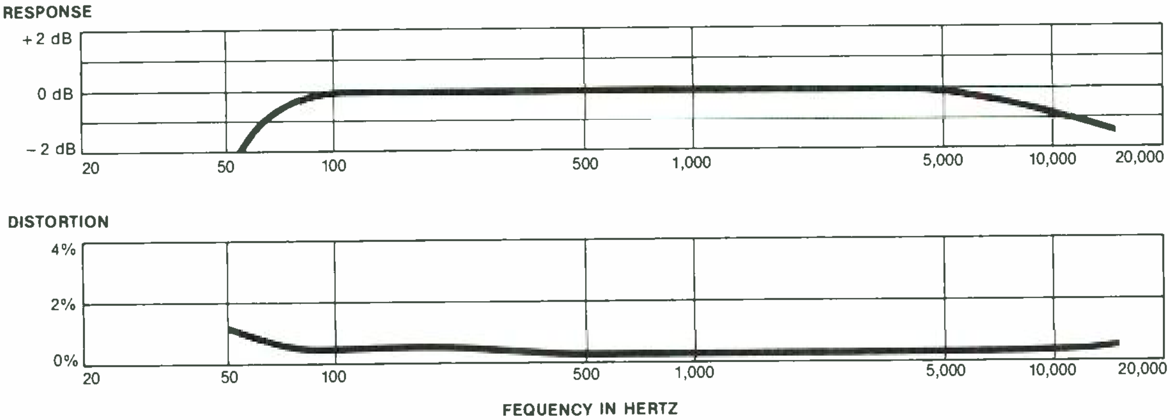


Fig. 4. L − R 85% modulation. −57 dB noise floor/85% modulation. Distortion measured at constant input due to L − R filtering in monitor 5 KS 315R − 1.

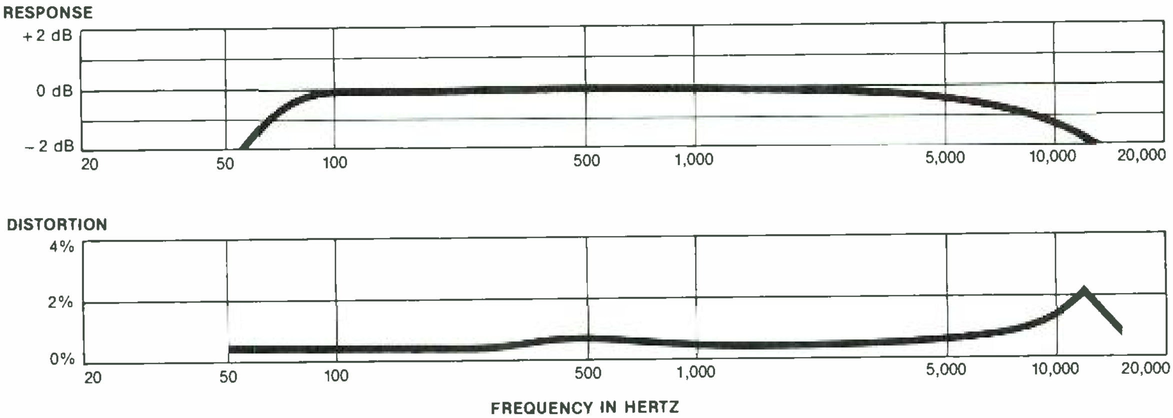


Fig. 5. Left channel 85% modulation (170% equiv. left mod). −60.5 dB noise floor/rel 85% modulation.

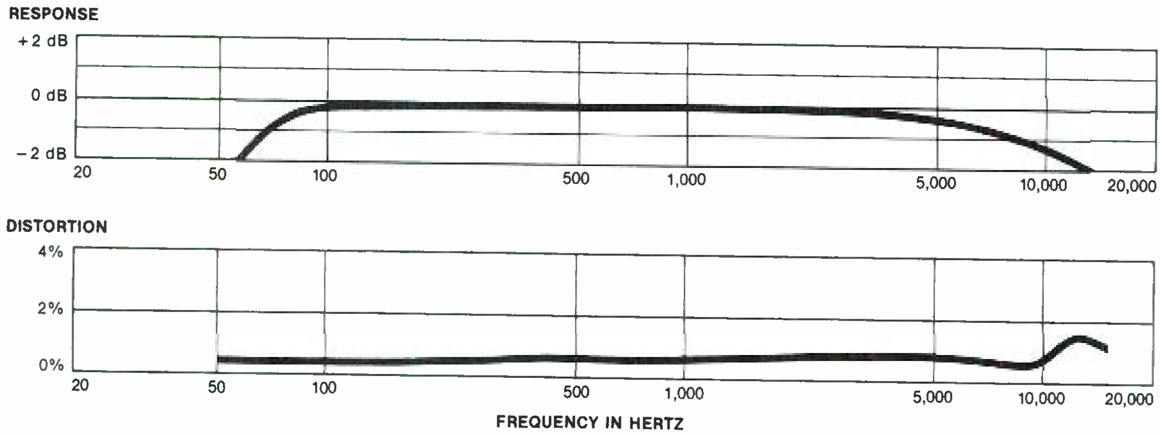


Fig. 6. Right channel 85% envelope modulation (170% equiv. right mod).

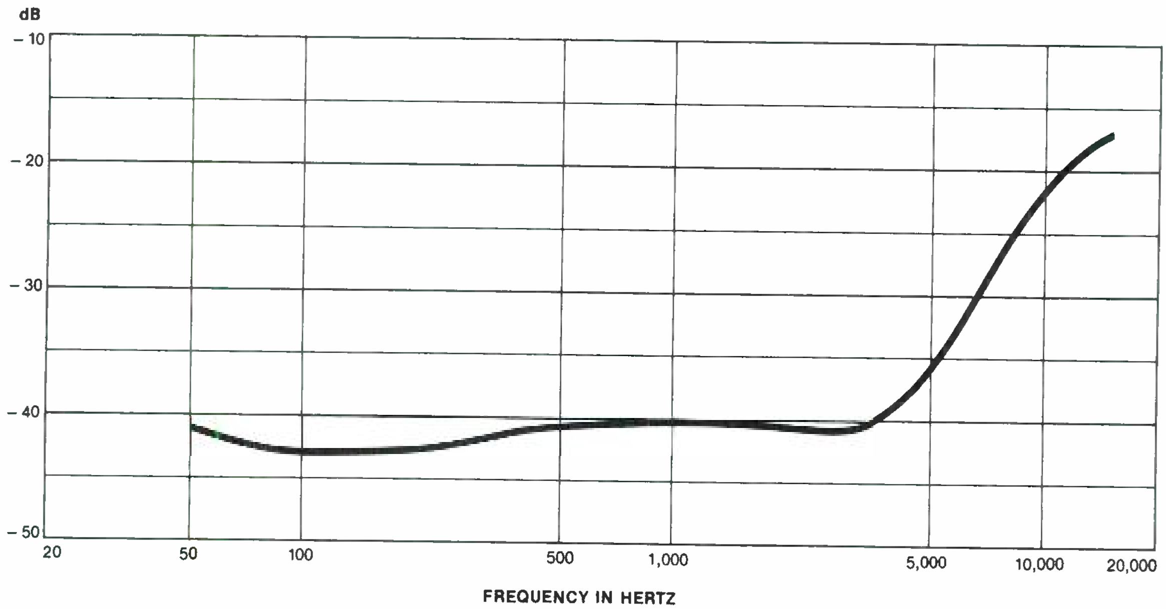


Fig. 7. Left channel detected, right channel transmitted at 85% envelope modulation (170% single channel).

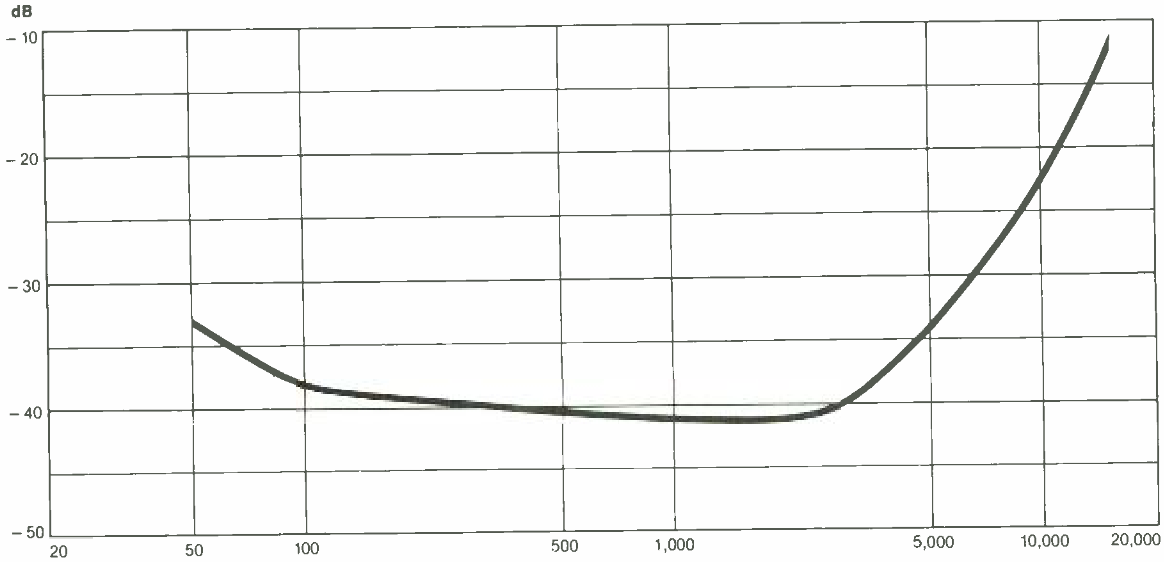


Fig. 8. Right channel detected, left channel transmitted at 85% envelope modulation (170% single channel).

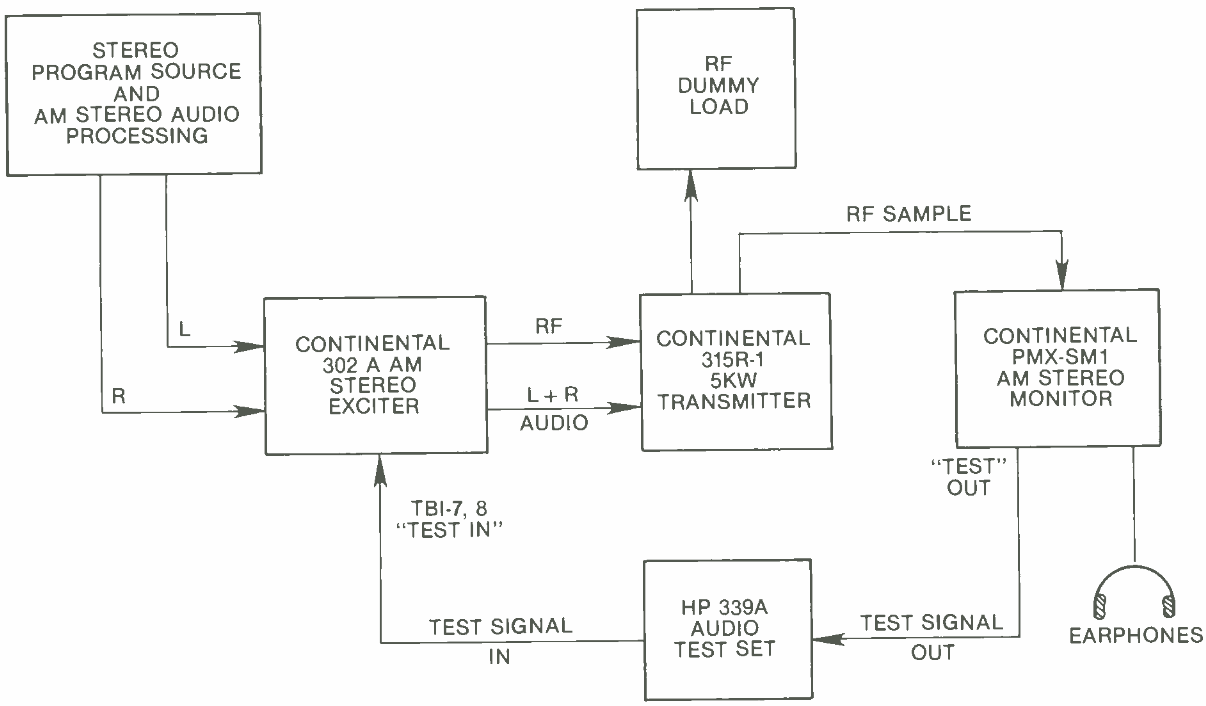


Fig. 9. PMX AM Stereo/315R-1 equipment interconnection.

3.3

FM Broadcast Transmitters

Geoffrey N. Mendenhall, P.E.
Vice President of Engineering
and
The Engineering Staff of Broadcast Electronics Inc.
Quincy, Illinois

With contributions by:
Warren B. Bruene, P.E.
Electrospace Systems Inc.
Richardson, Texas

HISTORY OF FM BROADCASTING

Although the mathematical principles explaining frequency modulation (FM) had been known for many years, the advantages and practical application to radio broadcasting were not realized until the 1930's, when Major Edwin H. Armstrong did extensive developmental work proving that FM radio transmissions were possible. Many theoreticians claimed to have proof that Armstrong's experiments were impossible based on mathematical models claiming that an infinite transmission bandwidth would be required. He never received proper credit for his many contributions to the radio communications industry during his lifetime. See reference (1) at the end of this chapter for more information about the career of Major Armstrong.

Among the advantages of FM are freedom from static, wide audio bandwidth, and the ability of an FM receiver to capture the stronger of two signals transmitted on the same carrier frequency.

In 1940, following extensive public hearings, the Commission established the FM Broadcast Service and set aside 40 channels in the 42 to 50 MHz band with commercial operation scheduled to begin January 1, 1941. Although World War II stopped all non-military radio construction, more than 40 FM stations continued to serve over 400,000 receivers. To eliminate the interference problems resulting from skywave reflection in the prewar FM band, the Commission moved the FM Broadcast Service to the 88 to 108 MHz band in 1945, thereby increasing the number of available channels to 100. However, the expected growth of FM broadcasting did not materialize. Since conversion of prewar FM receivers to the new band was not practical, purchase of a new receiver was the only way to receive the new FM stations. Television appeared to offer much more to the consumer than FM radio, since most FM stations merely duplicated the programming of an affiliated AM station. Despite the potential for a higher quality broadcast service, there was little public demand for new FM receivers and virtually no public reaction when FM stations dropped popular programs from their schedule or were off the air due to equipment failures. It is not surprising, therefore, that in May 1950, there had been only 16 new FM license applications during the previous 15 months in which 259 FM stations ceased operations. It was not until after the introduction of stereo multiplex FM broadcasting that public awareness of the FM band increased and then skyrocketed in the 1970's to make FM the dominant medium for musical programming.

CHARACTERISTICS OF FM COMPARED TO AM

Reduced Noise

The 88 to 108 MHz FM broadcast band is relatively free of atmospheric and other noise in-

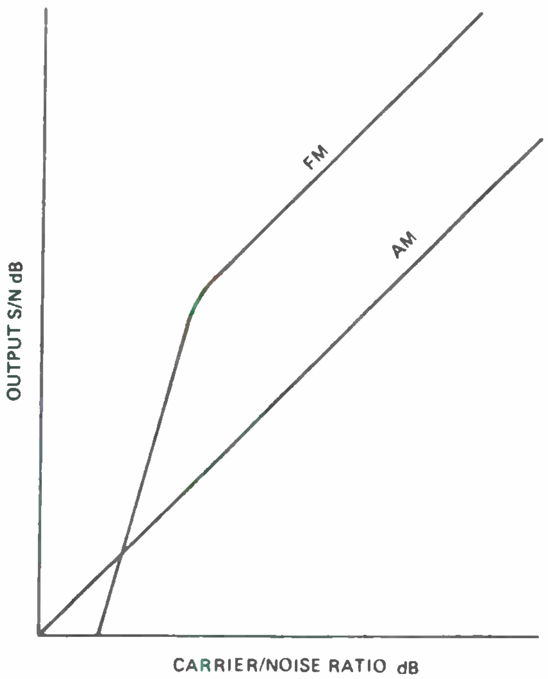


Fig. 1. FM improvement factor.

terference. Emission at these frequencies is not propagated great distances by the ionosphere as it is in the 535 to 1605 kHz AM broadcast band. Therefore, noise from lightning discharges is limited to line-of-sight distances and is almost negligible. Man-made noise is a far greater source of noise, particularly in urban areas. The level of man-made noise falls off at increasing higher frequencies so that the microvolts-per-meter noise level is about one-tenth as great in the FM band as it is in the AM band.

In addition, FM has an improved noise threshold characteristic when compared to AM. Limiting circuitry symmetrically clips the RF waveform in an FM receiver to remove any amplitude variations produced by static or impulse noise before they reach the demodulator which responds only to phase or frequency changes in the signal. The FM improvement factor is illustrated in Fig. 1. Note the sharp knee in the threshold curve above which noise and interference are suppressed, which results in an improved signal to noise (S/N) ratio. This same capture effect causes a weaker FM signal on the same channel to be suppressed, resulting in greatly reduced co-channel interference.

For even greater noise reduction, pre-emphasis (75 micro-second time constant) is employed in the transmitter whereby the audio frequency components above about 2.1 kHz are boosted in amplitude at the rate of 6 dB per octave before being applied to the modulator. Flat frequency response is restored in the receiver's de-emphasis network by attenuating the higher frequencies the same amount they were boosted in the transmitter. At the same time, the high frequency noise components which are characteristic to FM are also attenuated, resulting in greatly reduced background noise. Pre-emphasis is discussed in more detail in the FM theory and FM exciter sections of this chapter.

The net result of the above factors is a much better signal-to-noise ratio on FM than on AM and the ability to transmit much wider bandwidth information on FM. The lower background noise along with wider frequency response means that a signal of higher quality and wider dynamic range can be enjoyed.

Occupied Bandwidth

All the advantages of the FM broadcast medium over the AM broadcast medium do not come free of compromise. The primary compromise is that a standard FM broadcast channel occupies more than ten times the bandwidth of an AM broadcast channel. This is because the more complicated non-linear sideband structure of a frequency modulated carrier with wide deviation requires much more bandwidth than the simpler linear distribution of sidebands in an AM system. More details on the amplitude and spacing of FM sidebands are given in the section on FM modulation theory in this chapter. Fortunately, the FM broadcast band is located in the VHF portion of the frequency spectrum where more bandwidth is available than in the medium wave AM broadcast band.

High Fidelity

Fidelity is defined as the degree to which a system, or a portion of a system, accurately reproduces at its output the essential characteristics of the signal impressed upon its input. The large market for high-fidelity (HI-FI) equipment is evidence that the public recognizes and enjoys high quality reproduction. Many individuals spend a great deal of money to have high quality equipment in their homes.

Uniform frequency response over the audible range of at least 50 Hz to 15 kHz, very low amplitude distortion (harmonic and intermodulation), very low noise level, and good transient response (uniform time delay versus frequency) are necessary for HI-FI performance. The FM channel authorizations provide for adequate audio frequency response and a low-noise radio link to the listener. The rest of the performance is a matter of equipment design. The FCC has established minimum performance standards that assure HI-FI performance. Transmitter and receiver manufacturers strive to produce equipment that exceed these minimum standards and broadcast stations strive to maintain the maximum performance capability of their equipment to provide truly HI-FI programs for the listener.

Stereophonic Transmission

A higher level of realism and listener enjoyment is provided by stereophonic transmission. Again, the wide channel allocations and ability of FM to compatibly multiplex several audio channels on one carrier permitted development of a practical stereo broadcasting system. This provides a means for the broadcast industry to provide the public with the same or better quality reproduction than is available on stereo records or tapes. The advent of the "compact disc" and other digital audio source equipment will continue to challenge equipment manufacturers and station engineers to further improve the performance of the entire FM broadcast chain. Detailed technical information about stereophonic transmission theory and standards is provided in the chapter about FM stereo and subcarriers.

Subcarrier Transmission

The wide channel bandwidth authorized for FM broadcasting also makes it feasible to multiplex several audio or data channels as subcarriers together with the stereo transmission. These channels commonly called SCAs, provide an important source of revenue to many stations as well as provide many useful audio and digital services to the community. Detailed technical information about subcarrier transmission theory and standards is provided in the chapter about FM stereo and subcarriers.

FCC Transmission Standards

The Federal Communications Commission regulates and enforces the technical standards that apply to radio broadcasting in the United States. In theory, this will assure that the public is provided with a consistently high standard of transmission quality from station to station. The rules and regulations covering transmitter technical performance are set forth in Part 73 of the FCC Rules and Regulations available from the U.S. Government Printing Office in Washington, D.C. The rules and regulations are changed from time to time to keep pace with new technology and changes within the broadcast industry. Every broadcast engineer should have access to a current copy of these rules and regulations so that the station's technical performance is maintained within the prescribed limits.

FM MODULATION THEORY

Angular modulation

Frequency modulation (FM) and phase modulation (PM) are both special cases of angular modulation. In any angular modulation system both the frequency and phase of the carrier vary with time as a function of the modulating signal.

The relationship between the frequency deviation of the carrier, the phase deviation of the carrier, and the sinusoidal modulating frequency is defined as the modulation index (m).

$$\text{Where: } m = \frac{\text{frequency deviation } (\pm \text{ Hertz})}{\text{modulating frequency (Hertz)}}$$

Since frequency modulation and phase modulation are both subsets of angular modulation they are virtually indistinguishable from one another except in the modulator characteristics.

In a PM system, the modulating signal causes the phase of the carrier wave to vary according to the instantaneous amplitude of the modulating signal. A phase modulator generates a constant amount of phase deviation of the carrier with a constant amplitude modulating signal independent of the frequency of the modulating signal. The frequency deviation of the carrier produced by a phase modulator does increase as the modulating frequency is increased even though the level of the modulating voltage is held constant. The net effect is that the phase modulator behaves as if it were a frequency modulator with a 6 dB/octave rising slope on the modulating signal input.

A frequency modulator generates a constant frequency deviation of the carrier with a constant amplitude modulating signal independent of the frequency of the modulating signal. The phase deviation of the carrier produced by a frequency modulator decreases as the modulating frequency is increased even though the level of the modulating voltage is held constant. The net effect is that the frequency modulator behaves as if it were a phase modulator with a 6 dB/octave falling slope on the modulating signal input.

In FM broadcasting, we desire a signal that has frequency deviation that is proportional the amplitude of the modulating signal, but independent of the frequency of the modulating signal as is produced by a frequency modulator.

The instantaneous frequency (rate of change of phase) of the RF output wave differs from the carrier frequency by an amount proportional to the instantaneous value of the modulating waveform. For example, consider a 100 MHz carrier wave FM modulated by a 1000 Hz audio tone and assume that a 1 volt input to the modulator causes ±20 kHz of frequency deviation on the positive and negative peaks of this tone. If the audio input amplitude is increased to 2 volts, the peak deviation will become ±40 kHz varying in sine-wave fashion from one peak deviation to the other and back again at the 1000 Hz rate. In FM broadcasting, the FCC has established that a peak frequency deviation of ±75 kHz shall constitute 100 percent modulation.

It is interesting to note that when pre-emphasis is used ahead of the FM modulator, the system becomes a phase modulator at audio frequencies above the turnover point of the pre-emphasis network. This is because the frequency response of the pre-emphasis network rises at the rate of 6 dB/octave above this point. FM broadcasting with pre-emphasis really becomes a mixture of FM at low modulating frequencies and PM at high modulating frequencies.

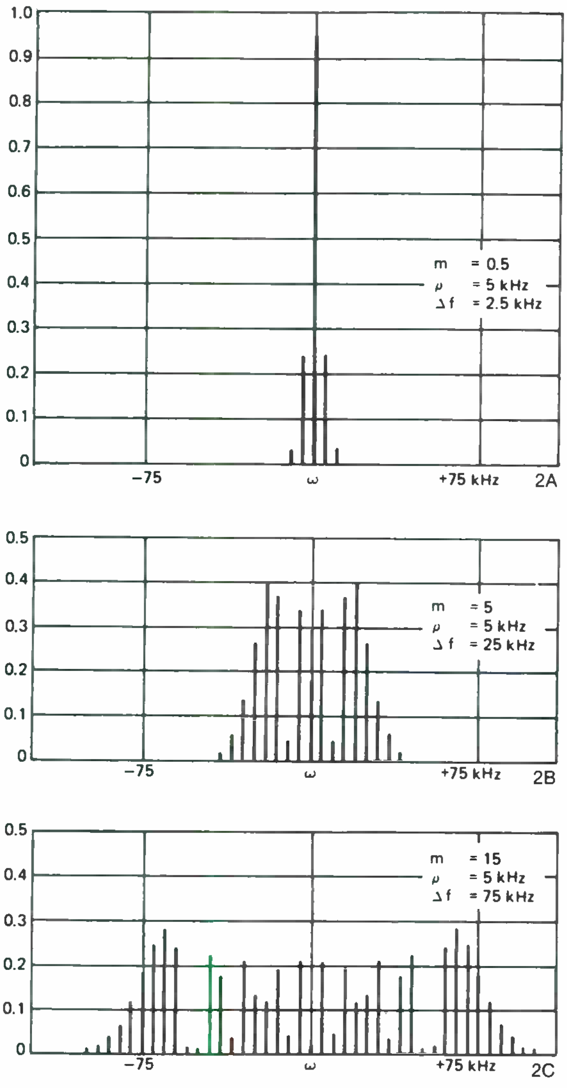


Fig. 2. RF spectrum with modulation indexes of 0.5, 5.0, and 15.

FM Sideband Structure

The frequency modulated RF output spectrum contains many sideband frequency components, theoretically an infinite number. They consist of pairs of sideband components spaced from the carrier frequency by multiples of the modulating frequency. When the modulation index is small ($m = 0.5$) the amplitude of the second and higher order sidebands is small so that the output consists mainly of the carrier and the pair of first-order sidebands, as illustrated in Fig. 2A. The total transmitter RF output power remains constant with modulation, but the distribution of that power into the sidebands varies with the modulation index so that power at the carrier frequency is reduced by the amount of power added to the sidebands.

As the modulation index is increased as in wide deviation FM broadcasting, the higher order sidebands become more prominent. The amplitude and phase of the carrier (J_0) as well as the sidebands (J_1 thru J_n) can be expressed mathematically by making the modulation index (m) the argument of a simplified Bessel function.

$E(t) =$	total RF output voltage
$A[J_0\ (m)\sin \omega c(t)]$	carrier amplitude
$+ [J_1\ (m)\sin(\omega c + \omega m)t]$	first order upper sideband
$- [J_1\ (m)\sin(\omega c - \omega m)t]$	first order lower sideband
$+ [J_2\ (m)\sin(\omega c + 2\omega m)t]$	second order upper sideband
$+ [J_2\ (m)\sin(\omega c - 2\omega m)t]$	second order lower sideband
$+ [J_3\ (m)\sin(\omega c + 3\omega m)t]$	third order upper sideband
$- [J_3\ (m)\sin(\omega c - 3\omega m)t]$	third order lower sideband
$\pm [J_n\ (m)\sin(\omega c \pm n\omega m)t]$	... higher order sidebands

Where:

A = The unmodulated carrier amplitude constant

J_0 = The modulated carrier amplitude

$J_1, J_2, J_3 \ldots J_n$ are the amplitudes of the nth order sidebands

m = The modulation index

$\omega c = 2\pi F_c$ (The carrier frequency)

$\omega m = 2\pi F_m$ (The modulating frequency)

The numeric values of the Bessel functions (J_0 thru J_n) which express the amplitudes of the various frequency components can be found in mathematical tables. Fig. 3 shows a graphical representation of how the Bessel function values for the carrier and the first eight pairs of sidebands vary with the modulation index.

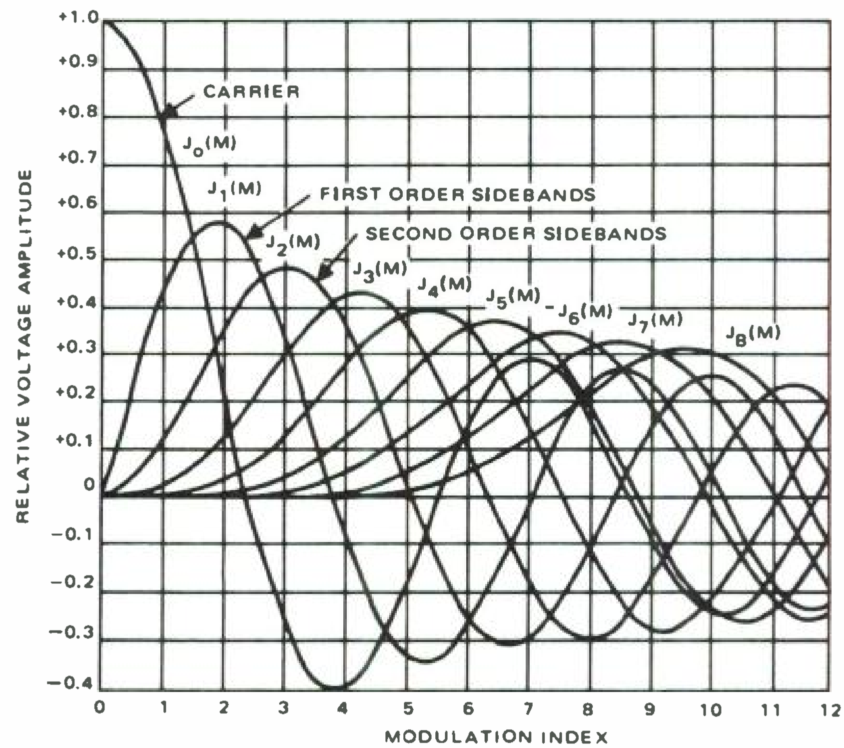


Fig. 3. Relationship of carrier and sideband amplitudes to modulation index.

In a monophonic FM broadcast transmitter, the modulation index can become very high at low modulating frequencies. With a 50 Hz audio input signal of sufficient amplitude to produce 75 kHz deviation (100 percent modulation), the modulation index is:

$$m = \frac{75000}{50} = 1500$$

With a 15,000 Hz input at the same deviation (also 100 percent modulation), the modulation index is only:

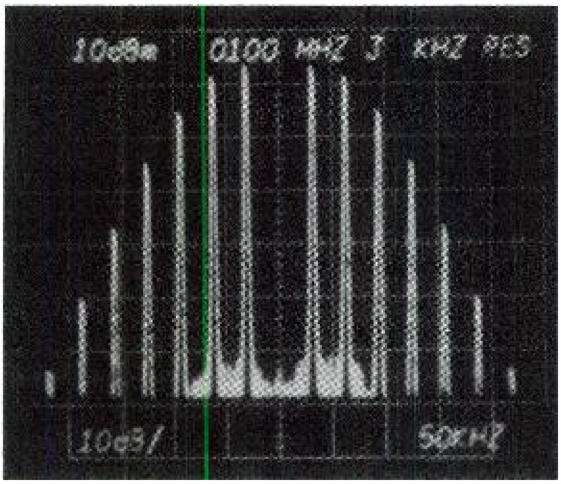
$$m = \frac{75000}{15000} = 5$$

Fig. 2B and 2C illustrate the frequency components present for modulation indices of 5 and of 15. Note that the number of significant sideband components becomes very large with a high modulation index. The total bandwidth occupied extends beyond ±75 kHz from the carrier depending upon the modulating frequency. This single tone modulating frequency analysis is useful in understanding the general nature of FM and for making tests and measurements. When program modulation is applied, there are many more sideband components present and they are varying so much that sideband energy becomes distributed over the entire occupied bandwidth rather than appearing at discrete frequencies.

Bessel Nulls

It is interesting to note that at certain modulation indices, the carrier amplitude goes to zero with all the transmitted power distributed at frequencies other than the carrier frequency. This carrier null phenomenon is useful as an extremely accurate method for measuring the frequency deviation and to check the calibration of modulation monitors. Referring again to Fig. 3, note that the carrier amplitude goes to zero and reverses sign at several values of modulation index including; 2.405, 5.520, and 8.654. Fig. 4 is a photograph taken from an RF spectrum analyzer showing the first Bessel null ($m = 2.405$) of a carrier at a frequency of 100 MHz.

If we want to determine the audio input level required to achieve 75 kHz deviation, we can apply an audio tone of exactly 8,667 Hz (75,000 divided by 8.654) and increase the audio level until the carrier disappears for the third time. At this audio level the deviation is exactly 75 kHz. The carrier amplitude (null) detector must have sufficient selectivity to separate the carrier from the sidebands and could be a spectrum analyzer or a receiver with a narrow IF bandwidth. The FM signal can even be heterodyned down to a convenient frequency for measurement. Heterodyning does not alter the modulation index while frequency division or multiplication does change the modulation index. When a frequency (or phase) modulated wave is multiplied or divided,



FOR M = 2.405, F_M = 31,185Hz, F_C = 100.00MHz

Fig. 4. FM RF spectrum showing occupied bandwidth and Bessel Null single tone @ ±75 KHz deviation.

this also multiplies or divides the frequency deviation and the modulation index by the same amount.

A listing of useful carrier and first order sideband nulls as function of the modulation index (m) and the modulating frequency (F_m) is given below.

Occupied Bandwidth

After examining the Bessel function and the resulting spectra, it becomes clear that the occupied bandwidth of an FM signal is far greater than the amount of deviation from the carrier that one might incorrectly assume as the bandwidth. In fact, the occupied bandwidth is infinite if all the sidebands are taken into account, so it is now clear that a frequency modulation system would require the transmission of an infinite number of sidebands for perfect demodulation of information. In practice, a signal of acceptable quality can be transmitted in the limited bandwidth assigned to an FM channel.

Effects of Bandwidth Limitation

Practical considerations in the transmitter RF circuitry make it necessary to restrict the RF bandwidth to less than infinity. As a result, the higher order sidebands will be altered in amplitude and phase. Bandwidth limitation will cause distortion in any FM system.

Consider the model shown in Fig. 5A, where a perfect FM modulator is connected to a perfect demodulator via an RF path of infinite bandwidth. The demodulated audio shown in Fig. 5B contains no distortion components.

In Fig. 6A, a passive LC bandpass filter is inserted between the modulator and demodulator in order to restrict the bandwidth. Audio distortion products now appear at the output of our perfect demodulator as shown in Fig. 6B. These distortion products are due solely to the bandwidth restriction (300 kHz bw_3) imposed by the passive bandpass filter.

Fig. 7A and 7B show the effects of a narrowband RF bandpass filter on the RF spectrum of a composite signal consisting of a stereophonic subcarrier modulated only on the left channel with 4.5 kHz plus a 67 kHz unmodulated SCA subcarrier. The only distortion evident on the RF spectrogram is the loss of some sidebands greater than 150 kHz from the center frequency and some amplitude differences between the upper and lower sideband pairs. Fig. 7B shows the corresponding effects observed on the demodulated baseband spectrum for the same signal. Note the creation of many undesired intermodulation terms which cause crosstalk into both the stereophonic and SCA subcarrier bands. The change in the RF spectrum is subtle, but the resulting spectrum after demodulation is clearly modified.

As one can see, the distortion in any practical FM system will depend on the amount of band-

NULL	MODULATION INDEX (m)		(F_m) IN Hz (FOR 75 kHz DEVIATION)	
	CARRIER	1st SIDEBANDS	CARRIER	1st SIDEBANDS
1st	2.405	3.852	31,185	19,470
2nd	5.520	7.016	13,587	10,690
3rd	8.654	10.173	8,667	7,372
4th	11.792	13.323	6,360	5,629
5th	14.931	16.470	5,023	4,554
6th	18.071	19.616	4,150	3,823
7th	21.212	22.760	3,536	3,295

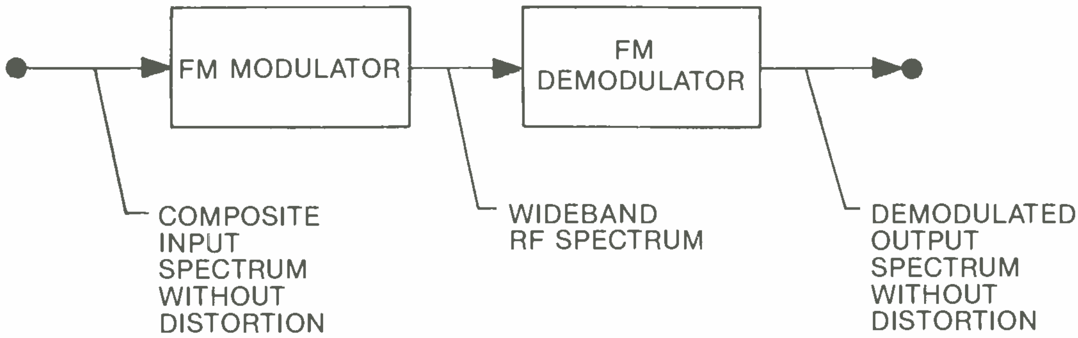
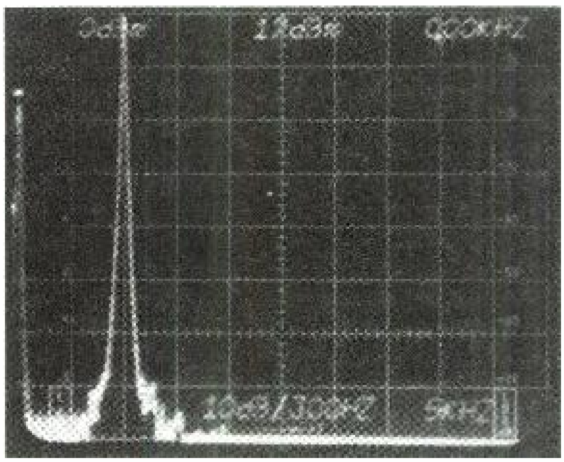
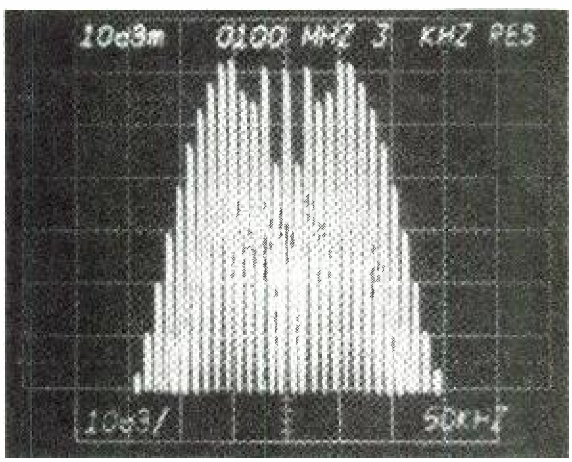


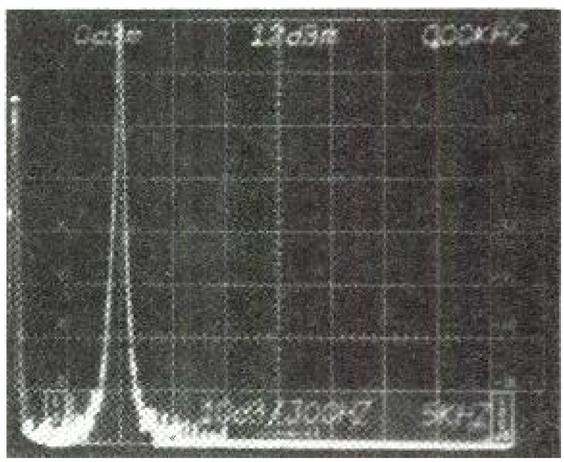
Fig. 5A. Wideband RF path.



BASEBAND SPECTRUM TO FM MODULATOR



RF SPECTRUM TO DEMODULATOR



DEMODULATED BASEBAND SPECTRUM

Fig. 5B. Single tone (10 KHz) modulation thru wideband RF path.

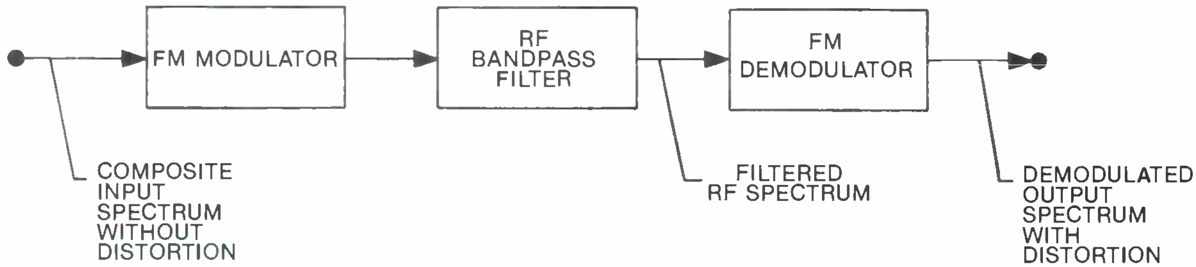
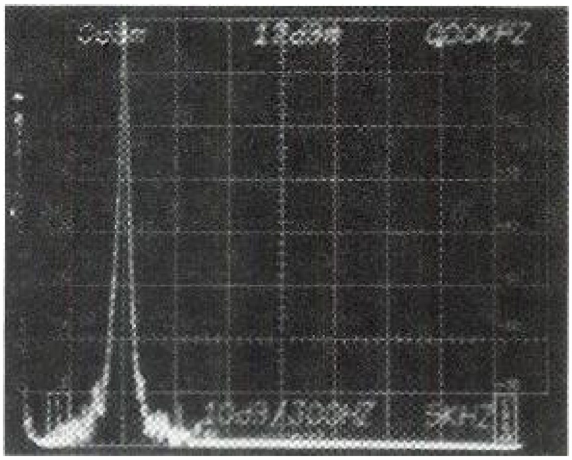
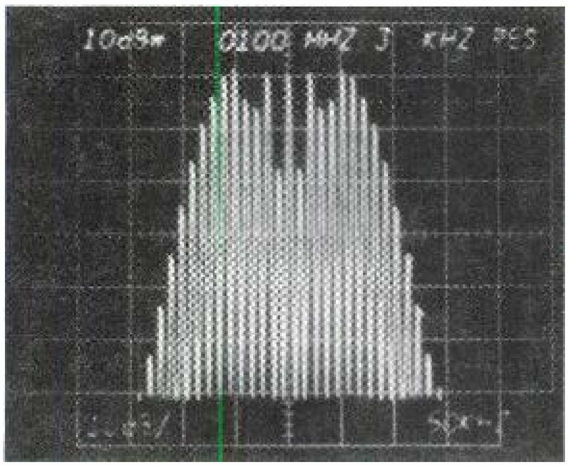


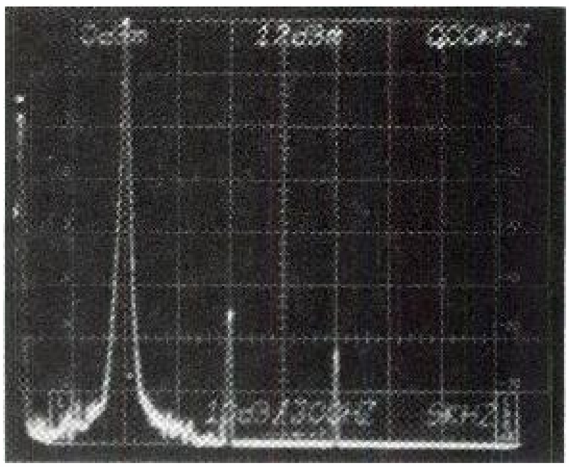
Fig. 6A. Bandwidth limited RF path.



BASEBAND SPECTRUM TO FM MODULATOR

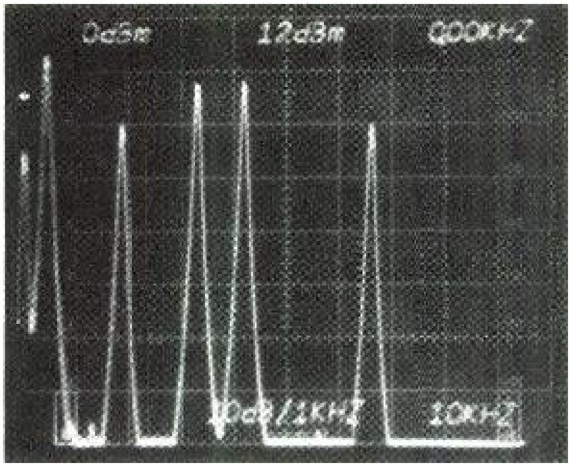


BANDWIDTH LIMITED RF SPECTRUM TO DEMODULATOR

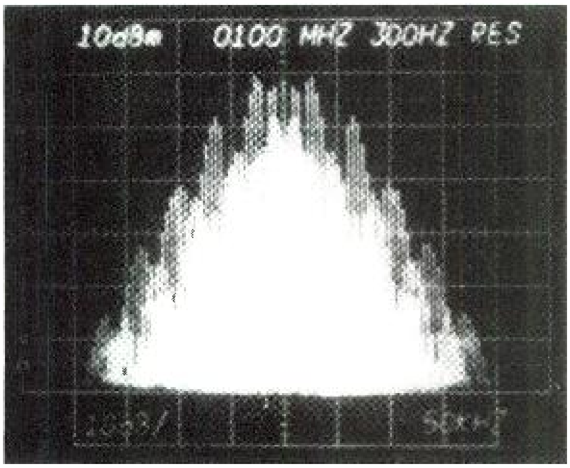


DEMODULATED BASEBAND SPECTRUM

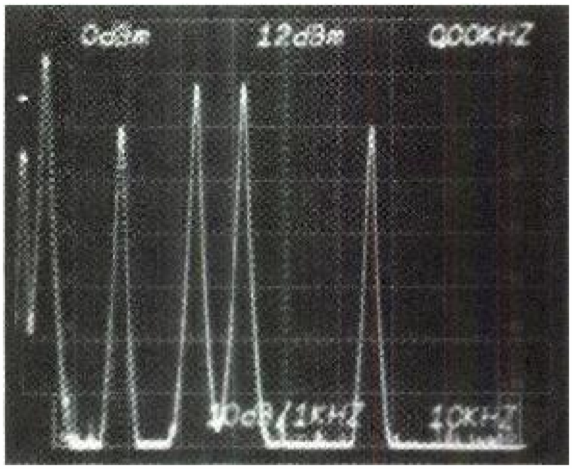
Fig. 6B. Singletone (10 KHz) modulation thru narrowband RF path.



BASEBAND SPECTRUM TO FM MODULATOR



RF SPECTRUM TO DEMODULATOR



DEMODULATED BASEBAND SPECTRUM

Fig. 7A. Stereo (L or R = 4.5 KHz) plus SCA (unmod.) modulation thru wideband RF path.

width available versus the modulation index being transmitted.

Limiting Factors Within An FM Transmitter

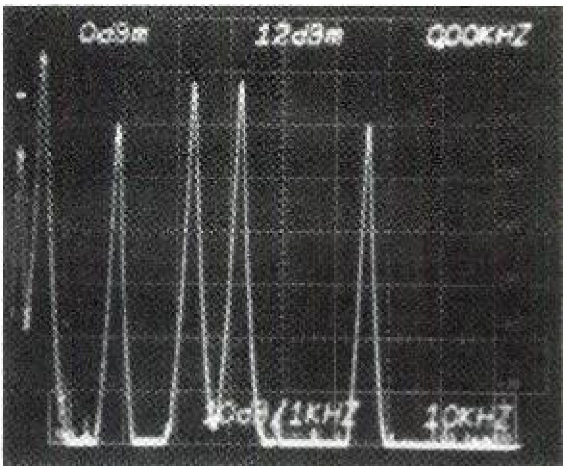
Relating the specific quantitative effect of the bandwidth limitations imposed by a particular transmitter to the actual distortion of the demodulated composite baseband is a complicated problem indeed. Some of the factors involved are:

1. Total number of tuned circuits involved.
2. Amplitude and phase response of the total combination of tuned circuits in the RF path.
3. Amount of drive (saturation effects) to each class "C" stage.
4. Non-linear transfer function within each amplifier stage.

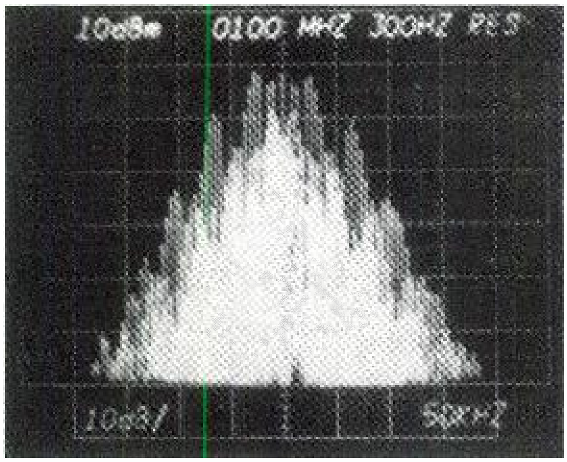
Improvement of The RF Path

The following design techniques can help improve the transmitter's bandwidth:

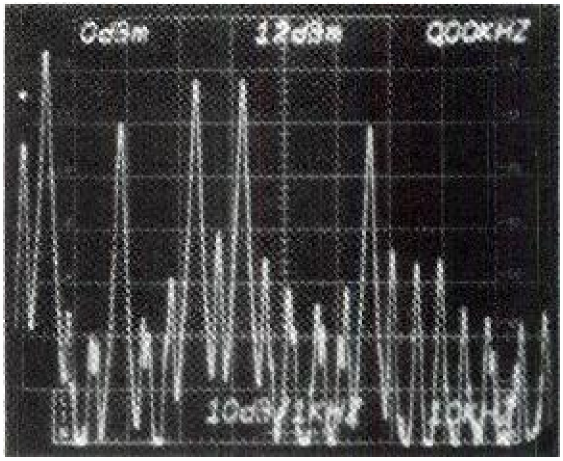
1. Maximize bandwidth by using a broadband exciter and a broadband IPA stage.
2. Use a single-tube design or a broadband, completely solid-state design where feasible.
3. Optimize both grid circuit and plate circuit of the tuned stage for the best possible bandwidth.
4. Minimize the number of interactive tuned networks.



BASEBAND SPECTRUM TO FM MODULATOR



BANDWIDTH LIMITED RF SPECTRUM TO SPECTRUM



DEMODULATED BASEBAND SPECTRUM

Fig. 7B. Stereo (L or R = 4.5 KHz) plus SCA (unmod.) modulation thru narrowband RF path.

5. Use a broadband antenna system with a low standing wave ratio on the transmission line.

For more detailed information about FM modulation theory, see references (2), (3), and (4) at the end of this chapter.

PRE-EMPHASIS

The standards adopted for FM broadcasting require the use of pre-emphasis. The standard pre-emphasis curve is defined as an ideal RC network with a time constant of 75 microseconds. The 3 dB point is at a frequency of:

$$f = \frac{1}{2\,\pi\, RC} = \frac{1}{2\,\pi\, 75 \times 10^{-6}} = 2{,}122 \text{ Hz.}$$

The 75 microsecond curve and the tolerance allowed by the FCC is shown in Fig. 8A.

The reduction in receiver output noise due to the use of pre-emphasis in monophonic transmission is illustrated in Fig. 9. The noise voltage in a narrow bandwidth (for example, 1 Hz) increases directly with frequency, therefore, the power spectral density increases as the square of frequency as shown. When de-emphasis is used, the noise voltage is attenuated above 2.1 kHz so that it remains constant with frequency. The power spectral density is also constant above 2.1 kHz. The area between these curves represents the noise power that is removed by the use of de-emphasis. This diagram indicates the importance of pre-emphasis for high-fidelity transmission because the high-frequency noise at the receiver would be very much greater without de-emphasis.

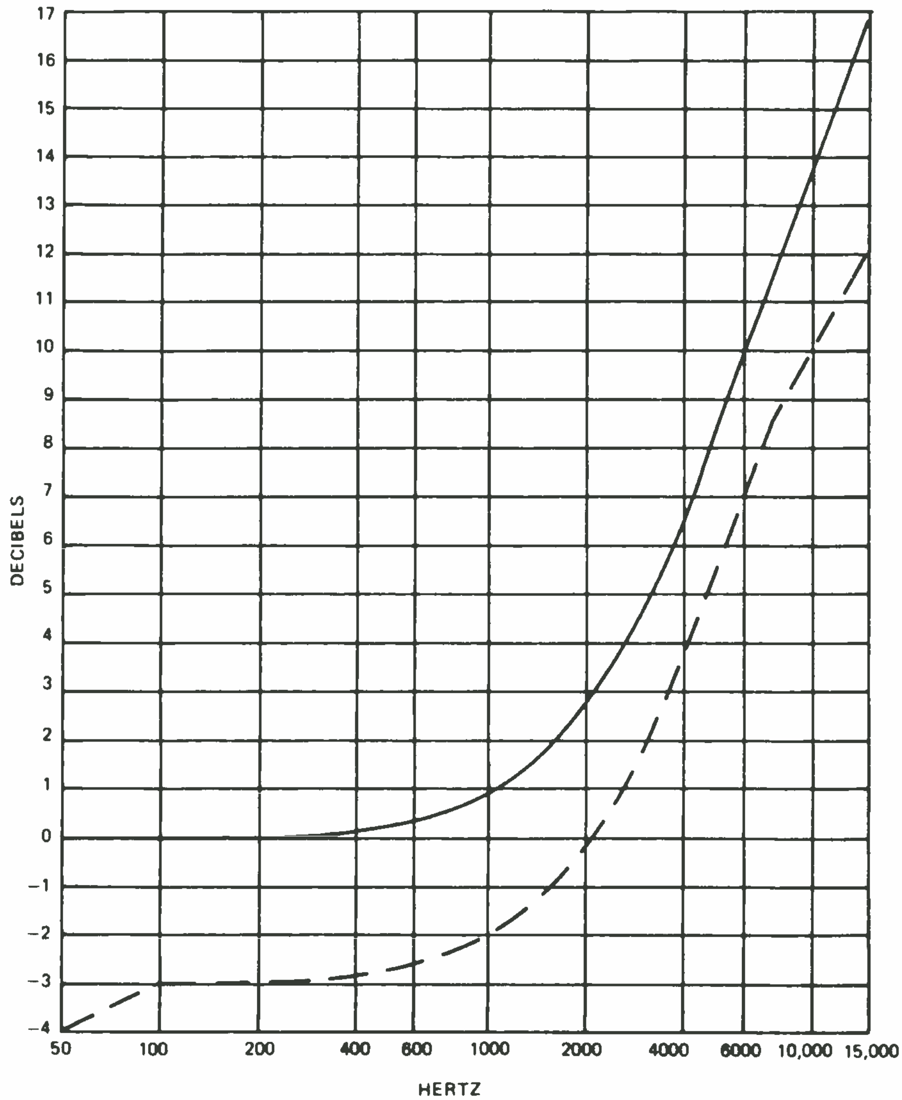


Fig. 8A. FCC Standard 75-microsecond pre-emphasis curve (solid line) and tolerance limits (solid and dashed lines).

Pre-emphasis is practical because program energy tends to peak at several kilohertz and then falls off fairly rapidly at the higher frequencies. For this reason, the higher frequencies can be boosted in amplitude without causing much increase in modulation level. There is some increase, however, so the net improvement due to pre-emphasis is the ratio of the areas under the two curves of the diagram less this reduction in audio input level required to keep within the 100 percent modulation limit. Modern audio processing equipment takes the pre-emphasis curve into account when controlling peak modulation levels.

In the section on FM modulation theory, it was mentioned that the use of pre-emphasis ahead of an FM modulator actually causes the system to behave like FM at low modulating frequencies and like PM at high modulating frequencies.

The location of the pre-emphasis network in the system depends on whether the station is operating in the monaural or stereo mode. In the case of monaural transmission, the pre-emphasis network is usually located in the FM exciter just ahead of the modulator stage. Stereo transmission requires that the FM modulator have a flat response to the composite baseband signal from the stereo generator, so the individual pre-emphasis networks for the left and right channels are located in the stereo generator before the left and right audio channels are multiplexed into the composite baseband signal. The pre-emphasis time constant may be reduced to 25 microseconds if Dolby type "B" noise reduction is being used by the station. This will assure reasonable compatibility at low modulation levels, with the standard 75 microsecond receiver de-emphasis.

FREQUENCY IN HERTZ	DECIBELS
400 HZ	0.15 dB
1,000	0.87
2,000	2.76
3,000	4.77
4,000	6.58
5,000	8.16
6,000	9.54
7,000	10.75
8,000	11.82
9,000	12.78
10,000	13.66
11,000	14.45
12,000	15.18
13,000	15.86
14,000	16.49
15,000	17.07

Fig. 8B. 75 Microsecond pre-emphasis response.

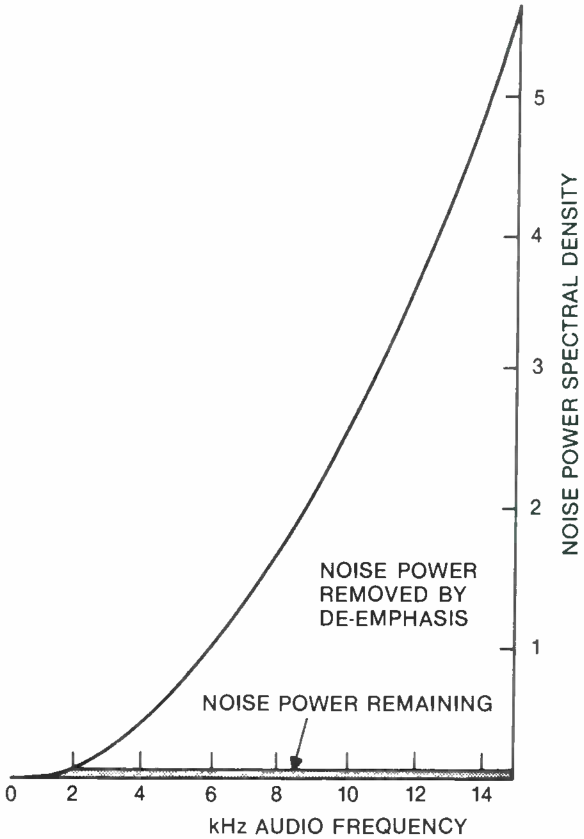


Fig. 9. Noise power spectral density before and after de-emphasis in receiver.

THE FM TRANSMITTER

The purpose of the FM transmitter is to convert one or more audio frequency (composite baseband) input signals into a frequency modulated, radio frequency signal at the desired power output level to feed into the radiating antenna system. In it's simplest form, it can be considered to be an FM modulator and an RF power amplifier packaged into one unit.

Actually the FM transmitter consists of a series of individual subsystems each having a specific function:

1. The FM exciter converts the audio/baseband into frequency modulated RF and determines the key qualities of the signal.
2. The intermediate power amplifier (IPA) is required in some transmitters to boost the RF power level up to a level sufficient to drive the final stage.
3. The final power amplifier further increases the signal level to the final value required to drive the antenna system.
4. The power supplies convert the input power from the ac line into the various dc or ac voltages and currents needed by each of these subsystems.
5. The transmitter control system monitors, protects, and provides commands to each of these subsystems so that they work together to provide the desired result.
6. The RF lowpass filter removes undesired harmonic frequencies from the transmitter's output, leaving only the fundamental output frequency.
7. The directional coupler provides an indication of the power being delivered to and reflected from the antenna system.

Fig. 10 shows a simplified block diagram of a typical FM transmitter.

FM EXCITERS

The heart of an FM broadcast transmitter is its exciter. The function of the exciter is to generate and modulate the carrier wave with one or more inputs (mono, stereo, SCA) in accordance with the FCC standards. The FM modulated carrier is then amplified by a wideband amplifier to the level required by the transmitter's following stage.

Stereo transmission places the most stringent performance requirements upon the exciter. Since the exciter is the origin of the transmitter's signal, it determines most of the signal's technical characteristics; including signal-to-noise-ratio, distortion, amplitude response, phase response, and frequency stability. Waveform linearity, amplitude bandwidth, and phase linearity must be maintained within acceptable limits throughout the baseband chain from the stereo and subcarrier generators to the FM exciter's modulated oscillator. From here, the FM carrier is usually amplified in a series of class "C" non-linear power amplifiers, where any amplitude variation is removed. The amplitude and phase responses of all the RF networks which follow the exciter must also be controlled to minimize degradation of the baseband.

Before the advent of stereo broadcasting, most of the FM exciters employed phase modulation techniques. Some of these were adapted to stereo but it was difficult to achieve and maintain the performance requirements for stereo transmission.

In the FM modulation theory section, the important relationship between PM and FM was discussed. If the audio frequency response is made to fall off at the rate of 6 dB/octave across the entire audio band at the input of a phase modulator, the resulting modulated output will be identical to that of a frequency modulated carrier. In 1948, James R. Day, an associate of Edwin H. Armstrong, demonstrated a "Serrasoid" phase modulator which used pulse circuits to achieve large phase shifts with low distortion. The "Serrasoid" modulation technique, in conjunction with the audio response shaping mentioned above, became the standard method of generating wide deviation FM with low distortion. This principle called "indirect FM" was used in many FM broadcast transmitters prior to the advent of stereo broadcasting. The advantage was that the carrier frequency could be generated by a stable crystal oscillator with the phase modulation occurring in later stages. The amount of phase deviation with low distortion was limited in most systems, so it was necessary to start with a low frequency oscillator and multiply it's frequency and modulation index many times to achieve 75 kHz deviation at low modulating frequencies. This technique has been abandoned in favor of direct FM systems.

Direct FM

Direct FM is a modulation technique where the frequency of an oscillator can be made to change in proportion to an applied voltage. Such an oscillator, called a voltage tuned oscillator (VTO), was made possible by the development of varactor tuning diodes which change capacitance as their reverse bias voltage is varied (also known as a voltage controlled oscillator or VCO).

If the composite baseband signal is applied to the tuning terminal of a VTO, the result is a "direct FM" modulated oscillator. Fig. 11 is a block diagram that fits most of the modern direct FM exciters on the market.

The signal-to-noise ratio of an FM exciter is dependent on the short-term stability of the modulated oscillator. That stability is determined by factors such as operating level, noise figure

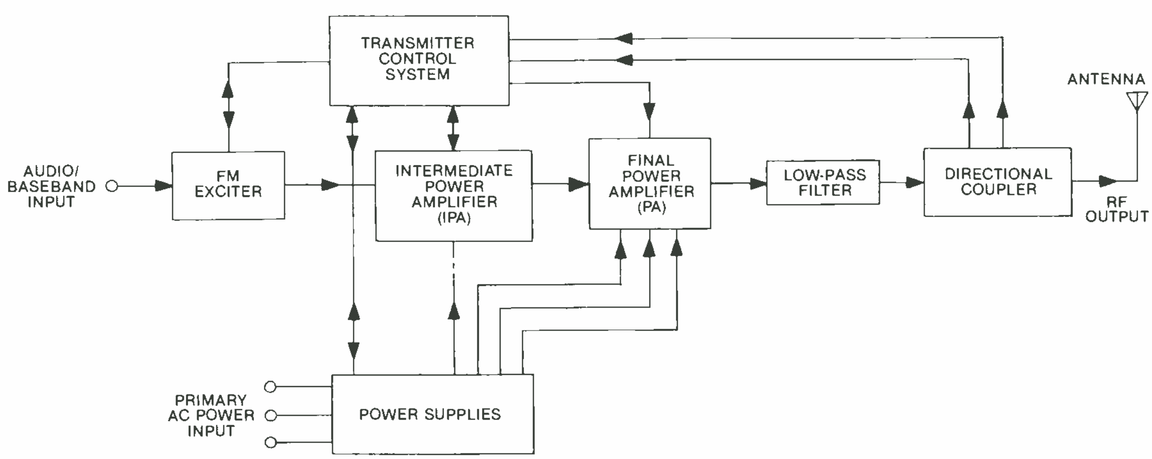


Fig. 10. Simplified block diagram of an FM broadcast transmitter.

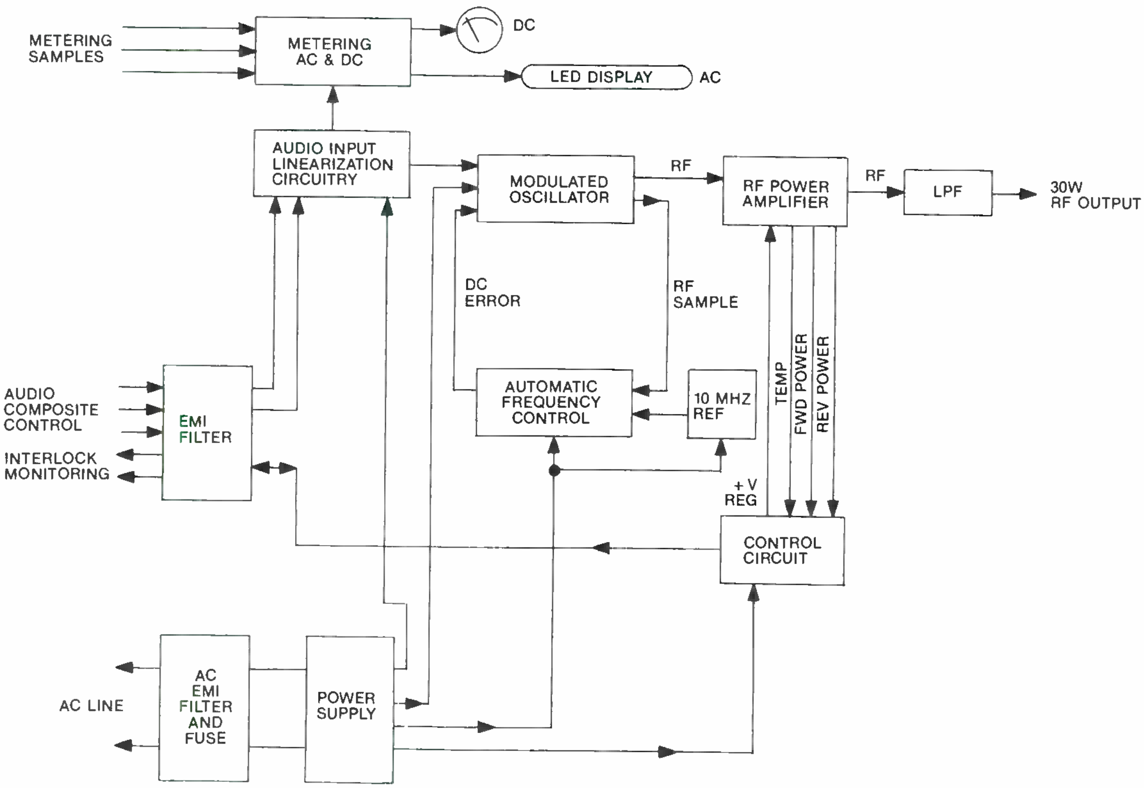


Fig. 11. FM exciter block diagram.

of the oscillator transistor, circuit configuration, method of amplitude limiting, loaded "Q" of the oscillator tank circuit, and the mechanical stability of components.

FM Modulator Linearity

Non-linearities in the FM oscillator can, by altering the waveform of the baseband signal, create distortion in the demodulated output at the receiver. A secondary effect of this distortion may include stereo crosstalk into the SCA.

The composite baseband signal is translated to a frequency modulated carrier frequency by the modulated oscillator. Frequency modulation is produced by applying the composite baseband signal to a voltage tunable RF oscillator. The modulated oscillator usually operates at the carrier frequency and is voltage tuned by varactor diodes, operating in a parallel LC circuit.

To have perfect modulation linearity, the RF output frequency (F_c) must change in direct proportion to the composite modulating voltage (V_m) applied to the varactor diodes (C_v). This requirement implies that the capacitance of the varactor diodes must change as nearly the square of the modulating voltage as shown in following relationships:

(F_c) is proportional to (V_m) (Desired linear voltage to frequency translation)

$$\text{if } F_c = \frac{1}{2\pi\sqrt{(L)(C_t)}}$$

$$\text{and: } C_v = \frac{K}{V_m^2} \ (\text{If } C_{\text{fixed}} = 0)$$

Where: F_c = Instantaneous carrier frequency
L = Inductance of resonant circuit
C_t = Total capacitance across L (C fixed + C varactors)
C_v = Capacitance of varactor tuning diodes
K = Varactor constant
V_m = Baseband modulating voltage

Unfortunately, the voltage versus capacitance characteristic of practical varactor diodes is not the desired square law relationship. All varactor-tuned oscillators have an inherently non-linear modulating characteristic. This non-linearity is very predictable and repeatable for a given circuit configuration, making correction by complementary predistortion of the modulating signal feasible. Suitable predistortion can be applied to the composite baseband signal by using a piecewise linear approximation to produce the desired complementary transfer function. Fig. 12 shows a typical network of switching diodes and resistors used for complementary predistortion of the composite baseband. Fig. 13 shows how the predistor-

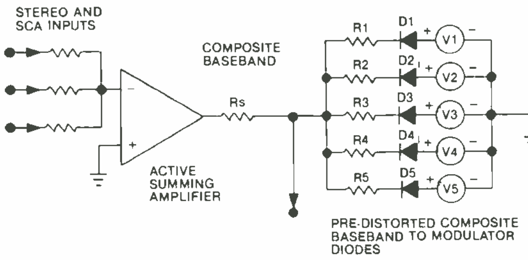


Fig. 12. Predistortion network.

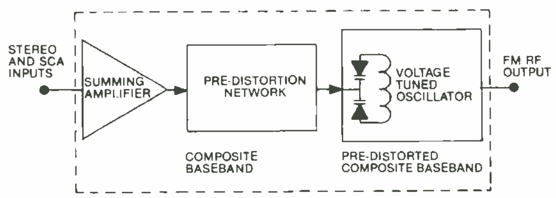


Fig. 13. Linearized FM modulator block diagram.

tion network is cascaded with a non-linear voltage-tuned oscillator to produce a linearized FM modulator.

It is also possible to improve both the linearity and signal to noise ratio of the modulated oscillator by demodulating its RF output to baseband and then feeding some of this baseband with the proper phase relationship back into the composite input of the modulator. This type of configuration places the entire modulated oscillator within a negative feedback loop and transfers the responsibility for maintaining linearity to the demodulator. Digital demodulation schemes can be made very linear, but the additional complexity and the potential problems with loop stability have limited the applications of this approach to linearization.

Modulator linearization has reduced harmonic and intermodulation distortion to less than .05% in the current generation of equipment. Any distortion of the baseband signal caused by the modulated oscillator will have secondary effects on stereo and SCA crosstalk, which are quite noticeable at the receiver in spite of the rather small amounts of distortion to the baseband. For example, if the harmonic distortion to the baseband is increased from .05% to 1.0%, as much as 26 dB additional crosstalk into the SCA can be expected.

For illustrative purposes, Fig. 14A, 14B, and 14C give representations of the fundamental and second order terms in the composite baseband spectrum with increasing amounts of harmonic distortion in the modulated oscillator. Fig. 14B shows this spectrum after 0.05% harmonic distortion has been added to each component. Note that the second order stereo (L-R) sidebands are 78 dB below 100% modulation or about 58 dB below a 67 kHz SCA with a 10% injection. With normal energy distribution in L-R and the SCA, crosstalk from stereo into the SCA will be more than 60 dB below the SCA subcarrier. Fig. 14C shows the same baseband spectrum with 1.0% harmonic distortion. The second order stereo sidebands are only 32 dB below the SCA. Crosstalk may now increase as much as 26 dB, depending on the respective energy distributions in (L-R) and the SCA.

Transient InterModulation (TIM) distortion is usually not a factor in varactor-tuned modulated oscillators. The modulation bandwidth capability is generally more than ten times the composite bandwidth and no negative feedback is used to maintain linearity.

Assuring that the composite baseband signal undergoes minimal distortion in the modulation process will suppress undesired harmonic and intermodulation products in the baseband, making the FM exciter transparent to the signals coupled into it. All exciter stages after the modulated oscillator operate as broadband amplifiers with minimal bandwidth limitations.

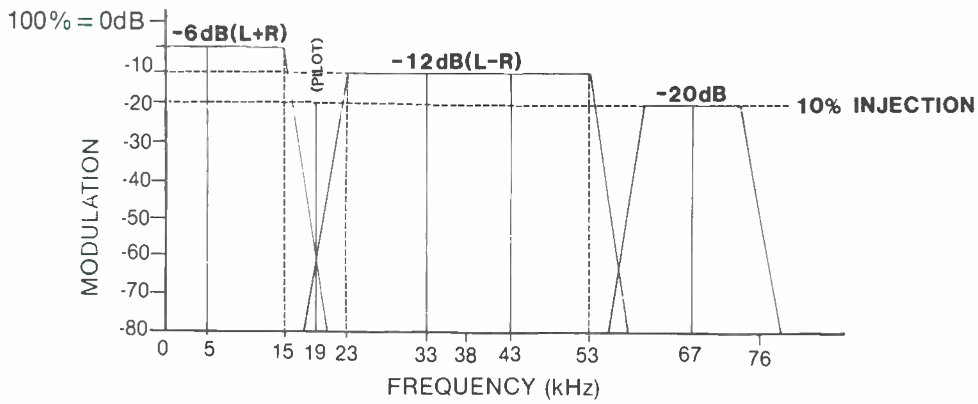
Automatic Frequency Control

The frequency stability of direct FM oscillators is not good enough to meet the FCC frequency tolerance of ±2,000 Hz. This requires an automatic frequency control system (AFC) that uses a stable crystal oscillator as the reference frequency.

The modulated oscillator need not have good long-term stability since the AFC feedback loop will correct for long-term drift to keep the average carrier frequency within limits. The modulated oscillator does need excellent short-term stability (less than 1 second) because the control loop time constant must be long enough so that the AFC circuit does not try to remove desired low frequency audio modulation. This means that the oscillator is essentially running open-loop at frequencies above a few Hertz so that the noise performance of the modulator will also be determined by the short term stability characteristics of the oscillator.

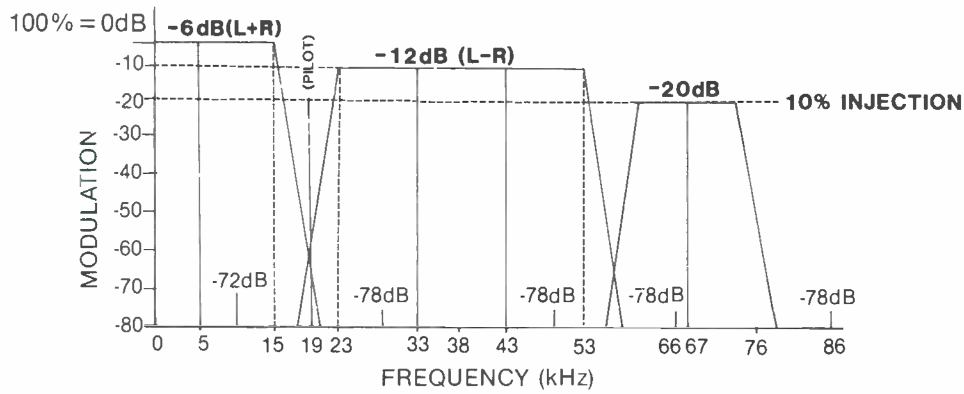
Phase-Locked-Loop Automatic Frequency Control

Phase-locked-loop (PLL) technology has provided a means of precisely controlling the carrier's average frequency while permitting wide deviation of the carrier frequency at baseband modulating frequencies. This implies that a PLL system behaves like an audio high-pass filter with higher modulating frequencies being ignored by



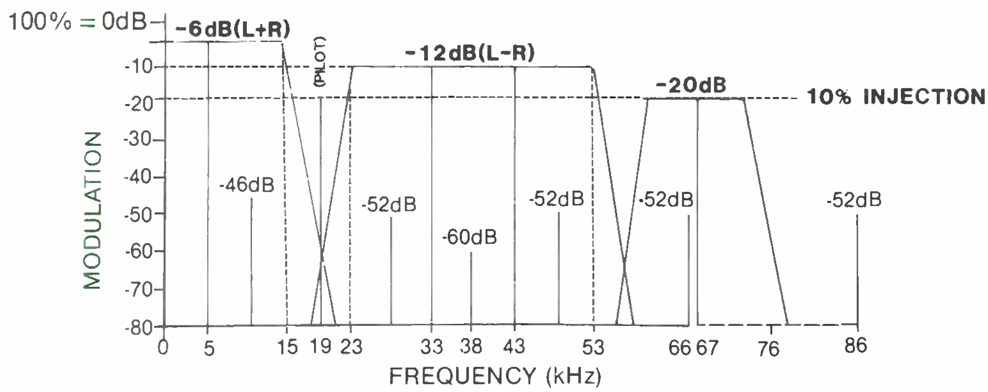
L OR R ONLY MODULATED 100% @ 5 kHz. UNMODULATED SCA @ 10% INJECTION

Fig. 14A. Ideal demodulated composite baseband spectrum with no modulator distortion.



L OR R ONLY MODULATED 100% @ 5 kHz. UNMODULATED SCA @ 10% INJECTION.
INTERFERING SECOND HARMONIC STEREO SIDEBANDS ARE 58 dB BELOW SCA.
ONLY FUNDAMENTAL AND SECOND HARMONIC TERMS ARE SHOWN.

Fig. 14B. Demodulated composite baseband spectrum with 0.05% harmonic distortion in modulator.



L OR R ONLY MODULATED 100% @ 5 kHz. UNMODULATED SCA @ 10% INJECTION.
INTERFERING SECOND HARMONIC STEREO SIDEBANDS ARE 32 dB BELOW SCA.
ONLY FUNDAMENTAL AND SECOND HARMONIC TERMS ARE SHOWN.

Fig. 14C. Demodulated composite baseband spectrum with 1.0 % harmonic distortion in modulator.

the control loop while lower frequencies are considered to be errors in the average frequency and are tracked out by the loop. An added advantage of the PLL is the ability to synthesize the desired frequency from a single reference oscillator, thereby eliminating the need to change crystals when changing the frequency of the exciter.

The block diagram shown in Fig. 15 includes the key elements in the PLL. The output of the modulated oscillator operating at the carrier frequency is digitally divided down to a frequency of a few kilohertz or even less, called the comparison frequency. Likewise, the reference crystal oscillator is also digitally divided down to the comparison frequency. The two frequencies are compared in a digital phase/frequency detector to develop an error voltage which corrects the carrier frequency of the modulated oscillator. The reason for dividing the modulated oscillator frequency so many times is to reduce the modulation index enough to limit the peak phase deviation at the comparison frequency to a value that will not exceed the linear range of the phase/frequency detector. If the linear range is exceeded, the loop will lose lock. This is why some exciters may lose AFC lock in the presence of low frequency modulation components.

The phase detector output is integrated and low-pass filtered to remove the comparison frequency and all other frequency components above a few Hertz so that the AFC circuit does not try to track-out low frequency modulation. Some FM exciters use a dual-speed PLL in order to keep the loop turn-over frequency low enough to maintain good amplitude and phase response at 30 Hz, while also providing quick lock-up time. The PLL error correction circuitry must respond quickly during the initial frequency scan of the FM band to achieve lock-up to the precision reference oscillator in a few seconds. The loop bandwidth is wide during acquisition and lock-up. After lock is achieved, the bandwidth is reduced to provide the optimum modulation characteristic.

The reference oscillator is usually temperature compensated and requires no warm-up to maintain ±3 PPM or better accuracy over the operating temperature range. 10 MHz is often selected as the reference frequency for convenient comparison to international frequency standards. For more information about PLL frequency synthesizers see reference (5) at the end of this chapter.

Exciter Metering

Metering of important operating parameters can be provided by a combination of analog metering and a digital LED display. Steady-state parameters are usually selected by multiposition

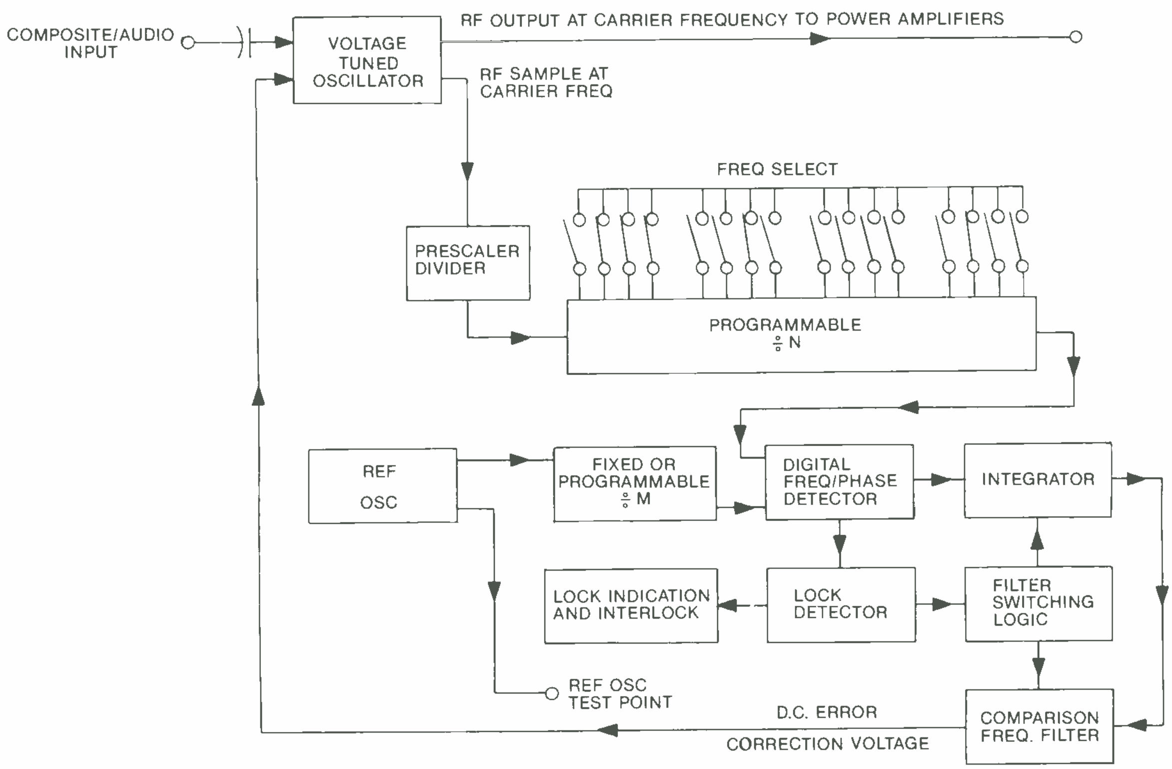


Fig. 15. Phase-locked-loop frequency synthesizer.

switch and displayed on a conventional analog multimeter. Typical steady-state functions include regulated, pre-regulated, and unregulated supply voltages; the AFC control voltage; RF power amplifier collector voltage and current; forward output power; and reflected power.

Either a color-coded LED display or a peak reading analog meter are usually provided to constantly monitor the time varying composite signal applied to the modulated oscillator. In either case, a high-speed peak detector gives accurate peak readings on signals from dc to 100 kHz. A one-shot multivibrator circuit provides a clear indication of short transient peaks exceeding 100 percent modulation.

Exciter Packaging

Protection of sensitive circuits within an FM exciter from external electromagnetic interference is important because the unit is often located in the near field of multiple broadcast antennas operating over a broad range of frequencies. The exciter should be protected from conducted EMI by use of RC and/or LC filters on all leads entering the cabinet, including the ac line. Additionally, the power transformer may have an electrostatic shield between the primary and secondary windings. The modulated oscillator is often shock mounted to prevent the transmission of mechanical vibrations from the transmitter's blower. This avoids microphonic pick-up by the modulator that would degrade the FM signal to noise ratio. Magnetic shielding of the modulator is also used to prevent hum pick-up from nearby transformers. In some cases, a hum-bucking circuit is provided to help cancel hum induced into the modulator.

The mechanical construction of most present day exciters is designed around a plug-in modular or semi-modular approach, which allows easy removal of sub-assemblies for repair or replacement.

The exciter chassis may be mounted on pull-out slides so that all sub-assemblies are accessible while the unit continues to operate. Front-panel test jacks are often provided to allow measurement of the composite signal without removing or opening the unit.

Exciter Output Stage

The broadband RF amplifier in the exciter amplifies the output of the modulated oscillator from a power level of a few milliwatts up to an output level in the range of 5 to 30 watts. The output stage should be protected against damage by an infinite VSWR at any phase angle.

The typical RF amplifier is designed to have a bandwidth of at least 20 MHz, using successive broadband impedance matching sections for each stage. Each group of matching sections consists of micro-strip or lumped elements.

The broadband performance of the RF amplifier eliminates the need for adjustments to any particular frequency within the FM band. The exciter output is transparent to the signal generated by the modulated oscillator and the amplifier stability is enhanced under varying load conditions.

A micro-strip directional coupler is often incorporated in the RF amplifier output network. This coupler supplies information to the exciter control circuitry which provides automatic control of power output level and provides protection against operation under high VSWR conditions.

All standard FM exciters will produce at least 10 watts output so they can be used as a complete transmitter for educational stations with the addition of a harmonic filter to the output. For higher power levels, the exciter is used to drive an external power amplifier.

RF POWER AMPLIFIERS

The remainder of the FM transmitter consists of a chain of power amplifiers, each having from 6 to 20 dB of power gain. Ideally, the transmitter should have as wide a bandwidth as practical with a minimum of tuned stages. Broadband solid-state amplifiers are preferred to eliminate tuned networks in the RF path. Higher powered transmitters in the multi-kilowatt range may use multiple tube stages each with fairly low gain such as in the grounded grid configuration or a single grid driven PA stage with high gain and efficiency. The dollars/watt economics of single-tube transmitters outweigh the bandwidth benefits of solid-state transmitters at the higher power levels with present technology. Design improvements in tube-type power amplifiers have concentrated on improving bandwidth, reliability, and cost effectiveness.

TRANSMITTER POWER OUTPUT REQUIREMENTS

The FCC regulates the power of FM broadcast stations in terms of effective radiated power (ERP). The authorized ERP applies only to the horizontally polarized component of radiation. Elliptical or circular polarization is also permitted where the ERP of the vertically polarized component may be as great as the authorized horizontal component. This means that twice as much total power may be radiated and twice as much transmitter power will be required.

The transmitter power requirement can be reduced by increasing the gain of the antenna. There is, of course, an economic trade off between the cost of a higher gain antenna versus the cost of a larger transmitter and the added primary power costs. For a high ERP, it is common to use antennas with up to 12 elements which provide a power gain of about 12.6 (or 6.3 in each polarization).

The long transmission lines associated with the tall towers commonly used are a source of considerable power loss. For example, the efficiency of 2000 ft. of 3-1/8 in. rigid coax at 100 MHz is only about 62 percent.

FM transmitters are designed to operate over a range of power outputs so that with a few basic sizes any required power output can be furnished. Popular maximum ratings range from 250 watts to 60 kilowatts. Most installations use a maximum transmitter power output (TPO) of thirty kilowatts because it is more economical to achieve the maximum 100 kilowatts ERP with circular polarization by means of sufficient antenna gain.

RF POWER AMPLIFIER PERFORMANCE REQUIREMENTS

The basic function of the power amplifier is to amplify the power of the exciter output to the authorized transmitter power output level. Most of the overall transmitter performance characteristics are determined by the exciter but a few are established or affected by the power amplifier characteristics:

1. The output at harmonics of the carrier frequency is almost completely a function of the attenuation provided by the output tank circuit and output low-pass/notch filters. The limit in decibels is [43 dB + 10(log watts) dB] or 80 dB whichever is less. (73 dB for 1 KW output or 80 dB for 5 KW and higher)
2. The major source of AM noise usually originates in the last power amplifier stage. The FCC limit is 50 dB below 100 percent equivalent AM modulation.
3. The RF power output control system which must keep the output within +5 percent and -10 percent of authorized output is usually achieved in the final power amplifier.
4. Inadequate passband, particularly with respect to phase linearity across the signal bandwidth, can reduce stereo separation and cause SCA crosstalk.
5. The presence of standing waves on the transmission line to the antenna may also react with the power amplifier to cause degraded stereo separation and SCA crosstalk.

The power amplifiers should provide trouble-free service and be easy to maintain and repair. Good overall efficiency is also desirable to reduce the primary power consumption.

Power Amplifier Bandwidth Considerations

As mentioned earlier, the FM signal theoretically occupies infinite bandwidth. In practice, however, truncation of the insignificant sidebands (typically less than 1 percent of the carrier) makes the system practical by accepting a certain degree of signal degradation. The input and output tuned circuits of the PA limit the bandwidth of the FM signal. The degree of bandwidth reduction is a design constraint which affects the gain and efficiency in all tuned PA stages.

The bandwidth of an amplifier is determined by the load resistance across the tuned circuit and the output or input capacitance of the amplifier. For a single-tuned circuit, the bandwidth is proportional to the ratio of capacitive reactance to resistance:

$$BW = \frac{K}{2\pi f R_L (C)} \propto \frac{K(X_c)}{R_L}$$

Where: BW = bandwidth between half-power points (BW3)
K = proportionality constant
R_L = load resistance (appearing across tuned circuit)
C = total capacitance of tuned circuit (includes stray capacitances and output or input capacitances of the tube)
X_c = capacitive reactance of C
f = carrier frequency

The load resistance is directly related to the RF voltage swing on the tube element. For the same power and efficiency, the bandwidth can be increased if the capacitance is reduced.

INTERMEDIATE POWER AMPLIFIERS

The intermediate power amplifier (IPA) is located between the exciter and the final amplifier in higher power transmitters that require more than about 30 watts of drive to the final amplifier. The IPA may consist of one or more tubes or solid-state amplifier modules.

Interstage Coupling Circuits

The separate IPA output circuit and the final amplifier input circuit are often coupled together by a coaxial transmission line. Impedance matching is usually accomplished at either end by one of the configurations shown in Fig. 16A, 16-B, 16-C, and 16-D.

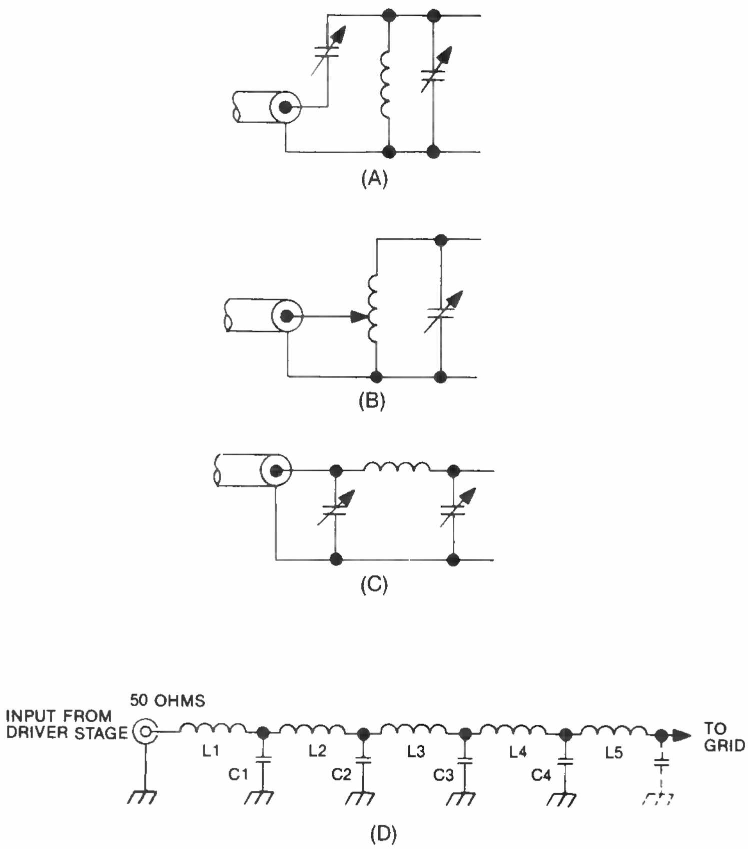


Fig. 16. Interstage RF coupling circuits.

All of the classical circuits shown in Fig. 16, except "D", require some interactive adjustment of the tuning and loading elements to provide a satisfactory impedance match for each operating frequency and RF drive level. The circuit in Fig. 16D utilizes multiple LC sections with each section providing a small step in the total impedance transformation. This technique provides a broadband impedance match without adjustment; thereby improving the transmitter's stability, ease of operation and maintainability. A single grid resonating control is sufficient to tune and match the 50 ohm driver impedance to the high input impedance of the grid over the entire 88-108 MHz FM broadcast band with a 4:1 range of RF power levels.

The interconnecting transmission line between the coupling circuits should be properly matched to avoid a high VSWR. Directional wattmeters are normally placed in the line to measure forward and reflected power from which standing wave ratio can be established. The VSWR is established by the match at the load end of the transmission line.

The transmission line matching problem is eliminated in some transmitters by integrating an IPA stage utilizing a tube(s) into the grid circuit of the final amplifier stage by having the plate of the IPA and the grid of the final tube share a common tuned circuit. This technique has the advantage of simplicity by not transforming the impedance down to 50 ohms and then back up to the grid impedance level, but does not allow the IPA to be connected directly to the antenna as a low power back-up system.

Solid-state RF power devices require a very low load impedance at the device output terminal, so that an impedance transformation that goes through the 50 ohm intermediate impedance level is required to couple these devices into the relatively high impedance of the final amplifier grid circuit. Therefore, virtually all solid-state IPA systems have a 50 ohm impedance point within the system that can be used to feed the antenna in an emergency.

High power transmitters utilizing a grounded-grid amplifier configuration in the output stage, require large amounts of drive power (typically greater than 1500 watts). The IPA may be a standard 3 KW transmitter that can also be used as a stand-by transmitter.

Most of the newer design high power transmitters only require between 150 and 600 watts of drive into a high gain final amplifier. This permits the use of a system of solid-state, wideband modules to boost the exciter's power up to the level required to drive the grid of the final tube.

Solid-State IPA Systems

A solid-state IPA almost always consists of a system of individual amplifier modules that are combined to provide the desired power output. The advantages of using several lower power modules instead of a single high power amplifier are:

1. Redundancy is provided by isolating the input and output of each module, permitting uninterrupted operation at reduced power if one or more of the modules fails.
2. The ability to repair or replace failed modules without having to go "off-the-air".
3. More effective cooling of each power device junction by splitting the concentration of heat to be dissipated into several areas instead of one small area.
4. Better isolation between the amplifier modules and the input circuit of the final power amplifier is provided by the combiner/isolator.
5. Redundant power supplies for each module improve overall reliability.

Each RF power amplifier module consists of one or more solid-state devices with broadband impedance transformation networks for input and output matching. A new generation of class "C" BIPOLAR and MOSFET devices permit the design of broadband amplifier stages that exhibit both high efficiency and the wide bandwidth necessary to cover the FM broadcast band.

Regardless of which type of solid-state device is used, the input impedance is always lower than the desired 50 ohm input impedance, so a broadband impedance transformation scheme is required. This is usually accomplished by a combination of coaxial baluns and push-pull coaxial line sections that are cross-coupled to provide 4:1 or higher transformation ratios over the FM band.

By operating two devices in push-pull, the input impedance (differential) is double that of a single ended circuit and the suppression of even order harmonics is enhanced. Two devices fed in this manner also provide some degree of redundancy within the module itself since partial RF output can be obtained with one device failed. In a similar manner, the low output impedance of these solid-state devices can be transformed up to the desired 50 ohm module output impedance where combining occurs. Fig. 17 illustrates a simplified schematic of a broadband IPA module utilizing the push-pull configuration.

IPA Splitting and Combining

There are two types of splitting/combining schemes used:

1. 90 degree hybrid splitter or combiner ("N-1" hybrids required to split or combine "N" inputs). (see section on transmitter output combining)
2. Wilkinson "N-way" in-phase splitter or combiner.

Either type of splitter/combiner must provide isolation between the individual power amplifier modules and low loss splitting or combining of the total power. By choosing the proper lengths for the coaxial interconnecting cables, either of the above methods can be configured to operate the individual modules "in-phase" so that the loading on each of the modules tracks the other modules when the impedance at the output of the combiner is varied.

The cascaded 90 degree hybrid system shown in Fig. 18 provides double isolation between the IPA and the grid circuit by first combining the

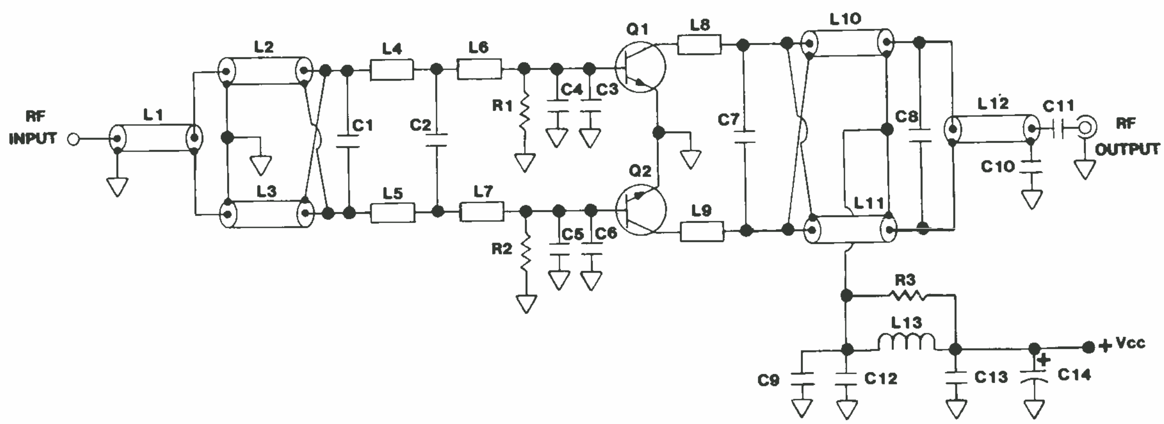


Fig. 17. Simplified schematic of a broadband intermediate power amplifier module.

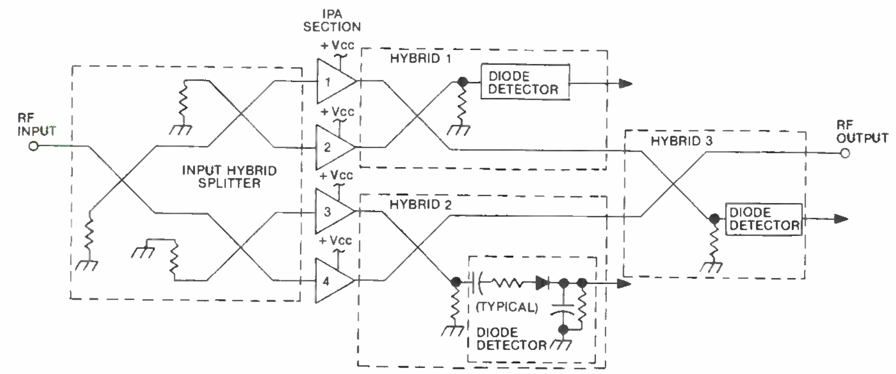


Fig. 18. Cascaded 90 degree hybrid splitting/combining system.

two pairs of amplifiers and then combining the outputs of the first two combiners. A portion of the reflected power, caused by a mismatch at the output, will be dissipated in the reject loads so that the IPA modules will operate into a lower VSWR than exists at the output. The unbalanced 50 ohm reject loads are accessible for monitoring of reject load power which is useful in determining the balance of the system. The coaxial interconnecting cables between module pairs must be offset in electrical length by one-quarter wavelength (90 degrees) at the operating frequency so that the modules operate in phase while the hybrids operate in quadrature.

The Wilkinson system shown in Fig. 19 is a simple and effective way to split and combine modules operating in phase, but usually requires a balanced reject load making reject power measurements more difficult. By adding additional coaxial balun sections to the Wilkinson/Gysel, e.g., it is possible to use unbalanced reject loads. Since the Wilkinson operates in phase, all of the coaxial interconnecting cables should be equal in length.

Since most IPA splitter/combiner systems are designed around a 50 ohm input and output impedance level, these systems can be easily used as a low power stand-by transmitter by routing the output to the antenna system. An RF low pass filter (LPF) is required only when directly feeding the antenna system. The harmonic suppression of the IPA is not as critical when driving a nonlinear power amplifier that also generates harmonics, because this stage will have its own LPF.

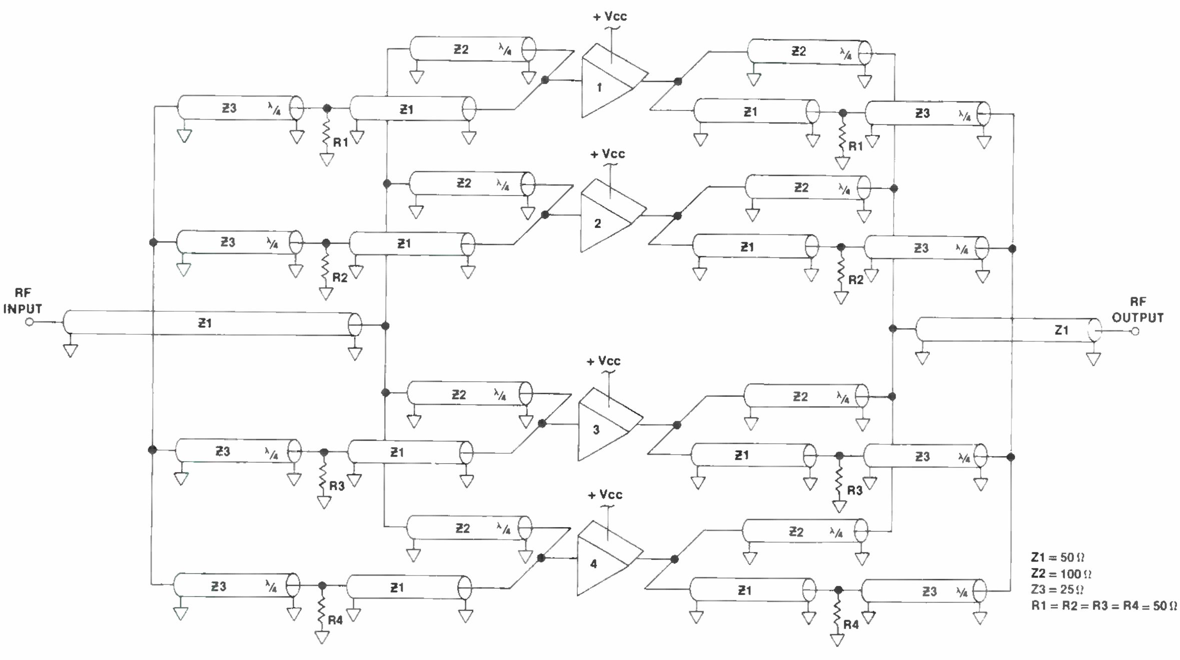


Fig. 19. Wilkinson/Gysel in-phase splitting/combining system with unbalanced reject loads.

Solid-State FM Broadcast Transmitters

The techniques used to construct IPA systems can also be used to construct a completely solid state transmitter using arrays of combined modules for the final output stage. An additional RF low pass filter is usually required to meet FCC emission requirements. Several manufacturers offer solid state FM broadcast transmitters with power outputs ranging from 100 watts up to several kilowatts, but present economic factors still favor the single tube FM transmitter for power levels above a few kilowatts. The primary advantages of a solid-state transmitter are the built-in amplifier redundancy and the ability to cover the entire FM band without the need for re-tuning. Even a tubeless transmitter is not entirely maintenance free, because high power solid-state RF devices do age and wear out like tubes, depending on the junction operating temperature and internal current density. The typical life of an RF transistor conservatively operated, is about ten years.

When broadband, solid state, amplifiers are combined for higher power, tuned output band-pass filters may still be necessary when operating in a dense RF environment in order to prevent RF intermodulation products from being generated in the PA modules.

For more information about solid state amplifiers and hybrid splitter/combiners, see references (6), (7), (8) and (9) at the end of this chapter.

VACUUM TUBE POWER AMPLIFIER CIRCUITS

The amplitude of an FM signal remains constant with modulation so that efficient Class "B" and "C" amplifiers can be used. Most exciters being manufactured at this time, provide 10 to 30 watts of output power. It is technically feasible to develop transistor amplifiers for any required power, but they are not yet economically competitive at high power levels. Additional circuitry is involved because it takes the combined output of many transistors to produce a few kilowatts of power output. For this reason, the following discussion will relate to vacuum tube amplifier circuits.

FM broadcast power amplifier circuits have evolved into two basic types. One type uses a tetrode or pentode tube in a grid-driven circuit while the other uses a high-mu triode in a cathode-driven (grounded grid) circuit.

Cathode-Driven Triode Amplifiers

The high-mu triodes being used in cathode-driven (grounded-grid) FM amplifiers were originally developed for linear SSB amplifiers. Their characteristics are well adapted to FM broadcast use because the circuit is very simple and no screen or grid bias power supplies are required. Fig. 20 shows the basic circuit configuration. In this case, the grid is connected directly to chassis ground. DC grid current is the difference between dc cathode current and dc plate current. The output tank circuit is a shorted coaxial cavity which is capacitively loaded by the tube output and stray circuit capacitance. A small capacitor is used for trimming the tuning and another small variable capacitor is used for adjusting the loading. A pi-network matches the 50 ohm input to the tube cathode.

The triodes are usually operated in the class "B" mode in order to achieve maximum power gain, which is on the order of 20 (13 dB). They can be driven into class "C" operation by providing grid bias. This increases the plate efficiency, but also requires increased drive power.

Most of the drive power into a grounded-grid amplifier is fed through the tube and appears in the stage's output. This increases the apparent efficiency so that the efficiency factor given by the manufacturer may be higher than the actual

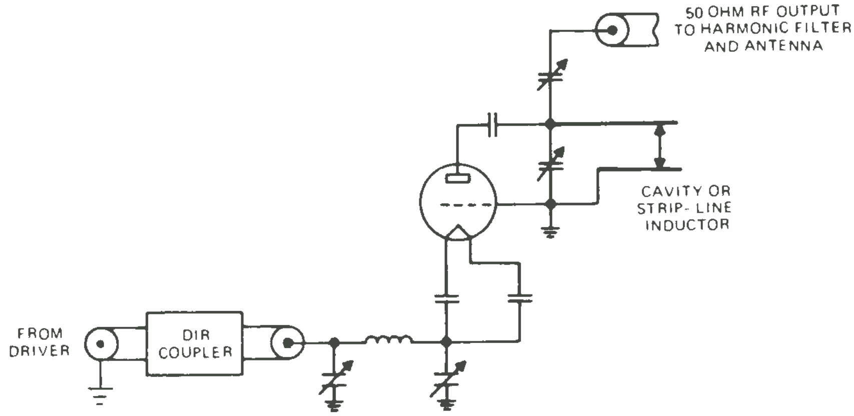


Fig. 20. Cathode-driven triode power amplifier.

plate efficiency of the tube. The true plate efficiency is determined by dividing the output power by the total input power, which includes both the dc plate input power ($I_p x E_p$) and the RF drive power. Since most of the drive power is fed through the tube, any changes in loading of the output circuit will also affect the input tuning and driver stage.

There is RF drive voltage on the cathode (filament) of the tube, so some means of de-coupling must be used to block it from the filament transformer. One method employs high current RF chokes since the inductance can be very low at this frequency range. The other commonly used method feeds the filament power through the input tank circuit inductor.

Cathode-driven stages are normally used only for the higher power stages. The first stage in a multi-tube transmitter is nearly always a tetrode because of its higher power gain.

Grounded-Grid Versus Grid-Driven Tetrode Operation

Tetrodes may also be operated in the grounded-grid configuration by placing both the control grid and the screen grid at RF ground. Higher efficiency and gain can be achieved by placing negative bias on the control grid while placing a positive voltage on the screen grid of a cathode-driven tetrode.

The input capacitance of a tetrode in a grounded-grid configuration is much less than a grid-driven configuration and the input impedance is lower providing better bandwidth.

Approximate input capacitances of some typical tubes are:

TUBE TYPE	Cin (PF)	
	Grounded Grid	Grounded Cathode
4CX 3,000A	67	140
4CX 3,500A	59	111
4CX 5,000A	53	115
4CX15,000A	67	161
4CX20,000A/8990	83	190

The typical drive power requirements, as a function of plate voltage, for a 5 kW power amplifier are:

CONFIGURATION	PLATE VOLTAGE/ DRIVE POWER	
	4500 VOLTS	5200 VOLTS
GROUNDED-GRID	340 WATTS	280 WATTS
GRID-DRIVEN	190 WATTS	140 WATTS

Television transmitters sometimes use a grounded-grid tetrode configuration to increase the input bandwidth of the amplifier. There are several trade-offs between the performance of grounded-grid and grid-driven operation of a tetrode PA with respect to gain, efficiency, amplitude bandwidth, phase bandwidth, and synchronous AM under equivalent operating conditions:

1. When driving the PA into saturation, the bandwidth of the PA is limited by the output cavity bandwidth in the grounded-grid amplifier. The PA bandwidth in the grid-driven amplifier is limited by the input circuit "Q", which is basically determined by the amount of swamping resistance.
2. Output bandwidth under saturation can be improved in either configuration by reducing the plate voltage. This involves a trade-off in efficiency with a smaller voltage swing. The bandwidth improvement can be obtained with a loss of PA gain and efficiency.
3. A grounded-grid saturated PA improves bandwidth over a grid-driven saturated PA at the expense of amplifier gain. Best performance for FM operation is obtained when the amplifier is driven into saturation where little change in output power occurs with increasing drive power. Maximum efficiency also occurs at this point.
4. The phase linearity in the 0.5 dB bandwidth is better in a grid-driven configuration. The class "C", grounded-grid, PA exhibits a more non-linear phase slope within the passband, yet has a wider amplitude bandwidth. This phenomenon is due to interaction of the input and output circuits because they are effectively connected in series in the grounded-grid configuration. The neutralized grid-driven PA provides more isolation between these networks, so they behave more like independent filters.

Grid-Driven Tetrode and Pentode Amplifiers

A small tetrode tube such as the 4CX250B is commonly used as the only amplifier stage in 250 watt transmitters and as the driver for higher power stages. The largest single stage transmitter presently available, uses a high gain 5CX1500A pentode to deliver 2.5 kW. Higher power levels require two stages (an IPA and the final PA).

Transmitters using tetrode amplifiers throughout usually have one less stage than those using triodes. Since tetrodes have higher power gain, they are driven into class "C" operation for high plate efficiency. Against these advantages is the requirement for neutralization, along with screen and bias power supplies.

Fig. 21 shows a schematic of a grid-driven tetrode amplifier. In this example, the screen is operated at dc ground potential and the cathode (filament) is operated below ground by the amount of screen voltage required. This is called grounded-screen operation. It has the advantage that stability problems due to undesired resonances in the screen bypass capacitors are eliminated. With directly heated tubes, it is necessary to use filament bypass capacitors. During grounded-screen operation, these bypass capacitors will need to have a higher breakdown voltage rating since they will have the dc screen voltage across them. The filament transformer must have additional insulation to withstand the dc screen voltage. The screen power supply provides a negative voltage in series with the cathode to ground and must have the additional capacity to handle the sume of the plate and screen currents. A coaxial cavity is used in the output circuit so that the circulating current is spread over large surfaces keeping the losses very low. This cavity is a shorted quarter-wavelength transmission line section which resonates the tube's output capacitance. The length is preset to the desired carrier frequency and then a small value variable capacitor is used to trim the system to resonance. Capacitive output coupling is used from the high RF voltage point to the 50 ohm transmission line. The 50 ohm input is capacitively coupled into the grid circuit inductor to provide the correct impedance match.

Pentode amplifiers have even higher gain than their tetrode counterparts. The circuit configuration and bias supply requirements for the pentode are similar to the tetrode since the third (suppressor) grid is tied directly to ground. The additional isolating effect of the (suppressor) grid eliminates the need for neutralization in the pentode amplifier.

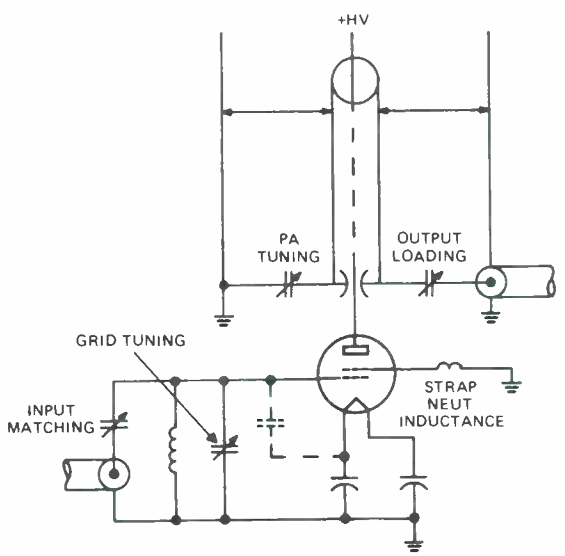


Fig. 21. Grid-driven grounded screen tetrode power amplifiers.

Impedance Matching into the Grid

The grid circuit is usually loaded (swamped) with added resistance. The purpose of this resistance is to broaden the bandwidth of the circuit by lowering the circuit "Q" and to provide a more constant load to the driver. It also makes neutralizing less critical so that the amplifier is less likely to become unstable with varying output circuit loading.

Cathode or filament lead inductance from inside the tube, through the socket and filament capacitors to ground, can heavily load the input circuit. This is caused by RF current flowing from grid to filament through the tube capacitance and then through the filament lead inductance to ground. An RF voltage is developed on the filament which in effect causes the tube to be partly cathode driven. This undesirable extra drive power requirement can be minimized by series resonating the cathode return path with the filament bypass capacitors or by minimizing the cathode to ground inductance by using a specially designed tube socket using thin-film dielectric "sandwich" capacitors for coupling and bypassing.

High power, grid driven, class "C", amplifiers require a swing of several hundred RF volts on the grid. To develop this high voltage swing, the input impedance of the grid must be increased by the grid input matching circuit. Since the capacitance between the grid and the other tube elements may be 100 picofarads or more, the capacitive reactance at 100 MHz will be very low unless the input capacitance is parallel resonated with an inductor. Bandwidth can be maximized by minimizing any additional circuit capacitance and utilizing a portion of the tube input capacitance as part of the impedance transformation network. Fig. 22A and 22B show two popular methods of resonating and matching into the grid of a high power tube. Both methods can be analyzed by recognizing that the desired impedance transformation is produced by an equivalent "L" network.

In Fig. 22A, a variable inductor (L_{in}) is used to raise the input reactance of the tube by bringing the tube input capacitance (C_{in}) almost to parallel resonance. Parallel resonance is not reached since a small amount of parallel capacitance (C_p) is required by the equivalent "L" network to transform the high impedance (Z_{in}) of the tube down to a lower value through the series matching inductor (L_s). This configuration has

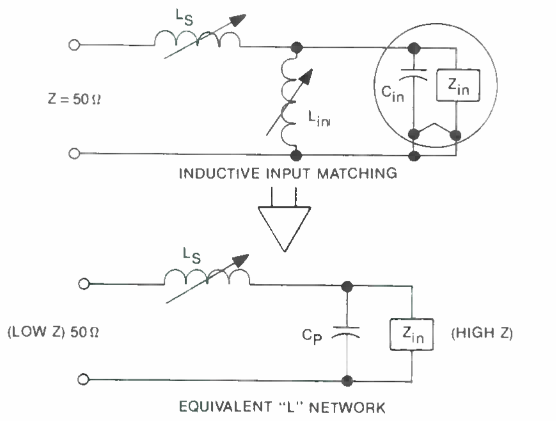


Fig. 22A. Inductive input matching.

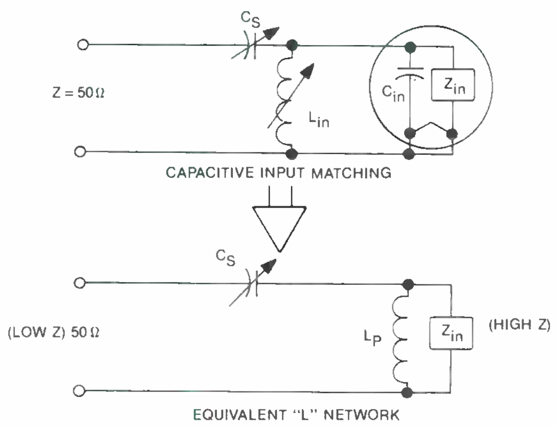


Fig. 22B. Capacitive input matching.

the advantage of providing a low-pass filter by using part of the tube's input capacitance to form (C_p), instead of an external bandwidth-reducing variable capacitor.

Fig. 22B uses variable inductor (L_{in}) to take the input capacitance (C_{in}) past parallel resonance so that the tube's input impedance becomes slightly inductive. The variable series matching capacitor (C_s) forms the rest of the equivalent "L" network. This configuration is a high-pass filter.

Neutralization

Apparently, none of the cathode-driven amplifiers utilizing triodes require neutralization. It is necessary that the grid-to-ground inductance, both internal and external to the tube, be kept very low to maintain this advantage. Omission of neutralization will allow a small amount of interaction between the output circuit and the input circuit through the plate-to-filament capacitance. This effect is not very noticeable because of the large coupling between the input and output circuits through the electron beam of the tube. Cathode-driven tetrodes have higher gain and may require some form of neutralization.

Grid-driven, high gain tetrodes need accurate neutralization for best stability and performance. Self-neutralization can be accomplished very simply by placing a small amount of inductance between the tube screen grid and ground. This inductance is usually in the form of several short, adjustable-length straps. The RF current flowing from plate to screen in the tube also flows through this screen lead inductance. This develops a small RF voltage on the screen, of the opposite phase, which cancels the voltage fed-back through the plate-to-grid capacitance. This method of lowering the self-neutralizing frequency of the tube works only if the self-neutralizing frequency of the tube/socket combination is above the desired operating frequency before the inductance is added. Special attention must be given to minimizing the inductances in the tube socket by integrating distributed bypass capacitors into the socket and cavity deck assembly. Pentodes normally do not require neutralization because the suppressor grid effectively isolates the plate from the grid.

POWER AMPLIFIER OUTPUT CIRCUITS

Usually, the output circuit consists of a "high-Q" (low loss) transmission line cavity, strip line, or a lumped inductor that resonates the tube output capacitance. A means of trimming the tuning and a means of adjusting the coupling to the output transmission line must also be provided by the output circuit. The tank circuit loaded "Q" is kept as low as practical to minimize circuit loss and to maintain as wide an RF bandwidth as possible.

The Power Amplifier Cavity

The vacuum tube power amplifier is constructed in an enclosure containing distributed tank circuit elements for minimum loss. The efficiency of the PA depends on the RF plate voltage swing, the plate current conduction angle, and the cavity efficiency. The cavity efficiency is related to the ratio of loaded to unloaded "Q" as follows:

$$N = 1 - \left(\frac{Q_L}{Q_u}\right) \times 100$$

Where: N = efficiency in percent
Q_L = loaded "Q" of cavity
Q_u = unloaded "Q" of cavity

The loaded "Q" depends on the plate load impedance and output circuit capacitance. Unloaded "Q" depends on the cavity volume and the RF resistivity of the conductors due to skin effects. A high unloaded "Q" is desirable, as is a low loaded "Q", for best efficiency. As the loaded "Q" goes up, the bandwidth decreases. For a given tube output capacitance and power level, loaded "Q" decreases with decreasing plate voltage or with increasing plate current. The increase in bandwidth at reduced plate voltage occurs because the load resistance is directly related to the RF voltage swing on the tube element. For the same power and efficiency, the bandwidth can also be increased if the output capacitance is reduced. Power tube selection and minimization of stray capacitance are areas of prime concern when designing for maximum bandwidth.

The methods used to improve the bandwidth of PA output circuits include minimizing added tuning capacitance. The ideal case would be to resonate the plate capacitance alone with a "perfect" inductor, but practical quarter-wave cavities require either the addition of a variable capacitor or a variable inductor using sliding contacts for tuning. The inherent mechanical and electrical compromises are the requirement for a plate dc blocking capacitor and the presence of maximum RF current at the grounded end of the line where the conductor may be non-homogeneous.

The Quarter-Wavelength Cavity

The quarter-wavelength coaxial cavity is the compact and popular output circuit illustrated in Fig. 23. The tube anode is coupled through a dc blocking capacitor to a shortened quarter-wave length transmission line. The tube's output capacitance is brought to resonance by the inductive component of the transmission line that is physically less than a quarter-wavelength long. Plate tuning can be accomplished either by adding end-loading capacitance at the high impedance end of the line with a variable capacitor or by sliding the ground plane at the low impedance end of the line. The plate tuning capacitor may be a sliding or rotating plate near the anode of the tube. The center conductor of the transmission line (chimney) is at dc ground while the anode of the tube operates at a high RF and dc potential. DC voltage is fed through an isolated quarter-wavelength de-coupling network inside the chimney to the anode of the tube, while the plate blocking capacitor prevents dc current flow from the anode into the chimney.

The Folded-Half-Wavelength Cavity

Another approach to VHF power amplification uses the re-entrant, folded-half-wave cavity design illustrated in Fig. 24. The dc anode voltage is applied to the lower portion of the plate line through a choke at the RF voltage null point.

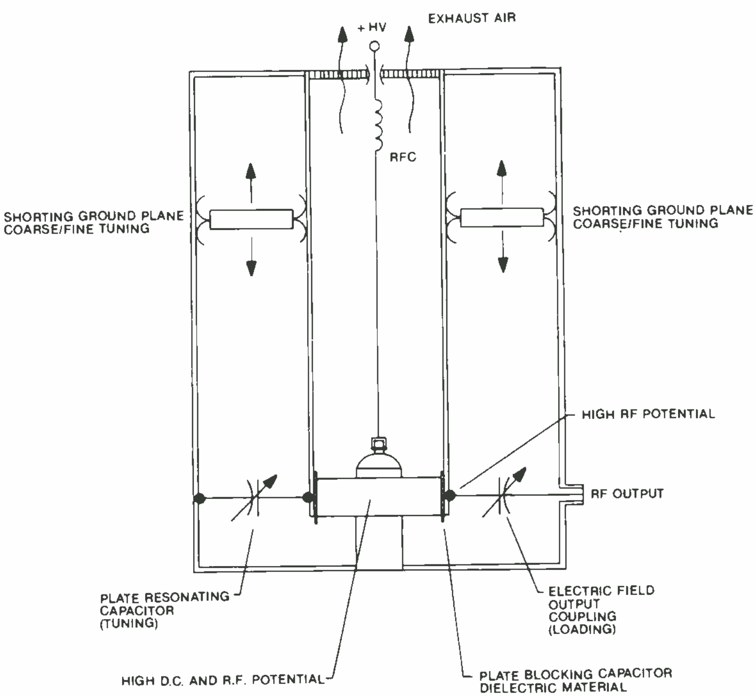


Fig. 23. The quarter-wavelength cavity.

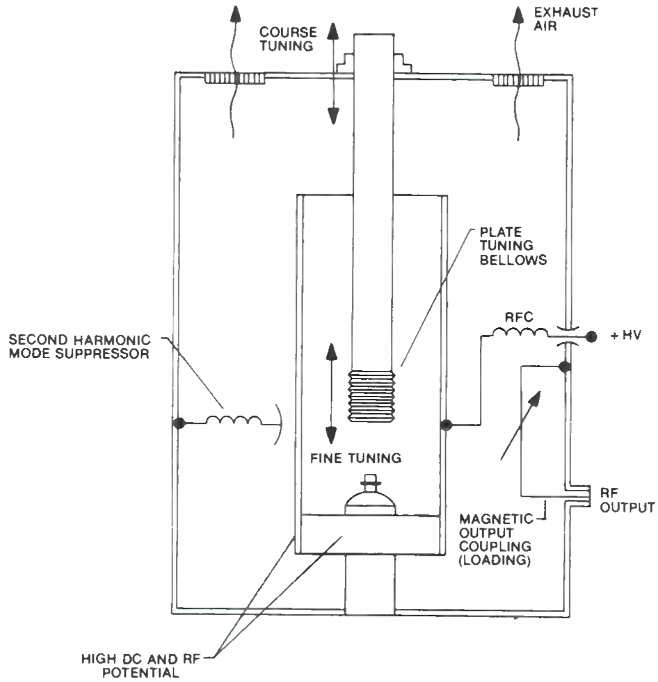


Fig. 24. The folded-half-wavelength cavity.

This RF voltage null point is reasonably consistent through the entire commercial FM broadcast band. The second harmonic voltage loop, which is located at this same point, exhibits a high source impedance and high voltage, allowing suppression with a simple series LC arrangement. This method provides about 50 dB of suppression to the second harmonic with little power loss at the fundamental frequency. The half-wave line is tuned by mechanically expanding or contracting the physical length of a flexible extension (bellows) on the end of the secondary transmission line stub, which is located concentrically within the primary transmission line (chimney). The bellows assembly is constructed of beryllium copper, providing high conductivity and long flex life.

Coarse frequency adjustment is accomplished by presetting the depth of the top secondary section of plate line into the tank cavity. A plate blocking capacitor is unnecessary since an air gap is provided between the primary line carrying the dc plate voltage and the secondary line at dc ground potential. Ohmic losses are minimized because the high RF current point is located in the central area of the homogeneous primary line where no joints, fasteners, or obstructions occur. Due to the folded nature, this configuration requires only slightly more physical space than the quarter-wave design.

Other power amplifier configurations may use lumped components or hybrid combinations with distributed transmission line elements to achieve similar results. The discrete circuit elements are chosen for their individual inductance or capacitance, instead of being operated in a purely quarter-wave or half-wave mode. Stray inductance and capacitance add to the component values resulting in the hybrid nature of these circuits.

The RF voltage and current distributions for the quarter-wavelength and the folded-half-wavelength cavities are shown in Fig. 25.

Regardless of the specific configuration, the output circuit must transform the high resonant plate impedance down to the output transmission line impedance of 50 ohms.

The bandwidth of either a quarter-wave or half-wave transmission line cavity is optimized by choosing the highest characteristic impedance mechanically allowable. The center conductor is sized for minimum impedance discontinuity and is clamped to the outer surface of the anode fins for better heat transfer. In the folded half-wave cavity, the secondary tuning line (with adjustable bellows) is sized to maintain a similar characteristic impedance without appreciable end-loading capacitance.

References (10) and (16) at the end of this chapter give detailed information about the design of tube type RF power amplifiers.

Output Coupling

Power may be coupled from a quarter-wavelength cavity to the transmission line by a capacitive probe located at the high RF voltage point located at the anode end of the quarter-wave line as shown in Fig. 23. The loaded "Q" of this circuit varies with the degree capacitive coupling. Another method of coupling power from the quarter-wavelength cavity uses a tuned loop located near the grounded (high current) end of the line.

Power may be coupled from the half-wave line by an inductive loop located in the strong fundamental magnetic field near the center of the cavity as shown in Fig. 24. One end of the output loop that couples energy to the transmission line is electrically grounded to the cavity wall. Output loading is mechanically controlled by changing the position of the loop in the magnetic field generated at the null point of the primary line structure. Multiple phosphor-bronze leaves provide the connections to the loop allowing mechanical movement without sliding contacts. The coupling to the cavity varies as the square of the effective loop area and inversely as the square of the distance of the loop center from the cavity center axis. As this loop is positioned so that it links more or less magnetic field, it determines the output loading of the transmitter.

Power Amplifier Source Impedance

At the milliwatt levels used in RF test equipment, it is customary to provide both 50 ohm source and load impedances at both ends of a coaxial transmission line. This approach minimizes any reflections on the line since both the transmitter (source) and the termination (load) will absorb reflected energy. A 50 ohm source impedance is usually provided by placing a 50 ohm build-out resistor in series with a low impedance voltage source (Thevenin equivalent). The closed circuit voltage with this configuration is exactly one-half of the open circuit voltage, meaning that half of the total available RF power is dissipated in the source resistance.

It becomes obvious that while an FM transmitter is designed to drive a 50 ohm load, it does not itself have an output source impedance of 50 ohms. In order to achieve high efficiency, the output stage is driven into voltage saturation which provides a low output source impedance. Since the low source impedance of the transmitter provides a mismatch to reflected power from the load, this power is almost totally reflected back from the transmitter output stage toward the load again.

POWER SUPPLIES

Power supplies provide the appropriate dc or ac voltages to the various subsystems with the transmitter. The range of voltages and currents provided can be from less than 5 volts at a few milliamperes, to over 10,000 volts at several amperes in a typical FM transmitter. Safety must be a prime consideration when working around potentially lethal power supplies. Power supplies must be designed with adequate bleeder resistors and interlocks to discharge high voltages before an operator can come in contact with these circuits. The degree to which the ac components are filtered out of the dc outputs of the power supplies will, in large part, determine the "asynchronous" (without FM modulation) AM noise of the FM transmitter.

FM transmitters usually contain multiple power supplies for each of the functional blocks within the system. These power supplies fall into two general categories:

1. Single-phase supplies (single input winding on the transformer).

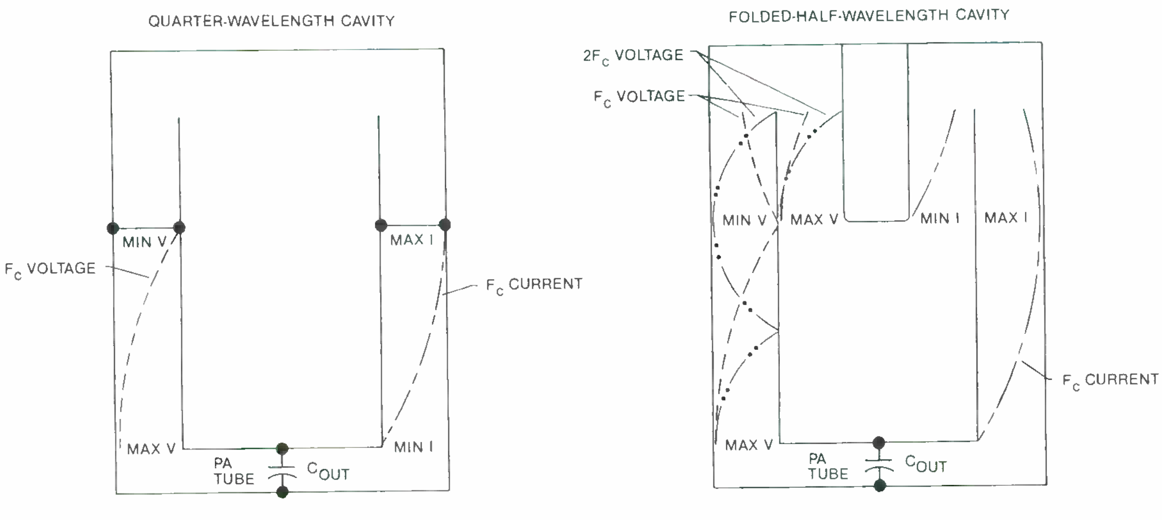


Fig. 25. Cavity RF voltage and current distributions.

2. Polyphase supplies (3 or more input windings on the transformer).

Single Phase Power Supplies

Single-phase power supplies with conventional full-wave rectification and filtering are most often used for the FM exciter, the control system, bias supplies, and the intermediate power amplifier (IPA). The filament transformer is also a single-phase transformer. A single-phase supply requires a larger filter choke to achieve the "critical inductance" requirement and a greater value of filter capacitance to maintain acceptably low ripple content compared to a polyphase supply. Large value filter components also mean that the greater stored energy in these components can have a more destructive effect if an arc-over occurs. Choke-input filter sections are normally used to help limit the in-rush current while the shunt capacitor is charging during turn-on. This also maximizes utilization of the transformer and rectifiers by keeping the charging current nearly constant, providing the best filtering action. Choke-input filters have the undesirable characteristic of poor voltage regulation over a wide range of loads. The output voltage will "soar" above the nominal value with no load unless there is enough current through the "bleeder" resistor to keep the choke in the constant current range. Fortunately, in a FM transmitter application, the load on the power supply is fairly constant since the power output of the transmitter does not vary significantly with FM modulation. In higher power transmitters where the main power source is three phase, care must be taken to balance each of the individual single phase loads among the three phases so that the total load on each of the individual phases is equal.

Polyphase Power Supplies

Polyphase power supplies are used for the final power amplifier high voltage supply in high power transmitters. Sometimes they are also used for tube or solid state IPA supplies. Large blowers used to cool transmitters are usually operated from a three phase power source. Care must be taken to make the three line connections to the blower motor in the proper sequence so that the motor will turn in proper direction.

The most common type of polyphase supply is three phase with full-wave rectification and LC filtering. Other polyphase systems encountered in broadcast equipment are usually multiples of the three phases with twelve phase rectifiers becoming more popular. The main advantages of a polyphase power supply are:

1. Division of the load current between three or more lines to reduce line losses and the size of each of the lines.
2. Greatly reduced filtering requirements after rectification due to the low ripple at the output of a polyphase full-wave rectifier.
3. Better voltage regulation with a choke input filter with typically 6% or less "soaring" from no load to full load.
4. Greater choice of output voltages from a given transformer by selection of either a DELTA or WYE configuration.

The main disadvantage of polyphase systems is their susceptibility to phase imbalance which causes degraded performance of the power supply. If significant imbalance exists in a polyphase system ripple rejection will be reduced in the polyphase rectifier with a resulting increase in AM noise.

The broadcast engineer should be particularly careful to be sure that the local utility does in fact, provide true three-phase power to the transmitter site. This can usually be verified by making sure that there are three transformers on the utility pole feeding the transmitter site. In many rural areas, the utilities are still synthesizing pseudo-three-phase service by providing the so called "open-delta" (V-V) or "Scott" (T-T) connection with two transformers instead of true three phase service. Operation on an "open-delta" service will degrade the transmitter's performance and increase the susceptibility of the transmitter to damage from transients on the line. Most transmitter manufacturers state that their warranty is void if the transmitter is connected to an "open-delta" system.

Special Power Supplies

Trends toward completely solid-state power amplification demand lower voltages at much higher currents. Voltage regulation of these high current supplies is necessary to suppress ripple, but the design of these specialized regulators is different from the typical high voltage power supply. Linear regulators are used at the lower power levels because they are low in efficiency, but they are simple and provide excellent ripple rejection without the need for suppression of switching transients. The linear regulators use series or shunt devices which change resistance dynamically to provide regulation with changes in load and therefore dissipate some of the power within the dynamic resistance.

Switch-mode regulators are used at higher power levels because they are high in efficiency, but they are more complicated and require additional suppression of the switching transients. The high efficiency comes from the digital "on" or "off" nature of the switching regulator which reduces resistive losses by using low loss reactive components to store energy during switching.

In some cases, phase control switching regulation is applied to the high voltage power supply feeding the final output tube. The regulation is accomplished by switching thyristors in the ac mains ahead of the primary winding of the transformer. As the switching duty cycle is reduced, the plate voltage is also reduced. Care must be taken to protect solid-state devices connected to the main power line from lightning transients. Heavy duty transient suppressors are available for this purpose.

Low voltage/high current power supplies contain extremely large amounts of stored energy that can be dangerous due to the high peak currents that can occur during a short circuit across a component with high stored energy. For this reason, special attention must be paid to methods of safely discharging these circuits without damage to components or injury to the operator. The voltage regulator should provide short-circuit protection with some type of current limiting. The main danger to the operator from this type of power supply is burns due to the nearly instantaneous heating of metallic tools or other conductors that accidentally get into a high current path such as a short across the filter capacitor.

For more detailed information on power supply theory, see references (2), (12), and (13) at the end of this chapter.

Step-Start

Step-start is often used in large power supplies where peak in-rush currents become high when the power supply is first turned on. These peak currents are caused by the need to overcome the hysteresis effect to initially magnetize the core of the transformer when ac power is first applied and to charge the filter capacitor. Step-start systems temporarily insert a resistance or reactance in series with the power lines to limit the current to a reasonable value until initial magnetization of the core and filter charging are completed. The system should be designed so that failure of the step-start mechanism will not prevent getting "on-the-air" without step-start. This is accomplished by using "fail-safe" systems where the current limiting components are only in series with the line during the starting interval until the main circuit is closed in parallel with the step-start circuit, then the connection to the current limiting components is opened.

TRANSMITTER CONTROL SYSTEMS

Transmitter control systems are often overlooked or given little priority in the selection and maintenance of a broadcast transmitter. The transmitter's control system does however, serve several important purposes:

1. Must provide basic on/off control of the transmitter.
2. Must provide overload protection to protect the transmitter from damage.
3. Must provide safety interlock protection to prevent injury to people and accessory equipment such as RF switching equipment or RF loads.
4. Must provide a means for controlling the transmitter output power.
5. Must provide remote control capability and interfacing at installations where the transmitter is not at the same location as the control operator.
6. Must provide warm-up and cool-down timing sequences of filaments or other time sensitive operations.
7. May provide status indications of overloads or other critical parameters.
8. May provide automatic regulation of the transmitter output power.
9. May provide diagnostic indications to aid in adjustment and maintenance.
10. May provide totally automatic operation of the transmitter plant (ATS).

The transmitter's ability to stay "on-the-air" will only be as good as the reliability of the control system, so it is easy to recognize the importance of the selection and correct operation of the transmitter control system.

Early transmitter technology relied on simple relay logic combined with electromechanical contactors to control the transmitter. The speed and variety of overload protection was limited and diagnostics were not available. While these systems had the virtue of being simple and immune to RFI, the relay logic required continuing maintenance in order to keep the contacts clean and adjusted.

Later, the transmitter control logic relays were replaced by solid-state digital logic. This provided much more flexibility and reliability, but also raised concerns about the ruggedness and ability of solid-state logic to survive lightning strikes. Although this concern is valid, modern solid-state logic systems are well protected against damage by optical isolation, shunt protection techniques, and RFI filtering. Operating experience with the current generation of transmitters has proven that a properly designed solid-state control system is far more reliable than its relay predecessors.

Automatic Power Control

Many transmitters also provide automatic power control (APC) circuitry to maintain the transmitter's power output within preset limits by correcting for changes in line voltage, component aging, or small amounts of drift in operating parameters. The APC circuitry compares a sample of the transmitter output power against a reference and then adjusts the RF drive or other voltages within the transmitter to bring the output power within tolerance. Some of the more sophisticated APC circuits also provide proportional VSWR fold-back of the transmitter output power. If a sample of the reflected power on the transmission line exceeds a safe limit, the transmitter output power is proportionally reduced to a safe level until the problem is resolved. This feature prevents lost air time during antenna icing or other limited VSWR situations. All standard APC circuits should provide fast-acting shutdown of the transmitter during a catastrophic failure of the antenna system such as a short or open circuit. A well designed APC system should operate like a damped analog feedback system with fast response and low overshoot.

Microprocessor Control Systems

Recently, microprocessor technology has been applied to broadcast transmitter control systems. Microprocessor based control systems can expand the functions of the transmitter controller from basic housekeeping duties into powerful self diagnostics, controller redundancy, integrated remote control, user customization, and even self-correction of operational problems.

Microprocessor technology lends itself well to industrial control applications like broadcast transmitters because the hardware can be made just as reliable as hard-wired digital logic, but has the advantage of allowing changes and growth in the operational features by simple changes in software instructions without requiring a complete redesign of the hardware.

Several transmitter manufacturers are marketing transmitters with microprocessor based control systems. Some of the typical features that distinguish these control systems from non-microprocessor systems are:

1. Built-in trouble tree with fault location and diagnostic readouts often using plain English messages (user-friendly).
2. Simultaneous readouts of all operating parameters.
3. Real time calculation of efficiency, dissipation, VSWR, and other parameters requiring mathematical operations.
4. Built-in clock/calendar for logging changes in operating status, power failures, and overloads.
5. Tolerance flagging on key operating parameters as warnings for logging and for preventive maintenance.
6. The ability to communicate with the outside world for remote control or logging purposes through a standard serial interface.
7. The customer has the ability to customize the system features to the individual requirements of the station through the use of software menus.
8. Tuning aids which will allow the operator to adjust the system for peak efficiency, minimum dissipation, and minimum VSWR by means of a real time display of these calculated parameters.

The method of communicating information to the operator varies from one system to another, but most use LED or LCD readouts with codes or alphanumeric messages. Some microprocessor controllers use CRT displays so that a large amount of information can be displayed in several different formats.

Controller Back-Up Systems

A certain degree of redundancy is desirable in the transmitter control system so that the transmitter can stay "on-the-air" even if a portion of the system fails. There are several approaches being used in present equipment to provide back-up systems. A multi-level hierarchy can be used which automatically hands over basic control functions to a primary controller in the event of a problem in the microprocessor hardware or software. Good system design separates diagnostic and supervisory functions from basic control functions so that a failure in a higher level function will not affect the ability of the system to remain "on-the-air" without interruption. Watchdog circuits and software are embedded within the control system to detect failures and initiate corrective action before an interruption in service occurs.

It is also possible to have two entirely independent microprocessor systems operating in parallel with only one system in charge at one time. Sophisticated software is required to determine if the two systems agree and if they don't agree, an algorithm must be used to decide which one is right so that control can be delegated to the properly functioning system.

The ability to quickly substitute another replacement control system or a simplified controller bypass unit is useful in the event of total failure of the control system.

Remote Control Interfacing

Regardless of the type of control system used, the ability to interface easily with standard remote control systems is very important. Most modern systems interface with parallel control lines for each individual function requiring a momentary contact closure of 24 volts or less at a current of 50 milliamperes or less. These levels are compatible with relay logic or optically isolated solid state logic. Analog levels output from the transmitter for remote meter readings generally are fully buffered and fall into the range of 0-10 volts dc for a full scale reading at an impedance level of less than 10K ohms.

With the advent of microprocessor-based control systems there is a trend toward using standard computer asynchronous serial interfacing instead of parallel interfacing. Serial interfacing reduces the number of connections to the transmitter and can carry both control functions and digitized meter readings through the same interface. By converting analog information into digital information before transmission to the remote control point, the need for calibration and re-calibration of the remote metering point is reduced. Microprocessor based control systems also allow the remote control system functions to be integrated directly into the transmitter itself giving the remote control point access to more in-depth information about the transmitter than is possible by interfacing with an external remote control system.

For more specific information, see the chapter on remote control systems.

RF OUTPUT LOW-PASS FILTERS

The high efficiency, non-linear RF power amplifiers used in FM broadcast transmitters generate significant amounts of energy on frequencies that are integer multiples (harmonics) of the desired fundamental frequency. The output circuit alone does not provide enough harmonic attenuation to meet FCC Regulations. To comply with Part 73 of the FCC Rules and Regulations and to prevent interference to other services, a low-pass filter must be installed in the transmission line at the output of the transmitter. The FM band is narrow enough that one low-pass filter design can be used for any FM channel carrier frequency. These filters usually consist of multiple LC sections arranged so that frequencies within the FM band are passed with little attenuation (typically 0.1 dB or less) while frequencies above the FM band are highly attenuated (60 dB or more).

The most common type of filter used in this application is called a "reflective" filter, meaning that the frequency components outside the passband are reflected back out of the filter toward the source because it provides a mismatch at these undesired frequencies. The filter can be constructed using either "lumped" inductors and capacitors or by using a section of non-constant impedance transmission line to form "distributed" inductors and capacitors. The filters designed for low power transmitters often employ "lumped" elements (coils and capacitors) because they are compact and can be integrated into the transmitter cabinet. The distributed type of filter is most often used with high power FM broadcast transmitters because of its simplicity, extreme ruggedness, and ability to handle higher power levels. The distributed filter does have the disadvantage of having larger physical dimensions than a similar lumped filter, which may necessitate mounting the filter external to the transmitter cabinet. Fig. 26 shows a cut-away view of a typical distributed low-pass filter. Note that the areas where the center conductor of the transmission line is smaller than that required for the input Z_0 are inductive, while the areas where the center conductor is larger in diameter are capacitive.

When two filters (such as the output cavity and the harmonic filter) are connected together by a transmission line, the total harmonic attenuation will vary with interconnecting line length. The attenuation characteristics of the harmonic filter are specified for the condition where both the source and load impedances are equal to the desired transmission line impedance.

In actual use, the source impedance at the output of the tank circuit is much less than the 50 ohm load impedance presented by a properly terminated filter. If an unfortunate length of line is selected, the harmonic attenuation may be insufficient and the transmitter tuning may be affected. This undesirable condition can be corrected by changing the line length by approximately one-quarter wavelength. The line length between the tank circuit and harmonic filter is

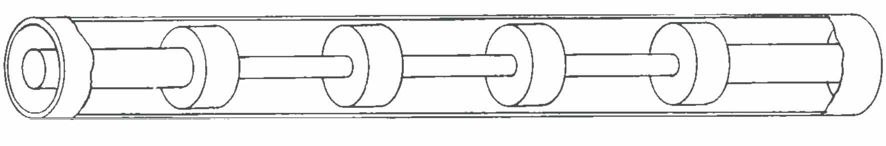


Fig. 26. Cut away view of a distributed low-pass filter.

usually supplied pre-cut to a value known to be satisfactory by the transmitter manufacturer.

Notch Filters

In some cases, a second harmonic notch filter is required in addition to the low-pass filter because the second harmonic component from amplifier is high and the cut-off slope of the low-pass filter is not steep enough to provide sufficient second harmonic attenuation. The additional attenuation required (typically 30 dB) can be provided by an absorptive notch filter which places a short circuit across the transmission line at the second harmonic while providing a high impedance at the fundamental. A one-quarter wavelength (at the fundamental frequency) shorted coaxial stub is often used for this function.

Some types of transmitters, including those using the folded half-wavelength cavity, have internal second harmonic mode suppressors which eliminate the need for an external notch filter.

TRANSMISSION LINE POWER AND SWR MEASUREMENTS

Directional wattmeters are instruments that measure the forward power (P_f) and reflected power (P_r) in a transmission line. The net power delivered to the load (antenna) is (P_f-P_r). If the transmission line is perfectly matched all the forward power will be absorbed by the load and there will be no reflected power. The peak voltage at each point along the line will be the same value and similarly the current at each point along the line will also have a uniform value. If the transmission line is mismatched, there will be reflected power with a resulting "standing wave" on the line. This means that the voltage and current distributions along the line will no longer be uniform with high values at certain points on the line and low values at points on the line that are one quarter wavelength away. The ratio of high value to the low value is called the standing wave ratio (SWR).

Standing Wave Ratio

The standing wave ratio (SWR) on the transmission line is related to the ratio of the forward to reflected power by the following formula:

$$SWR = \frac{1 + \sqrt{\dfrac{P_r}{P_f}}}{1 - \sqrt{\dfrac{P_r}{P_f}}}$$

This relationship is shown graphically in Fig. 27 so the approximate SWR can be obtained without computation.

The standing wave is due to the presence of two components of power, one traveling toward the load and the other reflected by the load mismatch, traveling back toward the generator.

These components are defined as:

$$P_f = \frac{(E_f)^2}{Z_o} = (I_f)^2(Z_o)$$

$$P_r = \frac{(E_r)^2}{Z_o} = (I_r)^2(Z_o)$$

$$P_n = (P_f - P_r)$$

The subscripts (f) and (r) are used to denote the forward and reflected values of power, voltage, and current; while Z_o is the characteristic impedance of the transmission line. (P_n) is the net power absorbed by the load (transmission line loss and antenna radiation).

Since the forward and reflected voltage and currents are traveling in opposite directions, they will add in phase at some point along the line length to produce a voltage maximum. One-quarter wave length along the line in either direction from this maximum, the forward and reflected components are out of phase and produce a voltage minimum. The forward and reflected components of current also add vectorially to produce a current standing wave. The magnitude of the standing wave is defined as:

$$SWR = \frac{E_{max}}{E_{min}} = \frac{I_{max}}{I_{min}}$$

The ratio of the highest voltage point on the line to the lowest voltage point on the line is a commonly used measure of system performance defined as the voltage standing wave ratio (VSWR). A VSWR of 1.0:1 means that a perfect match has been achieved while a VSWR of 2.0:1 means that a mismatch is causing approximately 11% of the power to be reflected.

At the point of reflection (the load mismatch), the phase of the reflected current is reversed 180 degrees from the forward current. The reflected voltage does not have this phase reversal. This displaces the voltage and current standing waves by 90 degrees along the line so that the (E_{max}) and (I_{min}) occur at the same points while (E_{min}) and (I_{max}) occur 90 degrees (one-quarter wavelength) away, in either direction from (E_{max}) and (I_{min}).

The fact that the reflected current is reversed in phase makes it possible to measure forward and reflected power separately using a device called

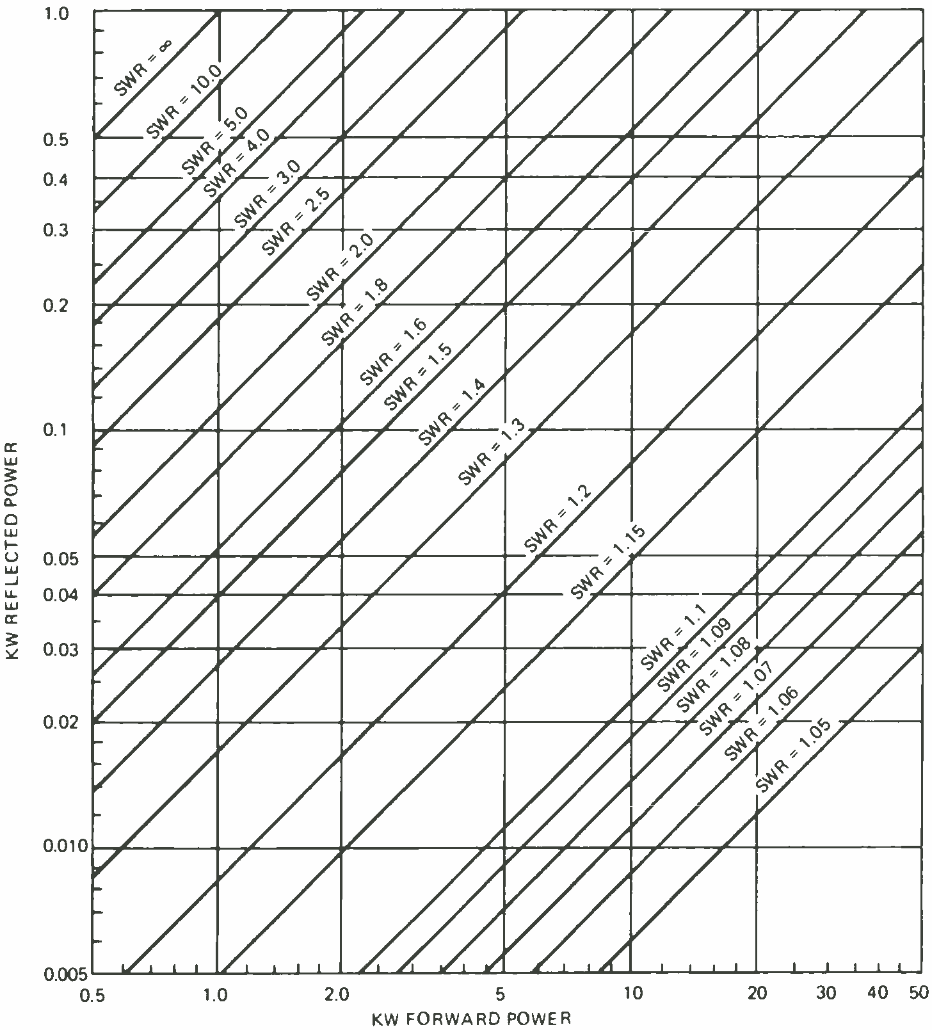


Fig. 27. Chart of SWR versus forward and reflected power.

a directional coupler. A small voltage is obtained by inductive coupling which represents the current in the transmission line. To this is added a sample of the voltage across the line that is simultaneously obtained by capacitive coupling. These two samples are adjusted to be exactly equal when the line is terminated with its characteristic impedance (no standing waves and no reflected components). The two RF samples are added, which gives a resultant RF voltage proportional to the forward components of voltage and current as illustrated in Fig. 28A.

The forward components of the samples are equal and in phase but the reflected components of voltage and current balance out. By having a second coupling section physically turned around in the opposite direction, the phase of the current sample is reversed and the reflected components add while the forward components balance out as illustrated by the vector diagram in Fig. 28B. These voltages output by the directional coupler representing the forward and reflected powers are usually rectified and buffered to feed the automatic power controller and a power indicating meter. Since power is proportional to the square of the voltage on the transmission line, the meter scale is calibrated to read the square of its input along with a diode correction factor so that forward and reflected power are read out directly.

VSWR Measurement

Although some FM transmitters can operate into a VSWR of greater than 1.8:1, the VSWR on an FM antenna transmission line should normally be kept to a value of 1.1:1 or less for good stereo performance. It takes very little reflected power to produce substantial VSWR as shown in Fig. 27. For this reason, the reflected power is usually read on a more sensitive meter position. Problems in the antenna system such as loose connections or icing may cause excessive VSWR. Instruments external to the transmitter

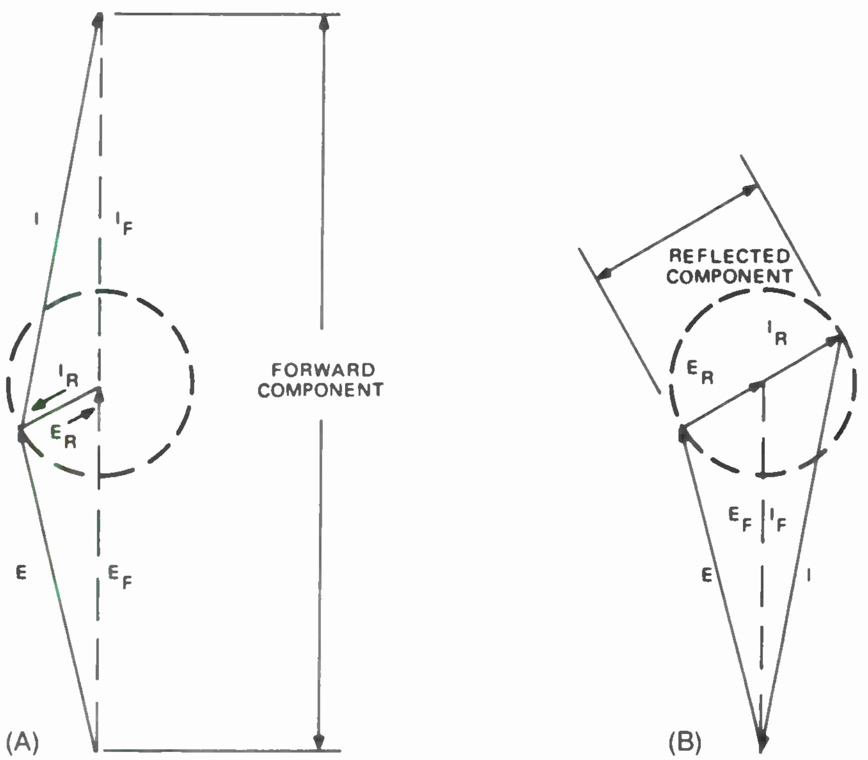


Fig. 28. Phasor addition of voltage and current samples to separate forward and reflected components.

are available that monitor reflected power and energize an alarm if it becomes excessive. As long as the transmitter power output is fairly constant, the use of reflected power to indicate excessive VSWR is simple and adequate.

COMBINED TRANSMITTERS

It is possible to combine the output of two RF power amplifiers for higher power levels. The important advantage is that the broadcast transmission is not interrupted if one amplifier fails. The radiated signal strength merely drops 6 dB until the failed amplifier is repaired and put back on the air. A dual amplifier system costs more than a single amplifier for a given total power output, but there are the economic advantages of reducing lost air time and eliminating the need for a separate standby transmitter. Automatic or manual output switching can be used to route the full power of the remaining amplifier directly to the antenna. Some stations go one step further and also install dual exciters with automatic switching so that if one exciter fails, the other unit is quickly switched into service.

Hybrid Couplers

Hybrid couplers are reciprocal 4-port devices that can be used either for splitting or combining RF sources over a wide frequency range. Fig. 29 shows an exploded view of a typical 3 dB, 90 degree hybrid coupler. The coupler consists of two identical parallel transmission lines that are coupled over a distance of approximately one-quarter wavelength and are enclosed within a single outer conductor. Ports at the same end of the coupler are in-phase while ports at opposite ends of the coupler are in quadrature (90 degree phase shift) with respect to each other.

The phase-shift between the two inputs or outputs is always 90 degrees and is almost independent of frequency. If the coupler is being used

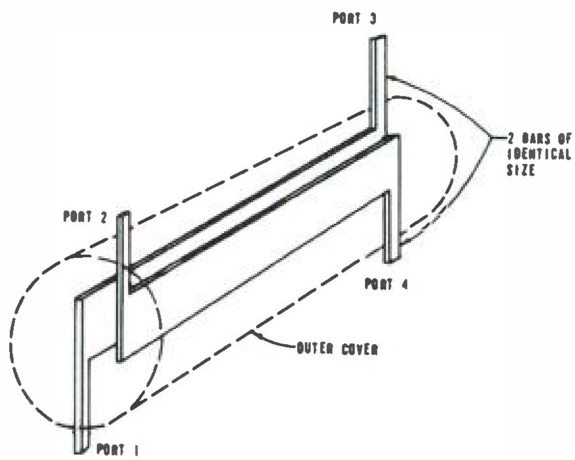


Fig. 29. Physical Model of 90° hybrid coupler.

to combine two signals into one output, these two signals must be fed to the hybrid coupler in phase quadrature. The reason this type of coupler is also called a 3 dB coupler is that when used as a power splitter, the split is equal or half-power (3 dB) between the two outputs.

Hybrid Combiners

The output hybrid combiner effectively isolates the two amplifiers from each other. Tuning adjustments can be made on one amplifier including turning it on and off without appreciably affecting the operation of the other amplifier. Good isolation is necessary so that if one transmitter fails, the other will continue to operate normally instead of in a mistuned condition.

Two of the ports on the hybrid coupler are the inputs from the power amplifiers, the sum port is the antenna output terminal, and the difference port goes to a resistive dummy load called the "reject load" since only the rejected power due to imbalance appears here. When the power fed to each of the two inputs is equal in amplitude with a phase difference of 90 degrees, the total power is delivered to the sum port (antenna). Very little of the power appears at the reject load if the phase relationship and power balance is correct. If the phase relationship is reversed between the two amplifiers, all the power is delivered to the reject load, so care must be taken to be sure that the proper one of the two possible 90 degree phase relationships is used. When all the ports on the hybrid combiner are properly terminated, isolation of 30 dB or more can be achieved between the power amplifiers. For perfect isolation between the amplifiers, the load impedance on the sum and difference ports must be exactly the same. This is approached in practice by providing a 1.0:1 VSWR with a resistive 50 ohm load for the termination (reject load) on the difference port and then reducing the VSWR on the antenna transmission line as low as possible by trimming the antenna match. This will keep the input port impedances from changing very much when one amplifier is not operating.

The input ports will present a load to each transmitter with a VSWR that is lower than the VSWR on the output transmission line. This is because part of the reflected power coming into the output port will be directed to the reject load and only a portion will be fed back into the transmitters. Fig. 30 shows the effect of output port VSWR on the input port VSWR and on the isolation between ports.

If the two inputs from the separate amplifiers are not equal in amplitude or exactly in phase quadrature, some of the power will be dissipated in the difference port reject load. The match in input power and phase is not extremely critical

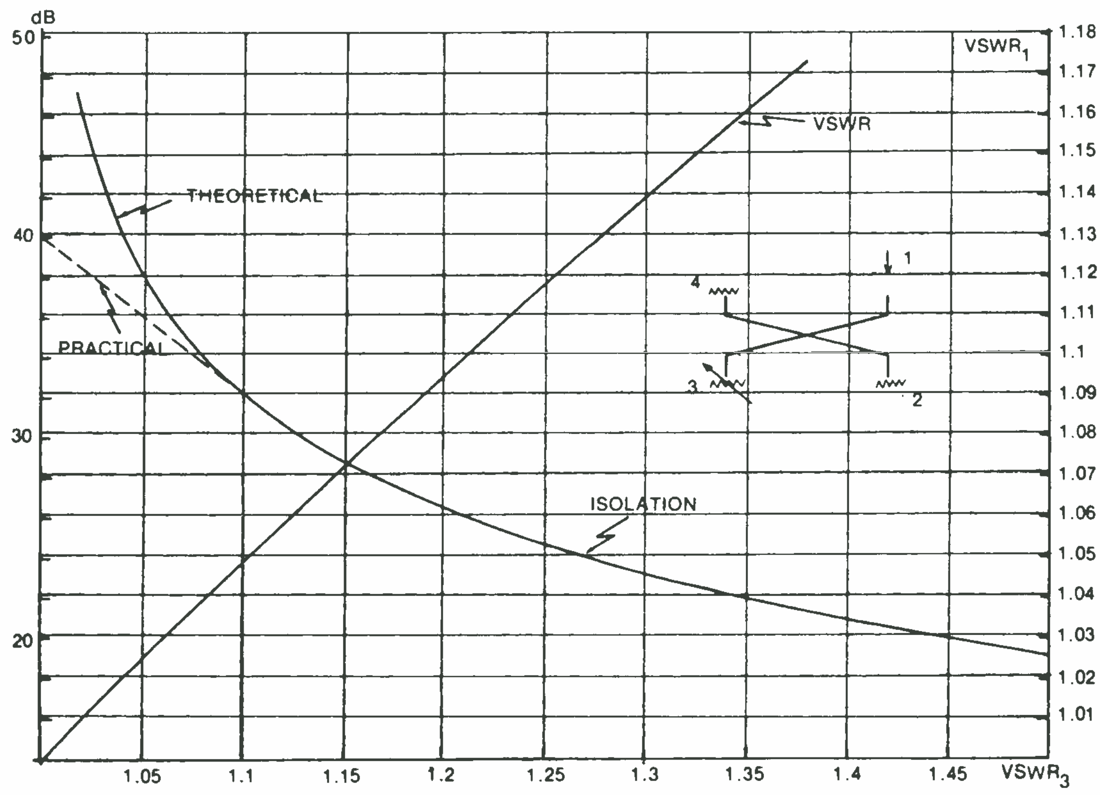


Fig. 30. Isolation of hybrid coupler.

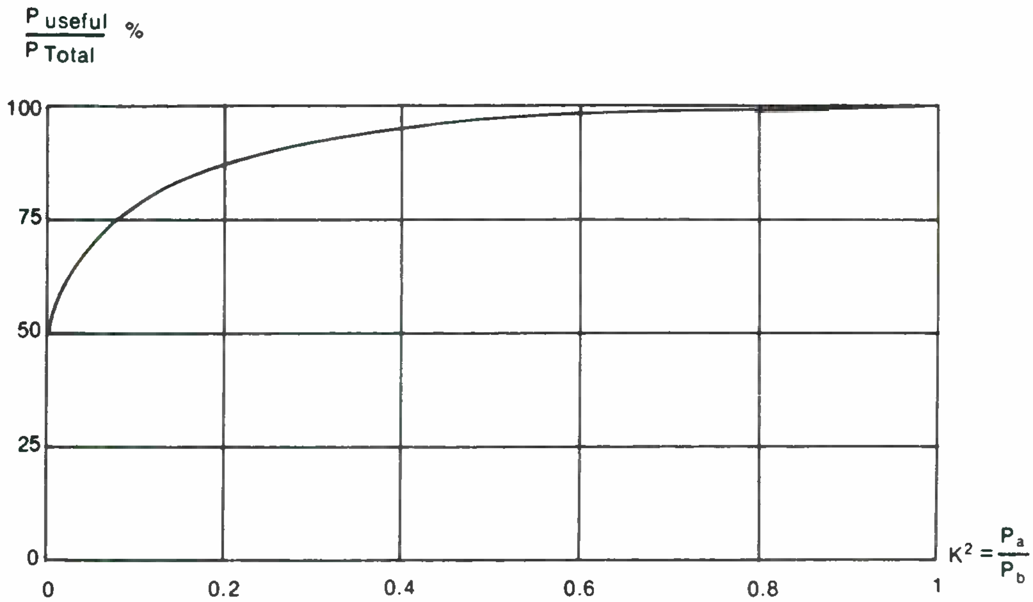


Fig. 31. Power imbalance in hybrid couplers.

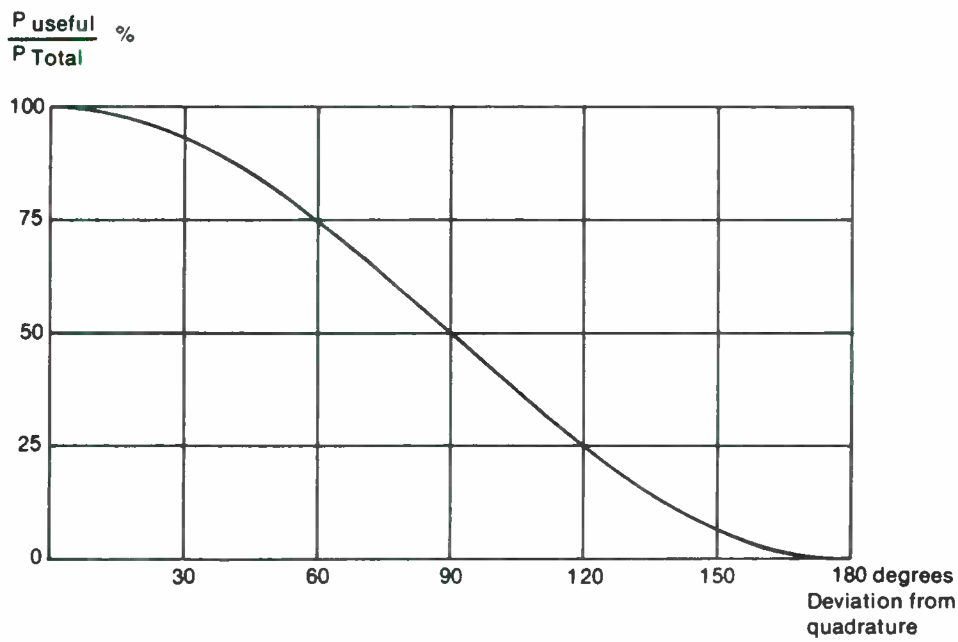


Fig. 32. Phase sensitivity, hybrid coupler.

as shown in Fig. 31 and Fig. 32. The power lost in the difference port reject load can be easily reduced to a negligible value by touching up the amplifier tuning and by adjusting the phase shift. For example, if one amplifier is delivering only half the power of the other amplifier, only about 3% of the total available power will be dissipated in the reject load and 97% is still fed to the output transmission line.

If one transmitter fails completely, half of the working amplifier's output goes to the antenna and the other half is dissipated in the difference port reject load. This is why the radiated output drops by 6 dB or to one-fourth of the original combined power. The reject load must be rated to handle a minimum of one-fourth of the total combined power, but often the reject load is rated to handle one-half the total power so that it can also be used a test load for one of the transmitters.

Hybrid Splitting of Exciter Power

Fig. 33 shows a block diagram of a pair of combined amplifiers with dual exciters. The exciters cannot be operated in parallel like the amplifiers because their RF outputs would have to be on exactly the same carrier frequency and almost exactly in phase under all modulation conditions. An automatic or manual exciter switcher is used to direct the output of the desired exciter to the combined transmitter while the other standby exciter is routed to a dummy load. The one exciter in use, feeds a hybrid splitter/phase shifter which transforms one 50 ohm input into two isolated 50 ohm outputs that have a 90 degree phase shift between them with half the power going to each output. The operation of this hybrid splitter is the reciprocal of the hybrid combiner described above. The exciter must have enough power output capability to drive both power amplifiers. In some cases an additional IPA is required between the exciter and the splitter to boost the drive level. The length of coax from the power splitter to each amplifier input must be cut to a precise length so that the amplifiers will be fed in the proper phase relationship.

Each of the power amplifiers is assumed to have equal gain and phase shift. In practice, it may be difficult to get the amplifiers tuned so that their gains and phase shifts are equal at the same time. For this reason, a line stretcher or variable phase shift network is usually included with the exciter splitter so that the station engineer will have the ability to adjust phasing independent of amplifier tuning.

For more detailed information on the theory of hybrid couplers, see the chapter on television transmitters or reference (7) at the end of this chapter.

FILTERPLEXING

The practice of having several FM stations share a single broadband antenna system is becoming more popular in recent years. In order to connect several transmitters on different frequencies together onto one antenna system, a special device called a filterplexer is required. The purpose of the filterplexer is to provide isolation between the various transmitters while efficiently combining their power into a single transmission line. This is usually accomplished by a system of band-pass filters, band-reject filters, and hybrid combiners. The isolation is required to prevent power from one transmitter from entering another transmitter with resulting spurious emissions, as well as to keep the rest of the system running in the event of the failure of one or more transmitters.

An important consideration in the design of a filterplexing system is the effect on the phase response (group delay characteristic in the passband) of each of the signals passing through the system due to individual bandwidth limitations on each of the inputs.

RF Intermodulation Between FM Broadcast Transmitters

Interference to other stations within the FM broadcast band, as well as to other services outside the broadcast band, can be caused by RF intermodulation between two or more FM broadcast transmitters. Transmitter manufacturers have begun to characterize the susceptibility of their equipment to RF intermodulation so this information will be available to the designers of filterplexing equipment.

The degree of intermodulation interference generated within a given system can be accurately predicted before the system is built if the actual mixing loss of the transmitters is available when the system is designed. Accurate data on "Mixing Loss" or "Turn-Around-Loss" not only speeds the design of filterplexing equipment, but also results in higher performance and more cost effective designs because the exact degree of isolation required is known before the system is designed. Filterplexer characteristics, as well as antenna isolation requirements, can be tailored to the specific requirements of the transmitters be-

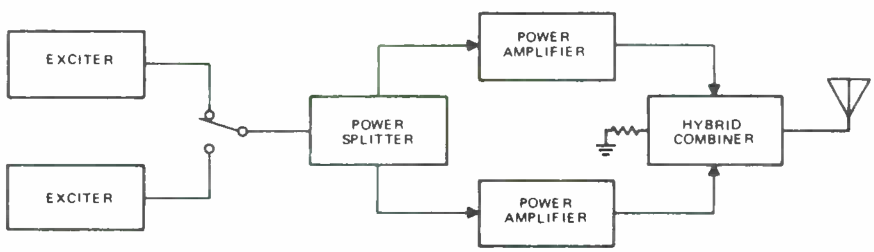


Fig. 33. Block diagram of transmitter with two power amplifiers, a hybrid combiner and dual exciters.

$$f_1 = 100.3\ Mhz. \qquad f_2 = 101.1\ Mhz.$$

$$2f_1 - f_2 = [2(100.3) - (101.1)\ ] = [200.6 - 101.1] = 99.5\ Mhz.$$

$$2f_2 - f_1 = [2(101.1) - (100.3)\ ] = [202.2 - 100.3] = 101.9\ Mhz.$$

OR

$$[f_1 - (f_2 - f_1)] = [100.3 - (101.1 - 100.3)] = [100.3 - 0.8] = 99.5\ Mhz.$$

$$[f_2 + (f_2 - f_1)] = [101.1 + (101.1 - 100.3)] = [101.1 + 0.8] = 101.9\ Mhz.$$

Fig. 34. 3rd order intermodulation products.

ing used. The end user is assured in advance of construction that the system will perform to specification without fear of overdesign or underdesign of the components within the system.

Mechanisms Which Generate RF Intermodulation Products

When two or more transmitters are coupled to each other, new spectral components are produced by mixing of the fundamental and harmonic terms of each of the desired output frequencies. For example, if only two transmitters are involved, the third order intermodulation (IM_3) terms could be generated in the following way. The output of the first transmitter (f_1) is coupled into the non-linear output stage of the second transmitter (f_2) because there is not complete isolation between the two output stages. (f_1) will mix with the second harmonic of (f_2) producing an in-band 3rd order term with a frequency of [2(f_2)-(f_1)]. In a similar fashion the other 3rd order term will be produced at a frequency of [2(f_1)-(f_2)]. This implies that the second harmonic content within each transmitter's output stage along with the specific non-linear characteristics of the output stage will have an effect on the value of the mixing loss.

It is possible, however, to generate these same 3rd order terms in another way. If the difference frequency between the two transmitters [(f_2)-(f_1)], which is an out-of-band frequency, re-mixes with either (f_1) or (f_2), the same 3rd order intermodulation frequencies are produced.

Empirical measurements indicate that the [2(f_2)-(f_1)] type of mechanism is the dominant mode generating 3rd order (IM) products in modern transmitters using a tuned cavity for the output network.

Fig. 34 shows an example of how the intermodulation product frequencies may be calculated. Fig. 35 and Fig. 36 show the resulting frequency spectra.

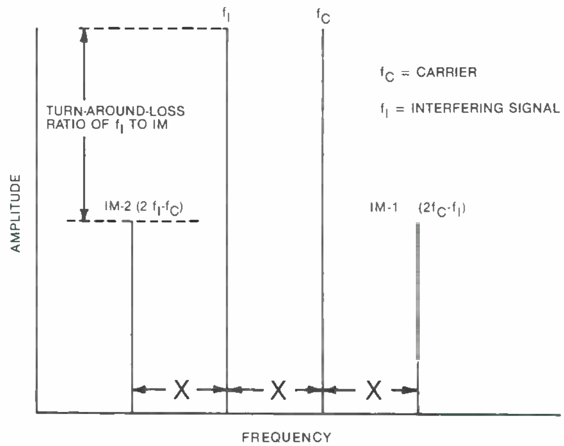


Fig. 35. Frequency spectrum of third order IM with the interfering level equal to the carrier level.

Intermodulation As A Function of "Turn-Around-Loss"

"Turn-Around-Loss" or "Mixing Loss" describes the phenomenon whereby the interfering signal mixes with the fundamental and its harmonics within the non-linear output device. This mixing occurs with a net conversion loss, hence the term "Turn-Around-Loss" has become widely used to quantify the ratio of the interfering level to the resulting (IM_3) level. A "Turn-Around-Loss" of 10 db means that the (IM_3) product fed back to the antenna system will be 10 db below the interfering signal fed into the transmitter's output stage.

"Turn-Around-Loss" will increase if the interfering signal falls outside the passband of the transmitter's output circuit, varying with the frequency separation of the desired signal and the interfering signal. This is because the interfering signal is first attenuated by the selectivity going into the non-linear device and then the (IM_3) product is further attenuated as it comes back out through the frequency selective circuit.

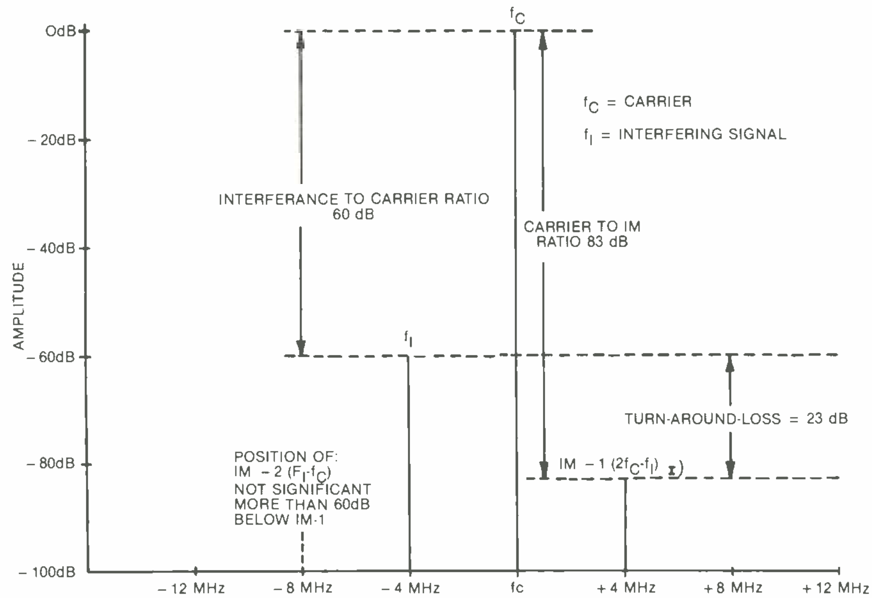


Fig. 36. Typical frequency spectrum of third order IM of a broadcast FM transmitter.

"Turn-Around-Loss" can actually be broken down into three individual parts:

1. The basic in-band conversion loss of the non-linear device.
2. The attenuation of the out-of-band interfering signal due to the selectivity of the output stage.
3. The attenuation of the resulting out-of-band (IM_3) products due to the selectivity of the output stage.

As the "Turn-Around-Loss" increases, the level of undesirable intermodulation products is reduced and the amount of isolation required between transmitters is also reduced.

The transmitter output circuit loading control directly affects the source impedance and therefore affects the efficiency of coupling the interfering signal into the output circuit where it mixes with the other frequencies present to produce (IM_3) products. Light loading reduces the amount of interference that enters the output circuit with a resulting increase in "Turn-Around-Loss". In addition, the output loading control setting will change the output circuit bandwidth (loaded "Q") and therefore also affect the amount of attenuation that out-of-band signals will encounter passing "into" and "out of" the output circuit.

Second harmonic traps or low-pass filters in the transmission line of either transmitter have little effect on the generation of intermodulation products. This is because the harmonic content of the interfering signal entering the output circuit of the transmitter has much less effect on (IM_3) generation than the harmonic content within the non-linear device itself. The resulting (IM_3) products fall within the passband of the low-pass filters and outside the reject band of the second harmonic traps, so these devices offer no attenuation to RF intermodulation products.

For more detailed information about RF intermodulation between transmitters, see the chapters on FM and TV antenna systems and reference (14) at the end of this chapter.

OPERATIONAL MEASUREMENTS

Certain operating parameters are important enough to justify regular observation. Especially important are: modulation level, carrier frequency, and output power level. The FCC requires that these operating parameters be observed as often as required to assure proper operation and to avoid causing interference to other broadcast services.

FM Modulation Measurement

The measurement of FM modulation can be accomplished with a broadcast type modulation monitor or with one of the newer modulation analyzers. Some FM exciters have built-in peak modulation displays for convenience in set-up and adjustment. Once the initial levels are correctly set, modern audio processing equipment will usually hold the modulation levels within the desired window.

Much interest and concern has developed throughout the broadcasting industry as to the best method for the determination of modulation percentage for complex program material. The ability of meter movements to follow short-duration, non-repetitive peaks accurately has received special attention. Standard modulation meter movements cannot follow modulation peaks with the required accuracy. For this reason, modulation monitors have a peak-indicating device that can be preset to flash at the particular level of interest. This device should be used instead of the meter to determine peak modulation conditions of the transmitter.

The reason for setting a peak deviation limit is so that the related occupied bandwidth does not increase to the point of interfering with stations on adjacent channels. The FCC presently enforces the modulation limit by monitoring the instantaneous peak deviation of the station as displayed on an oscilloscope. This method of measurement does not exactly correlate with the station's occupied bandwidth because the duty-cycle of the modulation peaks is not taken into account. As a result, many sophisticated "peak limiting" and "overshoot control" devices have appeared on the market to maximize loudness without exceeding the peak deviation limit, by removing the low energy peaks that would extend beyond 100% modulation. The use of these devices does cause some degradation of the audio quality and they would no longer be used if the method of modulation measurement were changed to one based on occupied bandwidth.

Carrier Frequency Measurement

The average carrier frequency must be measured and maintained to within ± 2 kHz of the assigned channel with an accurate frequency monitor. These monitors fall into two categories; analog display of the frequency error from the nominal carrier frequency and digital display of the absolute carrier frequency. The trend is toward the digital counter because of its inherent accuracy and ease of use.

The requirement that a station utilize type-approved modulation and frequency monitors has been eliminated by the FCC. Each station is still required to maintain its frequency, modulation, and audio performance within the FCC limits defined in Part 73 of the Rules and Regulations, but the responsibility for selecting the method of measurement and type of measuring equipment is now up to the station operator. Every quality conscious station will want to have the necessary equipment to accurately evaluate the signal being broadcast. There are new options available in high performance yet general purpose test equipment, now that monitor type-approval is no longer required. For instance, modern modulation analyzers provide frequency agility as well as greater measurement capability than the more specialized modulation monitors. General purpose frequency counters are now available with sufficient accuracy to measure the carrier, subcarrier, and stereo pilot frequencies directly. Spectrum analyzers provide a wide range of capability including; the measurement of harmonic and spurious frequencies at the carrier frequency, composite baseband, Bessel nulls, occupied bandwidth, stereophonic and SCA crosstalk, and AM noise.

Measurement of RF Power Output

The methods for determining RF output power are specified in Part 73 of the FCC Rules and Regulations. An accurately calibrated directional wattmeter provides an excellent means of making direct measurement of RF output power. The directional wattmeter is seldom used as the primary RF power determining method because of the requirement for recalibration to a traceable standard at regular intervals. Use of the indirect method of power measurement avoids this requirement. The FCC now permits the use of the transmitter power output meter directly, if it is periodically calibrated by comparison with the indirect method, instead of with a dummy load and standard wattmeter.

Using the indirect method, the output power is calculated from a measurement of the dc input power multiplied by the efficiency factor of the final amplifier stage. The efficiency factor is provided by the transmitter manufacturer on the final test data sheet or in the instruction manual and must be applicable to the particular frequency and power level in use. The power input to the final amplifier stage is normally defined as the product of plate voltage and plate current to this stage. Multiple output stages which are combined for the total power, must have their individual dc power inputs arithmetically summed to obtain the total power input.

The directional wattmeter can be used as a check when compared to the power output calculated by the indirect method to determine if the efficiency factor has changed due to incorrect tuning, changing antenna conditions, or a weak output device.

Logging of Transmitter Operating Parameters

FM stations may wish to log certain transmitter parameters at regular intervals in order to track equipment performance trends. The typical entries are:

1. Final amplifier plate voltage.
2. Final amplifier plate current.
3. RF transmission line current, voltage, or power.
4. Key tube parameters (grids, filament).
5. Carrier frequency.
6. Stereo pilot frequency.
7. Subcarrier(s) frequency.
8. Settings of tuning controls.

Microprocessor based control systems often provide automatic logging of all transmitter parameters including overloads and tolerance flagging.

Audio Performance Measurements

Most stations measure the performance of the entire audio system at regular intervals in order to verify the transmission quality and to locate any problem areas within the audio chain. Audio performance measurements are normally made from the audio console input to the output of the "off-air" de-modulator and include the following:

1. Audio frequency response.
2. Audio frequency harmonic distortion.
3. FM signal-to-noise ratio.
4. AM noise level.

Fig. 37 is a block diagram of a typical test set-up for audio performance measurements. All measurements are made on the composite system from microphone terminals of the console to the demodulated transmitter output. All normal program circuits, with the exception of audio limiting and compression amplifiers, must be included in the measurements. If the compression and limiting amplifiers have switches that convert them to linear, fixed-gain operation, the measurements can be made with the limiters and compression amplifiers in the linear fixed-gain mode of operation.

Measurement of Audio Frequency Response

The audio frequency response of the system is measured in reverse, that is, a constant percentage of modulation is maintained for all modulating frequencies by adjusting the amount of attenuation between the audio generator and the microphone input terminals. This is necessary because of the rising response due to pre-emphasis. Frequency response data is taken at three levels of modulation: 25 percent, 50 percent, and 100 percent. The audio voltmeter which measures the audio generator output voltage is used to maintain a constant voltage level versus frequency at the generator output terminals. The precision attenuator dials are adjusted for each modulating frequency to maintain the desired modulating level and the attenuator readings, in decibels, are recorded. Readings should be taken for the following modulating frequencies: 50, 100, 400, 1,000, 5,000, 7,500, 10,000, and 15,000 Hz.

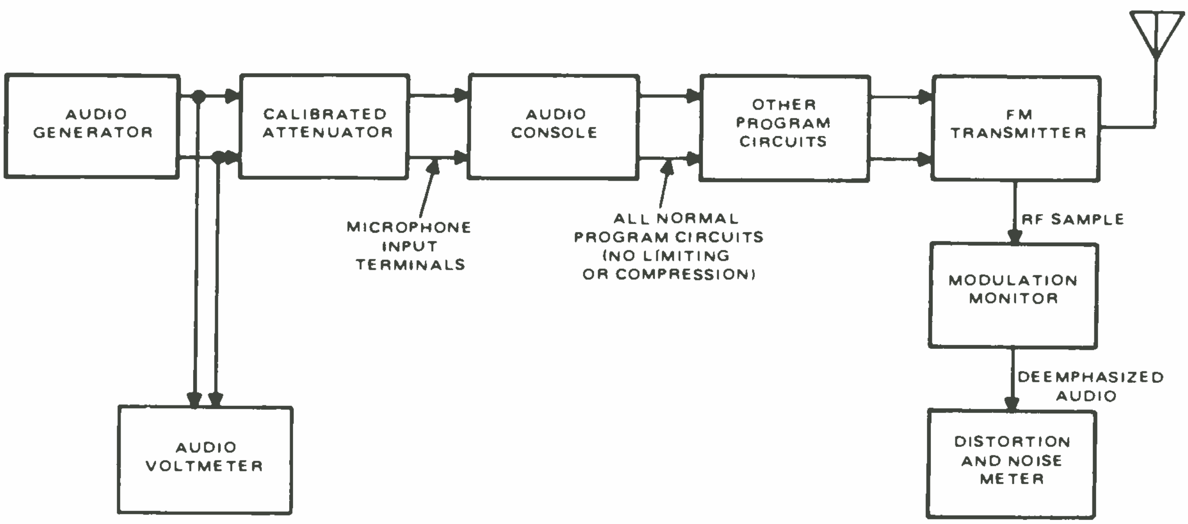


Fig. 37. Block diagram of test set-up for audio performance measurements.

When the attenuation, in decibels, is plotted versus frequency, the 75 microsecond pre-emphasis curve is obtained if the system frequency response is perfect. Deviations from the ideal response are permitted to an extent which allows the measured curve to be fitted between the upper and lower limit curves shown in Fig. 8A. The procedure for doing this is to offset the measured curve by subtracting or adding the same number of decibels from each of the measured values. This process may be repeated until a fit is obtained. If it is impossible to obtain a fit within the limits by subtracting or adding the same value of attenuation from all measured values, the system frequency response is inadequate and corrections must be made.

Measurement of Audio Harmonic Distortion

Total harmonic distortion (THD) of the system from microphone input terminals to transmitter output is measured by modulating the transmitter with sinusoidal modulating signals having low distortion and observing the harmonic content at the output of the modulation monitor. For this measurement, pre-emphasis is used in the transmitter and de-emphasis is used in the monitor. The distortion analyzer must respond to de-emphasized harmonics through 30 kHz.

The type of distortion meter normally used in this test reads not only harmonic distortion but also noise in the audio passband. For this reason, THD measurements above 5 kHz may be noise limited due to the effect of de-emphasis. If the total harmonic distortion and noise together are within the harmonic distortion limits, the system is assumed to meet its harmonic distortion requirements. A more accurate method of distortion measurement in the presence of noise is to use an audio frequency spectrum analyzer to determine the total RMS value of the individual distortion products.

The THD should be measured under the following conditions:

1. For modulating frequencies of 50, 100, 400, 1,000 and 5,000 Hz at modulating levels of 25, 50, and 100 percent modulation.

2. For modulating frequencies of 10 kHz and 15 kHz at a modulating level of 100 percent.

The preceding measurements should show that the system THD is less than 3.5 percent for modulating frequencies between 50 and 100 Hz, less than 2.5 percent for modulating frequencies between 100 and 7,500 Hz, and less than 3.0 percent for modulating frequencies between 7,500 and 15,000 Hz. Most modern equipment will pass these tests by a wide margin. If distortion levels greater than these are measured, the system requires adjustment or repair.

Measurement of Two-Tone Intermodulation Distortion

Audio intermodulation distortion (IMD) measurements are a quick and accurate way to evaluate the system performance before a complete set of THD measurements are made. If the system will pass a single 60 Hz/7 kHz, 4:1 (SMPTE-IMD) measurement at 100% modulation, it will probably pass all of the THD measurements.

Other types of difference-tone, swept two-tone, and sinewave/squarewave IMD measurements can reveal more subtle problems in the system such as transient IMD due to insufficient audio feedback bandwidth in the audio amplifier stages.

Measurement of FM Signal To Noise Ratio

The FM signal-to-noise ratio of the system is also measured from the microphone input terminals to the transmitter output. The residual noise level at the monitor output is measured with an audio voltmeter. For this measurement, pre-emphasis is employed in the transmitter and de-emphasis in the monitor. The residual audio noise level is referenced to the signal level produced by 400 Hz (L + R, mono) modulation at the 100 percent level (75 kHz deviation).

The procedure for making the FM signal-to-noise ratio measurement is as follows:

1. Modulate the transmitter with a 400 Hz sine wave applied at the microphone input terminals of the console and set the level for 100 percent modulation.
2. Read and record the audio signal level appearing at the modulation monitor output terminals. If the monitor has audio metering capability, the meter gain should be set for a 0.0 dB reference level according to the manufacturer's instructions.
3. Remove the modulation and terminate the console audio input terminals with a resistor equal to the normal microphone output impedance.
4. Read and record the residual audio noise voltage in decibels below the 400 Hz reference signal level. The measured signal-to-noise ratio should be at least 60 dB in a properly operating FM transmitter.

Measurement of AM Signal to Noise Ratios

The perfect FM transmitter will have a absolutely constant output, regardless of FM modulation or power supply variations. In practice, there is always some residual amplitude modulation of the FM transmitter. There are two types of AM signal to noise ratio that are of interest to the FM broadcast engineer:

1. Asynchronous AM signal to noise ratio measured without FM modulation is primarily related to power supply ripple.
2. Synchronous AM signal to noise ratio (incidental AM) measured with FM modulation is related to the tuning and overall bandwidth of the system.

Asynchronous AM

Residual amplitude modulation of the transmitter output, due primarily to power supply ripple, is measured with an AM envelope detector. Most FM modulation monitors include an AM detector for this purpose. The detector must include 75 microsecond de-emphasis of its output. AM noise measurements must be made directly at the transmitter output (or an accurate sample of its output). No amplifying or limiting equipment may be used between the transmitter output and the AM detector since this equipment would modify the residual AM noise level present. The residual AM noise in a properly operating FM transmitter will be 50 dB below the level which would represent 100 percent amplitude modulation of the carrier. Since the transmitter cannot be amplitude modulated, this reference must be established indirectly by a measurement of the RF carrier voltage. Refer to the instructions of the detector manufacturer to determine the reference level. Generally, the reference level is determined by setting a carrier level meter to a specified reading. If the transmitter is unable to meet the 50 dB requirement, the problem can usually be traced to a power supply component or to line imbalance in a three phase system.

Synchronous AM

Synchronous AM is a measure of the amount of incidental amplitude modulation introduced onto the carrier by the presence of FM modulation. This measurement is very useful for determining the proper tuning of the transmitter. Since all transmitters have limited bandwidth, there will be a slight drop-off in power output as the carrier frequency is swept to either side of the center frequency. This slight change in RF output level follows the waveform of the signal being applied to the FM modulator causing AM modulation in synchronization with the FM modulation. Minimizing synchronous AM will assure that the transmitter passband is centered on the FM channel.

Care must be taken when making these measurements that the test set-up does not introduce synchronous AM and give erroneous readings which would cause the operator to mistune the transmitter to compensate for errors in the measuring equipment. The input impedance of the envelope detector must provide a nearly perfect match so that there is a very low VSWR on the sampling line. Any significant VSWR on the sampling line will produce synchronous AM at the detector because the position of the voltage peak caused by the standing wave moves along this line with FM modulation.

Unfortunately the AM detectors supplied with some modulation monitors do not provide a good enough match to be useful for this measurement. Precision envelope detectors are available that present a good match (30 dB return loss) to the sampling line.

TUNING THE TRANSMITTER FOR BEST PERFORMANCE

The modern power amplifiers which have been discussed in the preceding sections, can operate with high reliability and power efficiency without compromising subcarrier performance, if they are properly adjusted.

All optimization should be done with the transmitter connected to the normal antenna system rather than to a dummy load. This is because the resistance and reactance of the antenna will be different from the dummy load and the optimum tuning point of the transmitter will shift between the two different loads. The tuning sequence is:

1. The transmitter is first tuned for normal output power and proper efficiency according to the manufacturer's instruction manual. The meter readings should closely agree with those listed on the manufacturer's final test data sheet if the transmitter is being operated at the same frequency and power level into an acceptable load.
2. A simple method for centering the transmitter passband on the carrier frequency involves adjustment for minimum synchronous AM. If the bandpass is narrow or skewed, increasing synchronous amplitude modulation of the carrier will result. A typical adjustment procedure is to FM modulate 100% at 400 Hz and fine-adjust the transmitter's grid tuning and output tuning controls for minimum 400 Hz AM modulation as detected by a wideband envelope detector (diode and line probe). It is helpful to display the demodulated output from the AM detector on an oscilloscope while

making this adjustment. Note that as the minimum point of synchronous AM is reached, the demodulated output from the AM detector will double in frequency to 800 Hz, because the fall-off in output power is symmetrical about the center frequency causing the amplitude variations to go through two complete cycles for every one FM sweep cycle. This effect is illustrated in Fig. 38. It should be possible to minimize synchronous AM while maintaining output power and efficiency in a properly designed power amplifier.

3. A more sensitive test is to tune for minimum intermodulation distortion in left-only or right-only stereo transmissions. Stereo separation will also vary with tuning.
4. For stations employing a 67 kHz SCA, transmitter tuning becomes very critical to minimizing crosstalk into the SCA. Modulate one channel only on the stereo generator to 100% with a 4.5 kHz tone. This will place the lower second harmonic (L-R) stereo sideband on top of 67 kHz SCA. Activate the SCA at normal injection level without modulation on the SCA. Tune the transmitter for minimum output from the SCA demodulator. This adjustment can also be made by listening to the residual SCA audio while normal stereo programming is being broadcast.

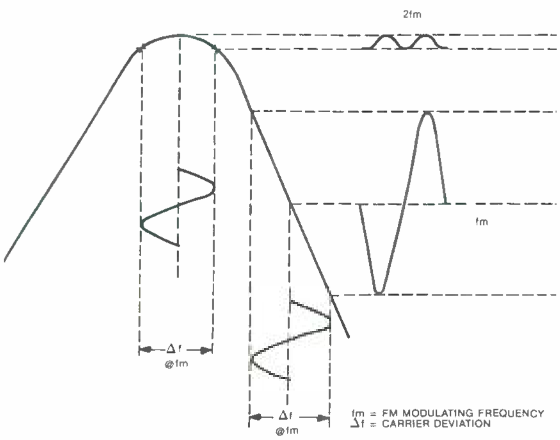


Fig. 38. Synchronous AM waveforms.

The new generation of power amplifiers have been designed to operate without compromising subcarrier performance. By providing broadband matching circuits, adjustment of these transmitters for optimal subcarrier performance (minimum crosstalk, maximum separation, etc.) is very repeatable and stable.

The field adjustment techniques are listed below in ascending order of sensitivity:

1. Tune for minimum synchronous AM noise.
2. Tune for minimum IMD in the left or right channel only.
3. Tune for minimum crosstalk into the unmodulated SCA subcarrier.

In any of these tests, the grid tuning is frequently more critical than the plate tuning. This is because the impedance match into the input capacitance of the grid becomes the bandwidth limiting factor. Even though the amplitude response appears flattened when the grid is heavily driven, the phase response has a serious effect on the higher order FM sidebands. For more information about tuning an FM transmitter for optimum performance, see references (15) and (16) at the end of this chapter.

INSTALLATION CONSIDERATIONS

Adequate planning and care in the installation of an FM broadcast transmitter and associated equipment will help avoid many problems that may be difficult and expensive to correct later. For example, poor grounds and ground loops may cause high noise levels.

Wiring the Transmitter Plant

Separate metalically shielded conduits or troughs should be provided for the audio and the ac wiring. A third conduit should be used if computer logic levels are employed for equipment control. These conduits or wiring troughs may be either overhead or below the cabinets. The ac wiring should be well separated from the audio pairs to prevent the induction of unwanted hum and noise into the audio circuits.

All audio shields should be grounded at only one point to prevent ground loops in the shields. This point may have to be found experimentally to give the lowest noise pickup. The equipment racks and transmitter should be connected together by copper straps at least 2 inches wide, tied to a good earth ground at one point. If a good ground screen is not available, a satisfactory ground can be provided by driving four or five copper ground rods 8 to 10 ft. long into the ground with a spacing of about 3 ft. These ground rods should be tied together with a wide copper strap. The straps connecting the equipment to the earth ground should be as short and direct as practical. It is often difficult to remove VHF-RF from the equipment by grounding because at FM carrier frequencies, nearly any connection to an earth ground has an appreciable impedance. The best way to keep RF out of sensitive low level circuits is by keeping them enclosed within an RF shield and by filtering leads that enter the shielded unit when necessary. Filters in the audio lines may be made up of small bi-filar RF chokes and disc capacitors.

For stereo transmission, it is necessary to keep the L and R audio lines phased properly. To insure proper monaural compatibility, correct audio phasing polarity must be maintained throughout the station from the microphones, tape machines, and turntables through all of the audio equipment to the stereo generator audio input terminals. Stereo phone line pairs or separate RF studio to transmitter links (STL) should also be checked for proper polarity and equal phase delay.

The equipment should be located and arranged to provide sufficient room around the front, sides, and rear for easy access during servicing and maintenance. Servicing of certain components may be easier by removing a side panel from the transmitter.

Transmitter Cooling

Almost all FM broadcast transmitters require forced air cooling to remove heat from the output stage and other assemblies within the cabinet. A very important consideration in locating the transmitter is the provision for adequate cooling air. As a rough approximation, it can be assumed that the overall efficiency of the transmitter is about 50 percent. In other words, it will generate about the same number of kilowatts of heat as it does RF power output.

Fig. 39A shows a transmitter located in an air conditioned room. This type of closed-loop system requires no special ducting and has the advantage that the transmitter intake air is usually much cleaner than outside air. The transmitter exhaust air places a substantial heat load on the air conditioner during the summer, but it becomes a source of heat in the winter. The transmitter manufacturer can usually supply data on the number of cooling BTU's required, so that the proper size air conditioner can be selected. This method is used frequently with the lower power transmitters. A protective system should be provided to prevent over-heating of the transmitter if the air conditioner fails.

Fig. 39B shows a transmitter located in a wall separating an air conditioned room and a ventilated, but not air conditioned room. A large exhaust fan is provided in the ceiling to remove the rising hot air while an adequate cool air intake is provided in the lower portion of an outside wall. Adequate air filtering is required to keep the transmitter interior clean.

Fig. 39C shows a transmitter located in an air conditioned room with intake and exhaust air ducts to the outside. An auxiliary blower or fan is normally required to overcome pressure drop in the ducting. This type of system requires careful design to make sure that the air flow through the transmitter is not impeded by the duct work. Additional air interlocks may be required to protect the transmitter from a failure of the external fan. The air intake and exhaust openings to the outside should be provided with rain shields, insect screens, and dust filters as dictated by the environment. The location of the air intake and exhaust openings should be arranged so that wind pressure will not impede the air flow.

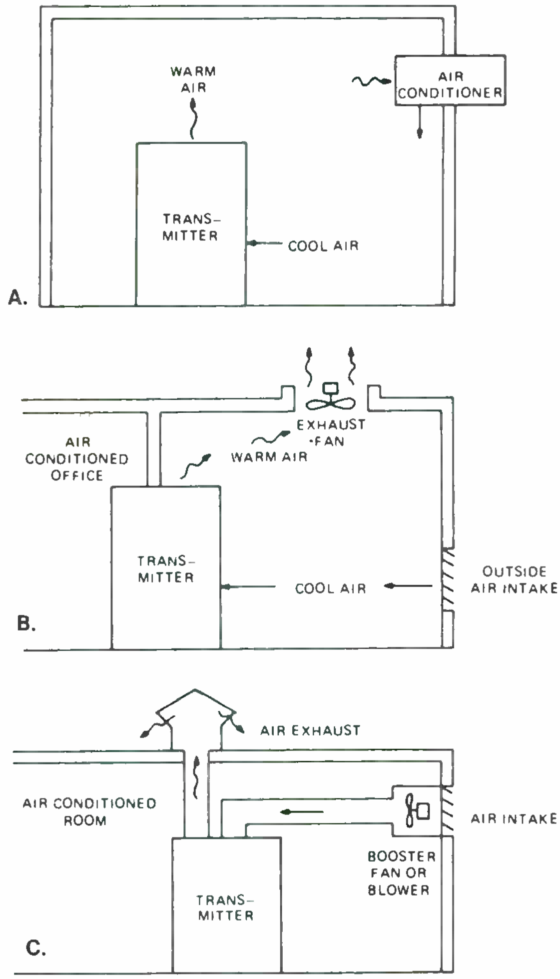


Fig. 39. Three methods of providing cooling air for the transmitter.

Air filters should be periodically cleaned or replaced according to the transmitter manufacturer's instructions. This is very important because dust or insect clogged air filters may reduce the cooling air flow enough to cause overheating of some of the components. The probability of component failure increases very rapidly when cooling is insufficient. Particular attention should be paid to removing dirt and dust from high voltage components during regular maintenance after all power is removed and all components are discharged.

Dust should be cleaned from the transmitter by means of a suitable brush and vacuum cleaner or as otherwise recommended by the transmitter manufacturer. Usually weekly cleaning is sufficient.

FM TRANSMITTER PLANT MAINTENANCE

Care of Power Tubes

The operating life of high power vacuum tubes can be extended by proper care. Most high power tubes utilize a directly heated cathode composed of a thoriated tungsten filament structure.

The key points to extending the life of RF power tubes are:

1. Store tubes upright (upside-down or right-side-up) along the axis of symmetry, not on their side. This will help to keep the internal elements concentrically aligned.
2. Use care when handling tubes to prevent mechanical shocks to the delicate internal structure. Don't set a tube on a hard surface without padding.
3. Keep the tube seals and anode cooler free of dust and dirt by cleaning once each week even in a clean environment.
4. Keep a spare tube on hand and rotate the tubes every few months to help keep the chemical "gas-getter" active so that the tubes remain gas free.
5. Keep a regular record of all tube operating parameters so that any trend of changes will be noticeable. In the event of a tube failure during the warranty period, this data will be essential to receiving credit on a replacement tube.
6. Monitor the filament voltage on a true RMS responding instrument and log any changes for future reference. The sampling point for this voltage measurement should be located as close to the tube's filament contacts as possible to minimize errors due to voltage drops in the filament wiring.

A tube that has been properly operated should gradually lose emission from the cathode until it is no longer useful because the emissive material is gradually consumed. The carcass of the tube can then be rebuilt with a new cathode and recycled back into service. Tube life is not directly related to plate dissipation (within the ratings), but is directly related to the current density (milliamperes per heater watt) emitted by a given size filament. This means that operating a given tube type at a lower plate current will proportionately increase the life of the tube.

Normally a new tube will deliver full output at a slightly reduced filament voltage. By operating the tube at the optimum filament voltage, the filament life can be significantly extended. The optimum value may be found by slowly reducing the filament voltage from the manufacturer's rated value until the RF power output drops about 2 percent and then increasing the voltage until the RF power increases back up 1 percent. In no case, should the filament voltage be reduced more than 5 percent below the manufacturer's rating. A brand new tube should be operated at the full rated filament voltage for the first 300 hours before the voltage is reduced to the optimum value for long life. This will assure that the "gas-getter" is properly activated. As the tube ages, the filament voltage will have to be increased to stay at the optimum value, until RF output power cannot be maintained at or above the rated value of filament voltage. At this point, the tube's useful life comes to an end. Check the manufacturer's data sheets and application notes for detailed information. An excellent guide to the proper care of power tubes is listed in reference (10) at the end of this chapter. Reference (17) gives detailed information about how to specify a proper forced-air cooling system for power tubes.

Preventive Maintenance

Preventive maintenance is equipment inspection and maintenance performed at regular intervals before an operational problem develops. The long term benefits are great because potential problems are discovered and solved while they are still easily manageable. A check-list of a few typical preventive maintenance items for an FM transmitter plant might include:

1. Weekly overall internal and external cleaning and inspection for damage or excessive wear.
2. Lubricate motors, tuning gears, and other moving parts at intervals recommended by the manufacturer.
3. Check and log all meter readings daily. Then compare these readings with the previous set of readings as an aid to diagnosing a developing problem.
4. Regularly exercise the automatic power control and any other servo systems.
5. Check the antenna lighting and de-icer systems.
6. Check the transmission line pressurization and VSWR.
7. Check all air filters in the transmitter plant and clean or replace as required.
8. Check the proper operation of all monitoring and remote control equipment.

9. Good overall housekeeping will pay big dividends in the long run by keeping the equipment clean and free of problems that would otherwise be caused by dirt build-up.

Maintenance Systems

The key to making any maintenance program work is to set up formal check-list, logging, parts inventory management, and repair scheduling systems. These systems provide the conscience and discipline required to keep the maintenance routine accurate and complete. Each station should develop a system suited for the particular physical plant involved. Often there will be more maintenance and repair work needing attention than there is time to do it all. In this case, the maintenance system should set the priorities for completing each item and assure that no item gets forgotten. Accurate notebooks describing all installation and maintenance work are a very helpful part of any maintenance system when working on the equipment years later after the human memory has faded.

REFERENCES

1. Lessing, Lawrence, *Man of High Fidelity: Edwin Howard Armstrong,* Bantam Publishing, 1959, The Armstrong Foundation, Columbia University, New York, N.Y.
2. *Reference Data for Radio Engineers,* ITT—Howard W. Sams and Co., Inc., New York, N.Y., Fifth Edition, 1970.
3. Terman, Frederick E., *Electronic and Radio Engineering,* McGraw-Hill Book Company, New York, N.Y., Fourth Edition, 1955.
4. Clarke, Kenneth K. and Hess, Donald T., *Communications Circuits: Analysis and Design,* Addison-Wesley Publishing Company, Reading, Massachusetts, Second Printing, 1978.
5. Rohde, Ulrich L., *Digital PLL Synthesizers-Theory and Design,* Prentice-Hall, Inc., Englewood Cliffs, New Jersey, 1983.
6. Krauss, Herbert L., Bostian, Charles W., and Raab, Frederick H., *Solid State Radio Engineering,* John Wiley and Sons, New York, N.Y., 1980.
7. Uyttendaele, Antoon G., "Design Requirements and Operational Features of the Gates Dualtran RF Switching System", Broadcast Products Division of Harris Corporation, Quincy, Illinois.
8. Johnson, Joseph, "Solid Circuits", Acrian/Communications Transistor Company, Cupertino, California, 1973.
9. Howe, Harlan Jr., "Simplified Design of High Power, N-Way, In-Phase Power Divider/Combiners", Microwave Journal, December, 1979.
10. Eimac, *Care and Feeding of Power Grid Tubes,* Eimac Division of Varian Corporation, San Carlos, California, 1967.
11. Hershberger, David and Weirather, Robert, "Amplitude Bandwidth, Phase Bandwidth, Incidental AM, and Saturation Characteristics of Power Tube Cavity Amplifiers for FM", Harris Corporation, Quincy, Illinois, 1982.
12. Motorola Silicon Rectifier Manual, Semiconductor Group-Motorola Inc., Phoenix, Arizona, 1980.
13. Hnatek, Eugene R., Design of Solid-State Power Supplies, Van Nostrand Reinhold Company, New York, N.Y., Second Edition, 1981.
14. Mendenhall, Geoffrey N., "A Study of RF Intermodulation Between FM Broadcast Transmitters Sharing Filterplexed or Co-located Antenna Systems", Broadcast Electronics Inc., Quincy, Illinois, 1983.
15. Mendenhall, Geoffrey N., "The Composite Signal—Key to Quality FM Broadcasting", Broadcast Electronics Inc., Quincy, Illinois, 1981.
16. Lyles, John T.M. and Shrestha, Mukunda B., "Transmitter Performance Requirements for Subcarrier Operation", Broadcast Electronics Inc., Quincy, Illinois, 1984.
17. Steinkamp, Jeffrey H., "Air Flow Measurements Using a Variable Frequency AC Generator", Broadcast Electronics Inc., Quincy, Illinois, 1984.
18. Robinson, John T. and Whipple, L. Glenn, "FCC Regulations Governing the Audio Fidelity Characteristics of FM Broadcast Equipment", Audio Engineering Society Journal, Vol. 16, No. 3, July 1968.
19. Bruene, Warren B., "An Inside Picture of Directional Wattmeters", American Radio Relay League — QST magazine, April 1959.

3.4

Subcarrier Transmissions and Stereophonic Broadcasting

John Kean
Senior Engineer
National Public Radio
Washington, DC

Stereophonic Sound FM broadcasting and SCA services are a form of *multiplexing* that had its origins fifty years ago, when high-fidelity audio and facsimile messages were simultaneously transmitted from the Empire State Building in New York City to experimental receive sites in New Jersey.[1] Those historic multiplexing efforts demonstrated the value of *wideband frequency modulation*, allowing modern FM broadcasting to provide services to the public of both quality and variety. More history of FM engineering and the origins of multiplex transmission are contained in the chapter "FM Transmitters". The purpose of this chapter is to explain the principles involved in multiplexing for FM broadcasting and to discuss the current state of the art.

THE COMPOSITE BASEBAND

This section contains definitions of the terms commonly used in the FM stereo and SCA systems. Since some of these terms are misused or ambiguous, the intention here is to establish meanings that will be used in the rest of this chapter.[2, 3] The list was selected to clarify certain terms and is not a complete glossary.

NOTE: Superscript numbers refer to footnotes and References at end of the chapter.

Glossary of Terms For Stereo and SCA

Multiplexing. In its simplest sense, *multiplexing* implies that two or more independent sources of information are combined for carriage over a single medium, namely, the radio frequency "carrier", and then are separated at the receiving end. In stereophonic broadcasting, for example, program information consisting of left and right audio signals are *multiplexed* onto an FM carrier for transmission to receivers which subsequently recover the original audio signals.

Channel. A transmission path. The distinction between the concept of a *channel* and a *signal* are not always clear. The usage herein distinguishes between transmission channels; e.g., main channel, stereophonic subchannel, etc., and left and right audio "signals".

Composite Baseband Signal. A signal which is the sum of all signals which frequency modulate the main carrier. The signal can be represented by a formula which includes all signal components: the main channel signal, the modulated stereophonic subcarrier, the pilot subcarrier and the SCA subcarrier(s).

FM Baseband. The frequency band from 0 hertz (Hz) to a specified upper frequency which contains the composite baseband signal.

Main Channel. The band of frequencies from 50 (or less) Hz to 15,000 Hz on the FM baseband which contains the main channel signal.

Main Channel Signal. A specified combination of the monophonic or left and right audio signals which frequency modulates the main carrier.

Stereophonic Sound. The audio information carried by plurality of channels arranged to afford the listener as sense of the spatial distribution of sound sources. Stereophonic sound includes, but is not limited to, biphonic (two channel), triphonic (three channel) and quadrophonic (four channel) services.

Stereophonic Sound Subcarrier. A subcarrier within the FM broadcast baseband used for transmitting signals for stereophonic sound reception of the main broadcast program service.

Stereophonic Sound Subchannel. The band of frequencies from 23 kHz to 99 kHz containing sound subcarriers and their associated sidebands.

Subchannel. A transmission path specified by a subchannel signal occupying a specified band of frequencies.

Subchannel Signal. Subcarrier(s) and associated sideband(s) which frequency modulate the main carrier. It is synonymous with *subcarrier*, as in the stereophonic subcarrier or SCA subcarrier.

Frequency Deviation. The peak difference between the instantaneous frequency of the modulated wave and the carrier frequency.

Percentage Modulation. The ratio of the actual frequency swing of the carrier to the frequency swing defined as 100 percent modulation, expressed in percentage. Although current FCC Rules conditionally permit greater than 100 percent modulation when SCAs are transmitted, a frequency swing of ±75 kHz is still defined as 100 percent modulation.

Injection. The ratio of the frequency swing of the FM carrier by a subchannel signal to the frequency swing defined as 100 percent modulation, expressed in percentage. The total injection of more than one subchannel signal is the arithmetic sum of each subchannel injection.

Crosstalk. An undesired signal occuring in one channel caused by an electrical signal in another channel.

Linear Crosstalk. A form of *crosstalk* in which the undesired signal(s) is created by phase or gain inequalities in another channel or channels. Such crosstalk may be due to causes external to the stereophonic generator; consequently it is sometimes referred to as *system crosstalk*.

Nonlinear Crosstalk. A form of *crosstalk* in which the undesired signal(s) is created by harmonic distortion or intermodulation of electrical signal(s) in another channel or channels. Such crosstalk may be due to distortion within the stereophonic generator or FM transmitter; consequently it is sometimes referred to as *transmitter crosstalk*.

Frequency Spectrum and Modulation Limits

The stereophonic transmission standards in Section 73.322 of the FCC Rules, and the SCA transmission standards in Section 73.319, were revised in 1984. The original definition of the "stereophonic subcarrier" was removed from the Rules and the definition of "stereophonic sound" was enlarged to cover any number of audio signals needed to convey the spatial distribution of sound sources.

Stereophonic sound broadcasting now includes, but is not limited to, the familiar biphonic (two channel) service described herein.

The composite baseband has been extended from 75 kHz (the old frequency limit) to 99 kHz and may be used in support of either stereophonic or SCA multiplex services. Within the frequency range of 23 kHz to 99 kHz any form of amplitude modulation (DSB, SSB, etc.), angle modulation (FM or PM) or frequency shift keying of a multiplex subchannel is permitted.

Under certain conditions when SCA multiplex subcarriers are operated, limits on the total maximum percentage modulation have changed. It is no longer required to "back off" the main channel and stereophonic subchannel levels by an amount equal to the SCA subcarrier injection in order to maintain 100 percent modulation; a total modulation of up to 110 percent is legal in some cases.

The term SCA is an acronym for a "Subsidiary Communications Authorization", once required in order for a station to begin broadcasting a multiplex service. Although the term "SCA" continues, the authorization is no longer required. Except for certain "common carrier" service, broadcast licensees may begin transmitting multiplex services without prior notification of or authorization from the FCC.

The figures below show the baseband spectrum in various modes, from monophonic to stereophonic plus fully-loaded SCA operation.

Fig. 1a represents the basic monophonic program mode, where the baseband width is limited to approximately 15 kHz and no other signals are multiplexed. In this case all the modulating energy is contributed by the main channel. Not more than 100 percent modulation is permitted in this case.

Fig. 1b shows the baseband with SCA operation in addition to monophonic main channel service. Total SCA injection up to 30 percent is permitted within the band from 20 kHz to 99 kHz. This injection figure may be contributed by one

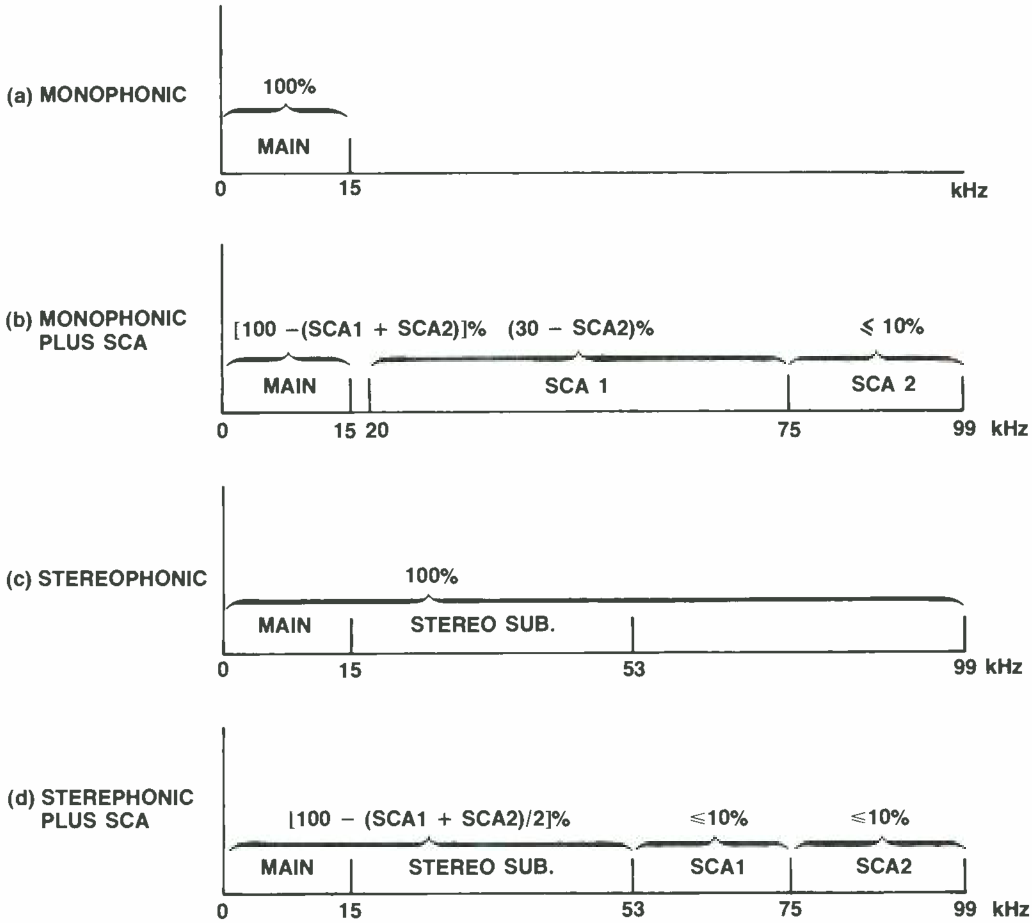


Fig. 1. Frequency spectrum and modulation limits for monophonic, stereophonic and SCA operation. Modulation percentages are referred to 75 kHz carrier deviation.

or more SCA subcarriers, provided that the arithmetic sum of all subcarrier injection above 75 kHz does not exceed 10 percent. Total modulation is not to exceed 100 percent. This is required to insure that the bandwidth of the main carrier (and its interference to other stations on adjacent and alternate channels) is not significantly increased.

When transmitting stereophonic programming, the main channel modulation must be limited to 100 percent minus one-half the total SCA injection. Total modulation, however, is not to exceed 110 percent. Modulation limits were established for the same reasons of protection from increased interference mentioned above.[4]

Note that the FCC Rules permit transmission of multiplex subcarriers when *no* broadcast program service is carried on the main channel, provided that the above modulation rules are met.

Fig. 1c shows the basic stereophonic sound program mode, without SCA operation. As is the case for monophonic program operation, the modulation must be limited to 100 percent. Frequencies up to 99 kHz are available for stereophonic sound program transmission. Actual bandwidths and characteristics of stereophonic signals are discussed in the next section of this chapter.

Fig. 1d adds a single band for SCA operation to the stereophonic mode. While total SCA injection may be up to 20 percent, no more than 10 percent total injection may be employed within the frequency bands from 53 kHz to 75 kHz and 75 kHz to 99 kHz. (There is one exception to determining total SCA injection involving subcarriers that are multiples of and synchronous to the stereophonic pilot. This will be discussed in the SCA multiplex section of the chapter.) The modulation contributed by the main channel and stereophonic subchannel signals must be no more than 100 percent minus one-half the total SCA injection. Since the total injection may be up to 20 percent, total modulation may be up to 110 percent.

STEREOPHONIC TRANSMISSION STANDARDS—PART 73.322

The stereophonic subchannel is the key to multichannel sound broadcasting. Its performance is strictly defined by the FCC Rules:

(a) The modulating signal for the main channel shall consist of the sum of the left and right signals.

(b) A pilot subcarrier at 19,000 Hz plus or minus 2 Hz shall be transmitted that shall frequency modulate the main carrier between the limits of 8 and 10 percent.

(c) The stereophonic subcarrier shall be the second harmonic of the pilot subcarrier and shall cross the time axis with a positive slope simultaneously with each crossing of the time axis by the pilot subcarrier.

(d) Amplitude modulation of the stereophonic subcarrier shall be used.

(e) The stereophonic subcarrier must be suppressed to a level of less than one percent modulation of the main carrier.

(f) Stereophonic sound subchannels must be capable of accepting audio frequencies from 50 Hz to 15,000 Hz.

(g) The modulating signal for the stereophonic subcarrier shall be equal to the difference of the left and right signals.

(h) The preemphasis characteristics of the stereophonic subchannel shall be identical with those of the main channel with respect to phase and amplitude at all frequencies.

(i) The following modulation levels apply to stereophonic sound transmission:

1. When a signal exists in only one channel of a two channel (biphonic) sound transmission, modulation of the carrier by audio components within the baseband range of 50 Hz to 15,000 Hz may not exceed 45 percent and modulation of the carrier by the sum of the amplitude modulated subcarrier in the baseband range of 23 kHz to 53 kHz may not exceed 45 percent.

2. When a signal exists in only one channel of a stereophonic sound transmission having more than one stereophonic subcarrier in the baseband, the modulation of the carrier by audio components which the audio baseband of 50 Hz to 15,000 Hz may not exceed 37 percent and modulation of the carrier by the sum of all subchannel components with the baseband range of 23 kHz to 99 kHz may not exceed 53 percent.

(k) At the instant when only a positive left signal is applied, the main channel modulation shall cause an upward deviation of the main carrier frequency; and the stereophonic subcarrier and its sidebands shall cross the time axis simultaneously and in the same direction.

(l) The ratio of peak main channel deviation to peak stereophonic subchannel deviation when only a steady state left (or right) signal exists shall be within plus or minus 3.5 percent of unity for all levels of this signal and all frequencies from 50 Hz to 15,000 Hz.

(m) The phase difference between the zero points of the main channel signal and the stereophonic subcarrier sidebands envelope, when only a steady state left (or right) signal exists, shall not exceed plus or minus 3 degrees for audio modulating frequencies from 50 Hz to 15,000 Hz.

(n) The separation between any two channels of a stereophonic transmission system must exceed 29.7 dB for all audio modulating frequencies between 50 Hz and 15,000 Hz; this separation will indicate compliance with the amplitude and phase tolerances given in paragraphs (l) and (m) above.

(o) Nonlinear crosstalk into the main program channel caused by signals in the stereophonic subchannel must be attenuated at least 40 dB (measured at RMS noise) below 90 percent modulation; nonlinear crosstalk into the stereophonic subchannel(s) caused by signals in the main channel must be attenuated at least 40 dB (measured as RMS noise) below 90 percent modulation; (nonlinear crosstalk does not include effects of phase delay differences in program audio circuits...these effects are represented by a loss of channel separation and by amplitude distortion in the monophonic reception of stereophonic programs).

GENERATING THE STEREOPHONIC BASEBAND SIGNAL

Fig. 2 shows the composite baseband that modulates the FM carrier for biphonic broadcasting. (SCA multiplex subchannels are not part of this band and will be discussed later.)

The two-channel stereo baseband has a bandwidth of 53 kHz, and consists of:

- A main channel *(L + R)* consisting of the sum of left plus right audio signals; this is the same signal broadcast by a monaural FM station. However, it must not modulate a full 75 kHz since it would lead to overmodulation and interference when the stereophonic signals are added. Instead, it is modulated 45 percent for a full left-only or right-only signal.

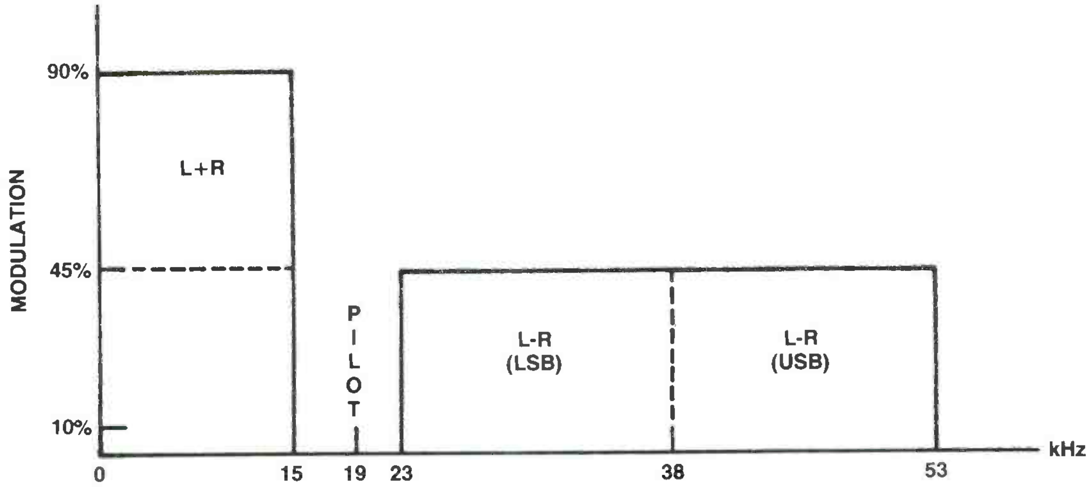


Fig. 2. Bi-phonic (two-channel) stereo baseband.

- A stereophonic sound subchannel *(L − R)* is required, consisting of a double sideband AM modulated subcarrier with a 38 kHz center frequency. The modulating signal is equal to the instantaneous difference of the left and right audio signals. The subcarrier is suppressed to avoid wasting modulation capability. As a result the AM sidebands have the same modulation potential as the main channel.
- A 19 kHz subcarrier *pilot* which must be exactly one-half the frequency of the stereophonic subcarrier and very nearly in phase to it. It supplies the reference signal needed by the stereo receivers to reinsert a 38 kHz carrier for DSBSC demodulation (or, more commonly, to synchronize the synchronous decoder circuitry in a receiver). The frequency tolerance is plus or minus 2 hertz and must modulate the main carrier between 8 and 10 percent.

In general two principles have been used to generate the stereophonic subchannel. The methods are called *time division multiplex* (TDM) or *switching method* and *frequency division multiplexing* (FDM) or *matrix method.*

Frequency Division Multiplexing

A popular method for generating the stereophonic baseband involves the direct generation of the double sideband suppressed L − R subchannel along with the L + R channel.

A simplified block diagram of the FDM system is shown in Fig. 3. Both left and right audio channels are preemphasized and low pass filtered just as a normal monaural signal would be. In the matrix, the left and right audio signals are both added and subtracted. The audio signals are added to form the L + R main channel which is also used as the monaural broadcast signal.

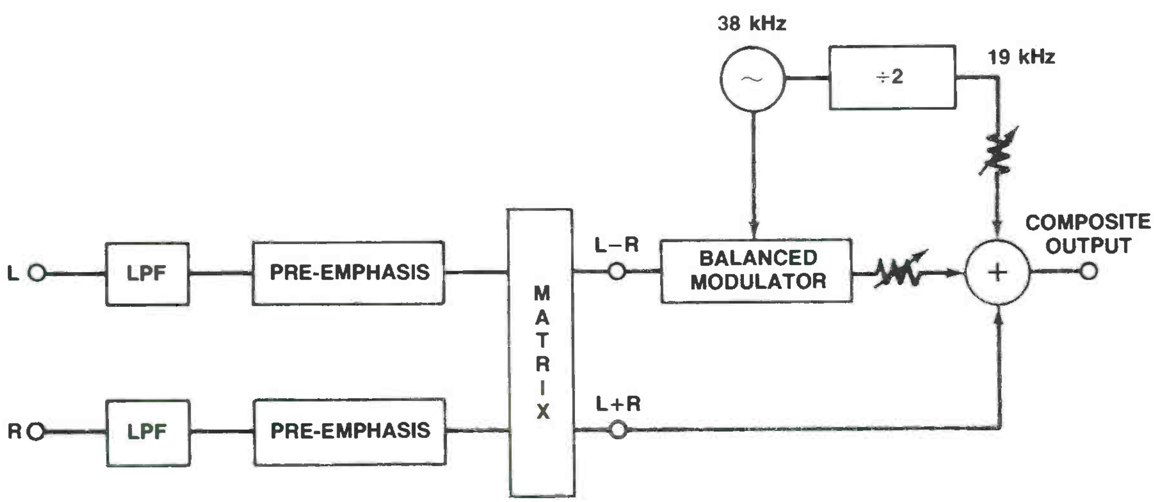


Fig. 3. Functional blocks of a frequency division multiplex stereo generator.

The subtracted signals are fed to a balanced modulator which generates the L – R subchannel. Since a balanced modulator is used, the carrier at 38 kHz will be suppressed, leaving only the modulated sidebands.

The 38 kHz oscillator is divided by 2 to make the 19 kHz pilot tone. Finally, the main channel stereophonic subchannel and pilot are combined in the proper (45, 45, 10) ratio to form the composite output.

An examination of the composite stereo waveform in the time domain (as displayed by an oscilloscope) will illustrate some basic situations. First consider a 1 kHz sine wave applied equally to the L and R audio inputs. This is shown in Fig. 4a without a pilot signal. The only frequency present is 1 kHz, since the matrix produces no difference signal necessary to generate sidebands in the stereophonic subchannel.

Fig. 4b illustrates the ideal composite signal when the same 1 kHz tone is applied to the L and R inputs but exactly out of phase. The pilot is still off. Two frequency components are generated: 37 kHz and 39 kHz. No L + R signal appears out of the matrix; thus, only the sidebands of the modulated 38 kHz subchannel are present.

The symmetrical envelope shown represents a double-sided band suppressed carrier (DSBSC) AM signal. Note that the amplitude of each sideband is one-half that of the L + R component in Fig. 4a. When demodulated in the receiver's stereo decoder, the sidebands are added together to produce an output equal to the full L + R signal.

Finally, consider the waveform in Fig. 4c, when the composite signal (still without pilot) is generated by applying a 1 kHz tone to the L in-

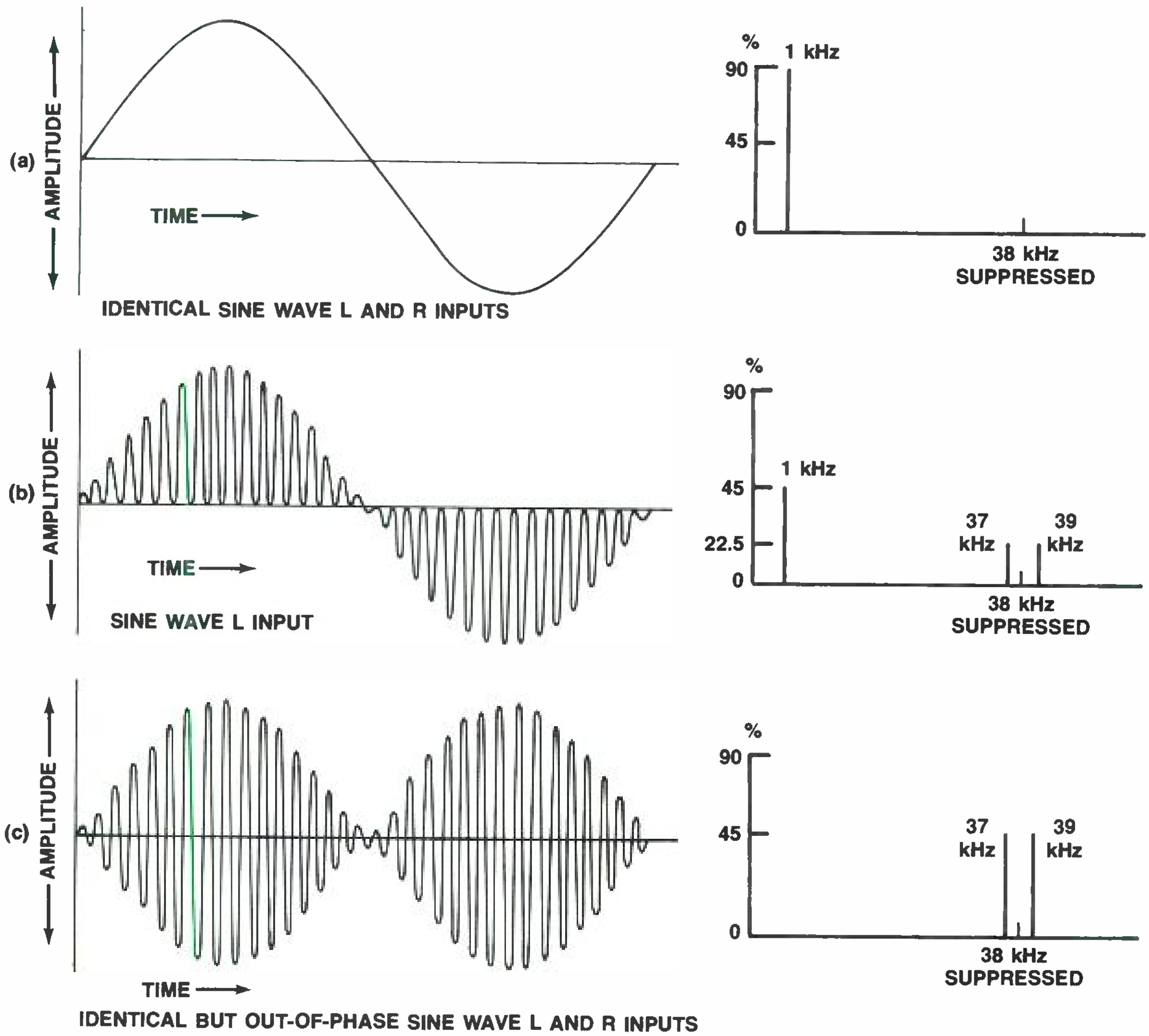


Fig. 4. Time domain (oscilloscope display) and frequency domain diagrams of stereo baseband signal with (a) identical sine wave L and R inputs, (b) sine wave L input, and (c) identical but out-of-phase sine wave L and R inputs.

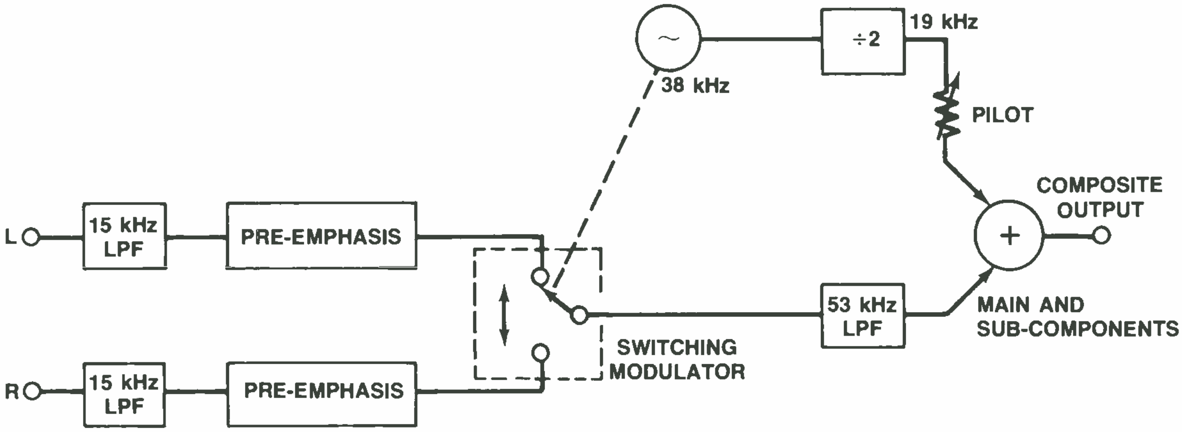


Fig. 5. Functional blocks of a time division multiplex stereo generator.

put alone. The baseline of the waveform envelope will be a straight if there is no amplitude or phase subchannel. Three frequency components are present: 1 kHz, 37 kHz and 39 kHz. These sidebands are each one-half the voltage amplitude of the 1 kHz signal in the main channel; together they equal the energy of the main channel in this instance.

The last diagram looks the same when an R-only 1 kHz tone is applied, although the phase of the two sidebands would be reversed with respect to the 38 kHz subcarrier (and the pilot). Adding the pilot, at its relatively low amplitude of 10 percent, produces the same overall waveforms, but its "spreading" effect makes display of the basic waveform less clear. That is true in practice, as well, which is a reason most stereo generators have provision for testing without the pilot.

Time Division Multiplex

A different type of stereo generator is in use which produces the same result as frequency division multiplexing by using a switching technique.

Generation of both the L + R and L − R channels is accomplished by an electronic switch which is toggled by a 38 kHz signal. The switch alternately samples one audio channel and then the other, as shown in Fig. 5. The original signal can be reconstructed from periodic samples, provided that the samples are taken at a rate at least twice the frequency of the highest audio frequency component (approximately 15 kHz in broadcast FM).

The output waveform for the TDM generator is next shown first in the time domain (as an oscilloscope would display the signal) for a sequence of input signals. The diagrams at the right

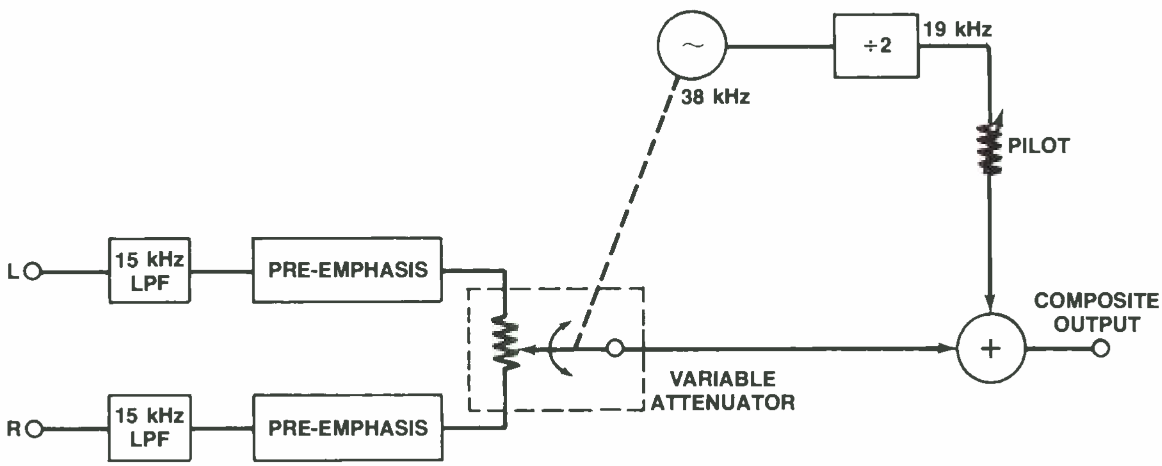


Fig. 6. Functional blocks of a time division multiplex stereo generator using a variable attenuator.

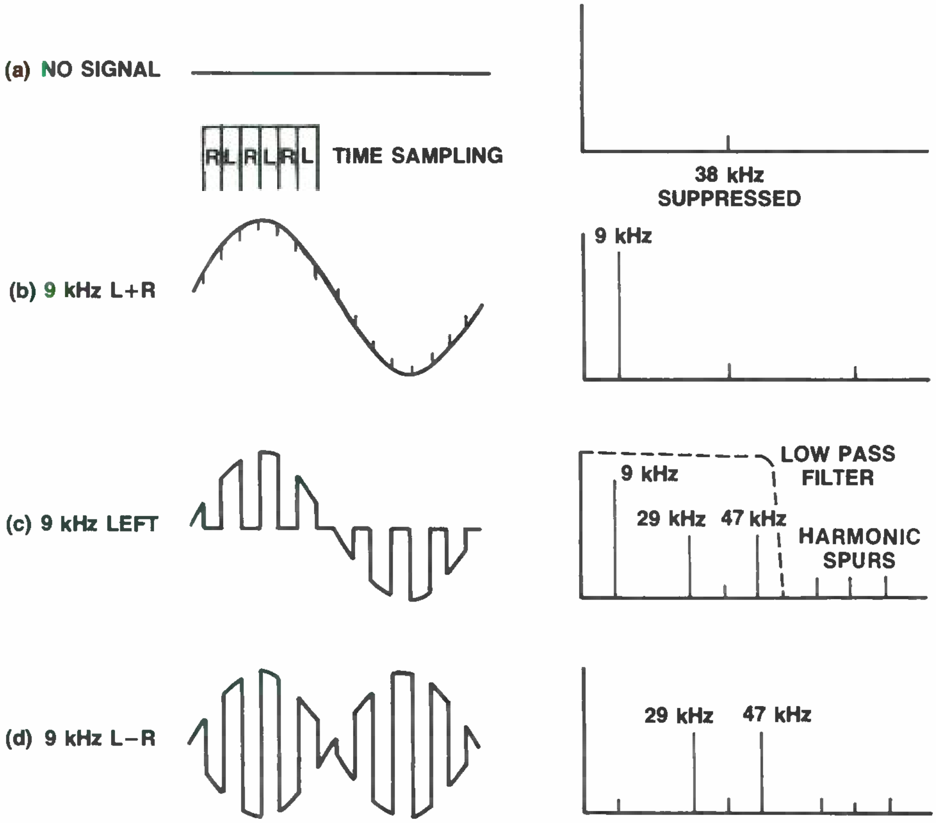


Fig. 7. Time domain and frequency domain diagrams of stereo baseband signals.

of the waveform show the same signal in the frequency domain (as would be displayed on a spectrum analyzer).

In Fig. 7a, there are no input signals present. Ideally, no output signals are possible, and in practice only a small amount of leakage of the switching transients are present. Since the transfer time of the switching signal is extremely quick, harmonics of the fundamental 38 kHz are possible.

A 9 kHz audio tone is applied to the L and R inputs in Fig. 7b. The 9 kHz input signals are combined at full amplitude (90 percent modulation) and no subchannel sidebands are generated.

In Fig. 7c, only the left channel has a signal present. As the switch selects the L audio line, samples are passed along to the composite output. Therefore, the output waveform shows the same signal, chopped into segments of 1/38,000th of a second. Since the total area under the waveform has been divided in half, it should be apparent that the energy of the 9 kHz signal in the L + R channel is only half the amplitude that it would be if an equal 9 kHz signal were also present at the right channel. The equation for the output signal e for an input signal σ at any instant t is

$$e = 1/2 \sin \sigma t$$
(main ch. audio)
$$+ 1/\pi \, [\sin(\phi + \sigma)t + \sin(\phi - \sigma)t)]$$
(DSBSC)
$$- 1/3\pi \, [\sin(3\phi + \sigma)t + \sin(3\phi - \sigma)t]$$
(3rd Harmonic)

. . . etc. *(higher harmonics)*

Fig. 7c shows the original 9 kHz signal (at half amplitude), and a pair of sidebands centered about the 38 kHz switching frequency. No 38 kHz signal is generated if the switching waveform has perfect symmetry, that is, if the switch is connected to the left and right channels for precisely equal periods. Note that a harmonic of the stereophonic subcarrier is shown, centered around 114 kHz which is three times the switching frequency. Only one extra term was shown in the equation, however, other terms at the 5th, 7th, etc. are present.

In addition to the odd-order harmonics of the 38 kHz subchannel, asymmetry in the switching signal or other circuit imbalances can create some sidebands centered about the second harmonic at 76 kHz. All these harmonics must be removed by filtering, as shown in the diagram.

When the odd harmonics are filtered out, the proper DSBSC waveform results. However, it is slightly greater in amplitude than the L + R signal. (Because the fundamental component of the square wave is $4/3\pi$ times larger than the square wave amplitude, the DSB component is larger than the L + R component.) This is easily corrected by adding enough of the L and R audio to the output to equalize the amplitude.

In Fig. 7d, the TDM signal is shown when the L and R signals are equal in amplitude and exactly reversed in phase. This waveform matches the composite stereo signal shown in Fig. 4c.

The composite lowpass filter should have very steep cutoff characteristics but should have flat amplitude response and linear phase shift with frequency (equal time delay at all frequencies) below 53 kHz. While this approach to stereophonic generation is simple and stable, the filter can degrade stereo separation, especially at higher audio frequencies.

A significant improvement on the original switching concept is shown in Fig. 6. As mentioned earlier, the higher order terms of the square wave-driven switch are responsible for generating the harmonics of the 38 kHz subchannel which must be removed by filtering. By using a "soft switch" to connect back and forth between the L and R channels it is possible to eliminate the lowpass filter and its side-effects. This is accomplished by using a variable attenuator, shown in the diagram by a potentiometer. The slider is driven from end to end of the potentiometer by a sinewave. Since a sinewave is represented only by a single, fundamental frequency, the signal output at the slider has the proper DSBSC characteristics without the harmonics. The equation for the composite signal generated in this way is

$$\begin{aligned} e = {} & 1/2 \sin \sigma t && \text{(L + R audio)} \\ & + 1/\pi\,[\sin(\phi + \sigma)t + \sin(\phi - \sigma)t)] && \text{(38 kHz DSBSC)} \end{aligned}$$

As the equation shows, only the fundamental sidebands of 38 kHz are present in the sampled signal, along with the main channel component. Like the fast-switching TDM system, the L + R and L − R channels are generated in one operation so that the circuit remains relatively simple. No filter of the output is required, provided that the 38 kHz sinewave is free from harmonics and the variable attenuator has good linearity.

Stereo Decoder Circuits

Stereo FM receivers include a circuit to convert the multiplexed composite signal at the FM detector into the original Left and Right audio channels transmitted by the FM station. There are at least as many ways to decode the stereophonic signal as there are ways to encode (generate) the composite signal. In practice, only one type of decoder is commonly used, the so-called "phase-locked loop" integrated circuit.

The circuit in Fig. 8a is seldom used, but is shown for comparison. It is the closest complement to stereo generators using frequency division multiplexing. At the input, the composite signal is split with three filters into the main (L + R) channel, pilot signal and stereo (L − R) subchannel. Next, the pilot is doubled to 38 kHz and this regenerated carrier is reinserted into the double-sideband AM signal from the subchannel filter. This AM signal is demodulated to yield the L − R (difference) audio. Finally, the L + R and L − R signals are combined in a sum and difference matrix to produce to original Left and Right audio channels.

Because of the costly filters needed to separate the composite spectrum, the frequency division multiplex circuit is not used in consumer equipment. Similar circuits have been used in broadcast modulation monitors, where metering of the separate channels is required.

The circuit shown in Fig. 8b is universally used, due to its simplicity, high performance and low cost. While commonly referred to as a "phase locked loop" (PLL) stereo decoder, its distinction is really as a time division demultiplexer (shown in the dashed box as a toggle switch).

Following a buffer amplifier, the composite baseband signal is sampled by a phase locked loop within the IC. A voltage controlled oscillator, usually running at 76 kHz (four times the pilot frequency) is held in phase with the pilot by a reference signal from the phase comparator and loop filter. When divided by two, the result is a square wave at 38 kHz having nearly perfect duty cycle (high and low states have equal timing) and very fast rise and fall times. This signal is ideal for driving the output audio switcher (demultiplexer). This stage is a transistor matrix designed to rapidly transfer the composite baseband to the Left and Right audio output in time with the switch in the station's stereo generator. Fast, clean audio switches are relatively easy to make, and because they are only "on" or "off", they are not subject to drift.

Because PLL stereo decoders normally use square wave switching, the circuit is able to demultiplex (demodulate) baseband signals which are odd harmonics of 38 kHz. The third harmonic (114 kHz) is most troublesome, since noise and spurious signals near this frequency are shifted to the audio baseband, as is the 38 kHz stereophonic subchannel.

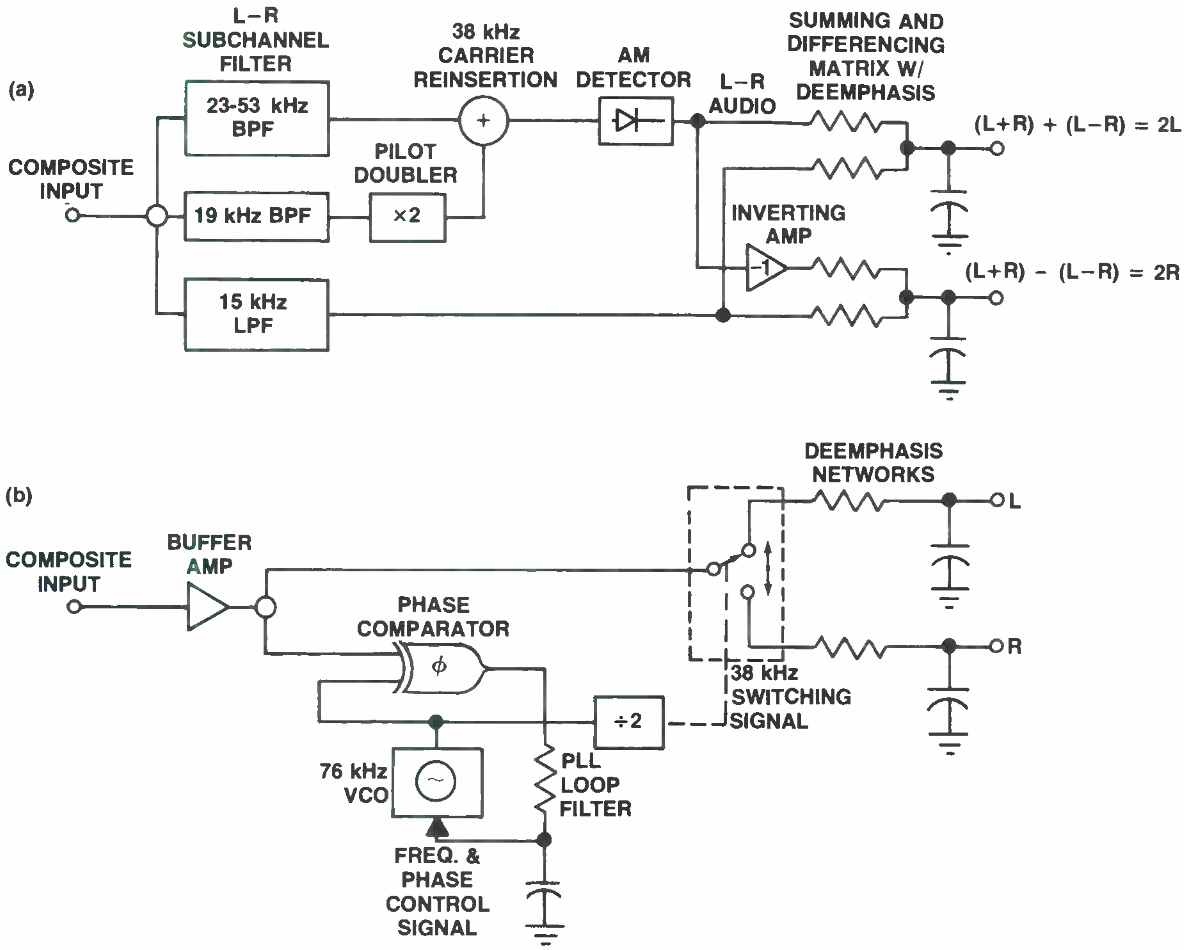


Fig. 8. Functional blocks of stereo decoders using (a) L+R and L−R matrixing and (b) phase-locked time-division multiplexing.

Engineers should be watchful of the frequency band centered on 114 kHz in their transmitted signal since audible noise may occur in consumer receivers. Some recent stereo decoder ICs utilize a second composite audio toggle switch operated at 114 kHz. The demodulated product is inverted and mixed equally with the 38 kHz switching outputs, cancelling the response to signals in the 114 kHz range.

FM SCA TRANSMISSION

Background

From its beginning as a broadcast service, people have recognized the potential of FM for multiplexed services. As early as 1940, the FCC permitted multiplex facsimile transmission on FM stations, but not until much later did auxiliary FM services attain a wide acceptance among FM broadcasters.

In 1955 the FCC established the Subsidiary Communications Authorization and created an entirely new industry. The original intent of the authorization was to permit programming of background music to offices, stores, restaurants, etc. where it was uneconomical to provide this service via telephone lines. For many commercial FM stations, the SCA operation became a major source of revenue which enabled them to survive in the 1950s.

Because educational broadcasters had not indicated a real desire for additional programming services, they were specifically excluded from providing any SCA services. However, in 1961 the FCC authorized SCA operation in a manner consistent with their noncommercial status.

By the early 1980s, improvements in transmitter and receiver technology and the desire for new revenue prompted commercial and noncommercial broadcasters to seek changes in the SCA rules. In a series of rulemakings in 1982 and 1983 the Commission made numerous changes to expand technical opportunities and reduce legal regulation:[5]

- elimination of the requirement for an FCC authorization (Subsidiary Communications Authorization) prior to operating multiplexed services;
- extension of the baseband frequency limit from 75 kHz to 99 kHz;
- allowance of any type of subcarrier modulation to be used for multiplexed services;
- changes in the subcarrier injection requirements to permit multiple services;
- increased limits for the total modulation during SCA multiplex operation to reduce main channel modulation loss;
- operation of SCA services on noncommercial FM stations under the same rules as commercial FM stations.

FCC Rule Requirements for SCA Operation

The FCC rules have no standards for SCA subcarrier performance. This is left for the broadcaster or lessee of the service to determine. However, the Commission is quite specific in defining the transmission conditions under which subcarriers may be operated, in order to minimize interference to the main channel and stereophonic subchannel or to other FM stations.

The constraints of injection and subcarrier bandwidth will be considered in detail for the stereophonic mode, since stereo transmission is by far the most common FM service. Fig. 9 shows the upper portion of the composite baseband, from approximately 50 kHz to 100 kHz. This is the same as the upper portion of Fig. 2 with the addition of two hypothetical subcarriers. (Any number of SCA subcarriers may be operated within this frequency range provided that each subcarrier has adequate bandwidth and injection. This will be discussed later in the chapter.)

On the baseline are several frequency markers at the upper and lower frequency limits of the two subcarrier ranges. The actual frequencies of the subcarriers are not specified, but are referred to as "fc_1" and "fc_2".

Alongside each subcarrier are arrows marking the level in dB below 100 percent modulation of the main carrier. For example, -20 dB marks the injection at the center frequency of both subcarriers. Since 10 percent is the maximum injection permitted under the Commission Rules within each SCA subchannel,

$$\text{injection} = 20 \times \log_{10}(0.1) = 20 \times (-1) = -20 \text{ dB}$$

At 53 kHz, an arrow marks a level of -60 dB. The FCC requires that any frequency modulation of the main carrier due to the SCA operation shall be at least 60 dB below 100 percent modulation in the frequency range of 50 Hz to 53 kHz when stereo is transmitted. This figure must include spurious and intermodulation products as well as subcarrier sideband energy.

At 99 kHz, the level of -20 dB is marked, denoting the FCC requirement that "instantaneous sidebands" be restricted within this frequency limit. The Commission has not officially defined instantaneous sidebands, but it is normally considered to be the instantaneous frequency of the subcarrier at its peak deviation (for frequency modulated subcarriers) or the highest sideband frequency (for amplitude modulated subcarriers).

Interference Between FM SCA Subcarriers

In practice, two subcarriers should be separated as far apart in frequency as possible, while observing the limit of spurious energy below 53 kHz and the instantaneous sidebands at 99 kHz. While the FCC Rules are silent on the choice of

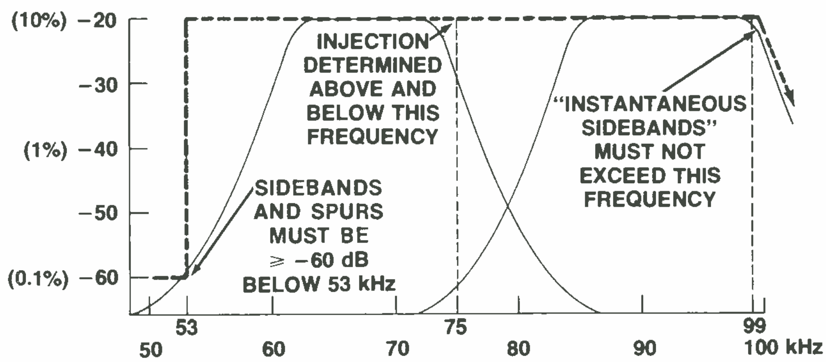


Fig. 9. Injection, channel bandwidths and spurs limits for SCA operation when stereo is transmitted. Two possible subcarriers (at 67 and 92 kHz) are shown.

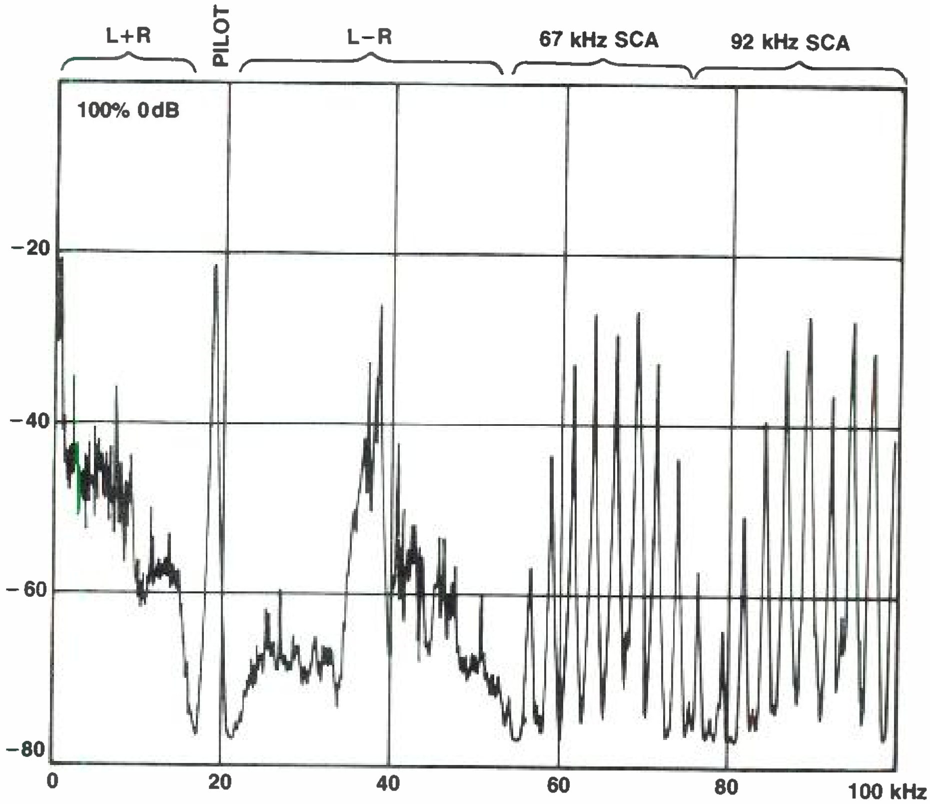


Fig. 10. Composite baseband of FM station with 67 kHz and 92 kHz FM-SCAs, modulated with 2.5 kHz tone at 5 kHz and 7 kHz peak deviation, respectively. Note overlap of sidebands between subcarriers. Station is carrying stereo programming.

SCA frequencies, the industry has two defacto standards for frequency modulated subcarriers: 67 kHz and 92 kHz. The first frequency was adopted when the original stereophonic standards placed an upper limit of 75 kHz on the baseband. The second frequency was recommended to situate the instantaneous sidebands below 99 kHz while remaining safely distant from a 67 kHz subcarrier.[6]

Some overlap of the subcarrier sidebands does occur, as depicted in Fig. 10. This does not cause significant interference between the two SCA subchannels when the systems use frequency modulation:

- unweighted 67 kHz SCA S/N ratio referred to 5 kHz deviation: 57.5 dB
- same, with 4 kHz tone modulation of 92 kHz subcarrier at 7 kHz dev.: 56.5 dB
- same as first case w/ tone mod. of main channel at 67.5 deviation: 52.5 dB

(Receiver used for data: McMartin TRE-6 with regulated power supply substituted for standard equipment)

The above data shows that crosstalk from main-to-subchannel is significantly higher than crosstalk from one highly modulated subcarrier in the adjacent subchannel.

Other Types of Subcarrier Modulation

Other frequencies and total number of subchannels are permissable, according to the occupied bandwidths and interference margins required. An example of an SCA baseband spectrum combining five amplitude-modulated subcarriers is shown in Fig. 11.

Here, a variety of bandwidths and injection levels are used. At 57 kHz, a very narrow double-sideband AM subchannel is shown. This system was developed in Europe for "Automotive Road Information" (ARI), to identify stations which broadcast traffic information and to activate specially-equipped FM car radios during traffic announcements. (The system is now being introduced on selected stations in this country by Blaupunkt.) An interesting effect of adding the 57 kHz subcarrier in phase to the stereophonic pilot is that there is virtually no increase in peak modulation of the main carrier. (A third-order

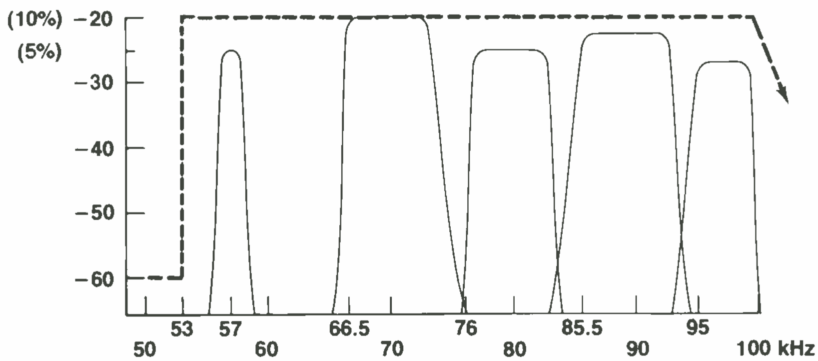


Fig. 11. Carrier envelope spectrum, DSB-AM and SSB-AM SCA subcarriers. Vertical markings are dB referred to 75 kHz deviation.

harmonic does not increase the peak amplitude when in-phase and at an amplitude one-half the fundamental.) This allows 57 kHz double-sideband AM subcarrier to be added to the baseband without being counted toward the injection from 53 kHz to 75 kHz and without requiring the back-off of main channel modulation. However, this subchannel must have very narrow bandwidth to operate with an acceptable margin of reliability, and thus it has limited information capacity.

Note in Fig. 11 that upper-sideband, suppressed carrier AM (SSBSCAM) systems are used for the four other subchannels. On the first (at 66.5 kHz), nearly 10 percent peak envelope injection is being used, while the other three use a combination which adds up to ten percent to satisfy the limit for the 75 to 99 kHz range.

Examples of SCA Operation

Fig. 12a shows the functional blocks of a standard aural SCA generator using frequency modulation. At the input, audio is lowpass filtered to limit the generation of high-order sidebands which could cause excessive bandwidth and possibly interfere with the stereophonic subchannel or an upper SCA subchannel.

A 5 kHz audio cutoff is frequently used so that fidelity compares favorably to unequalized telephone service. Using a Bessell function analysis, one can determine the peak deviation which may be used to maintain FM sidebands below 53 kHz within FCC limits:

SCA Modulating Frequency	SCA Peak Deviation
3.5 kHz	5.0 kHz
5.0	3.5
7.5	2.0

For the above protection requirements, equipment manufacturers often recommend 3.5 or 4.0 kHz peak deviation with a 5 kHz lowpass filter when stereo is transmitted. In practice, slightly higher deviation may be used since the calculations assume single-tone modulation which has greater modulation power than program audio.

Preemphasis follows the lowpass filter in the sample FM SCA generator. A value of 150 Sec is commonly used to combat the rising high-frequency noise characteristic in the FM SCA channel. Unfortunately, this extreme boost (almost 17 dB at 5 kHz, compared to 400 Hz) requires substantial amounts of peak limiting to allow full modulation at low and middle audio frequencies.

The processed audio signal is fed to the FM modulator, usually with a provision for subaudible (less than 30 Hz) channel for telemetering return. The telemetry channel takes advantage of the extended low frequency response capability of FM subcarriers by sending tone signals in the 10 to 30 Hz range at a level of 15 to 25 dB below reference peak deviation. A highpass filter is inserted ahead of the audio input when subaudible telemetry is used to avoid interference to the low frequency channel.[7]

Ideally, no filtering of the frequency-modulated subchannel is required. However, some of the FM generator circuits produce spurious or harmonic energy. For this reason, a bandpass or lowpass filter is used which passes all the significant sidebands but suppresses spurious signals to acceptable levels. The modulated subchannel signal is amplified to a level required by the subcarrier input of the FM exciter or studio transmitter link equipment.

A simplified block diagram of a SSBSCAM subchannel generator is shown in Fig. 12b. The input audio is bandpass filtered to control subchannel bandwidth (to 5 kHz, for example), and may be compressed according to specific amplitude and spectral standards. A complementary expander circuit in the receiver restores the

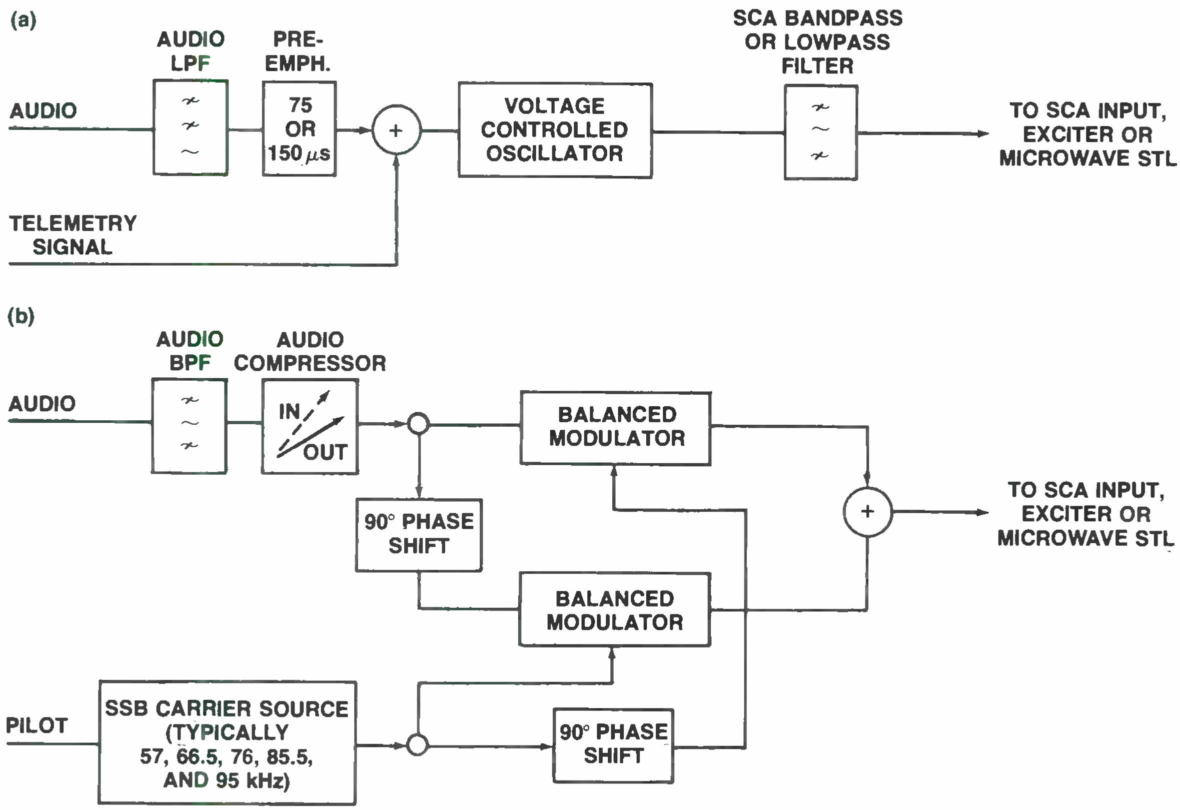


Fig. 12. Functional blocks of (a) a frequency-modulated SCA subcarrier generator, and (b) a single-sideband AM subcarrier generator.

original program dynamics and reduces noise in the channel. Large amounts of preemphasis are not as important since the spectral characteristic of the noise is flat with a linear modulation system, not rising with frequency as in angular modulation systems.

The processed audio signal is fed to a pair of balanced modulators, one path going through an allpass filter which shifts the audio phase 90 degrees throughout the required audio range. The balanced modulators are driven by a subcarrier frequency source which may be frequency-locked to a harmonic of the stereophonic pilot (and its half-frequency of 9.5 kHz, as shown). One of the modulators is driven by a carrier signal which is shifted 90 degrees. When the two modulated carriers are summed, cancellation of the lower (or upper) sidebands creates the desired single-sideband signal.

Since SSBSCAM requires a carrier for detection, the pilot provides an excellent source for reinsertion at the receiver. With monophonic audio, no pilot is transmitted, so another carrier reference must be used. One approach is to carry the 19 kHz reference at a reduced level which will not trigger the stereophonic decoders in FM receivers.

SCA DATA SYSTEMS

SCA subchannels have characteristics which are quite favorable for data transmission: while the channels are noisy for high fidelity audio services, they can provide good to excellent data error performance, and moderately fast data rates are possible without complicated or expensive receivers. Combined with the high power and elevation of FM broadcast stations, SCA data systems may be operated over large service areas.

Indirect Data Modulation

There are many methods of putting data on an SCA subchannel, but most can be separated into two broad groups: direct and indirect carriage. Indirect methods have been used on SCAs for years and are the most similar to data circuits over phone lines. Here, an audio oscillator is switched between two frequencies, e.g., 1070 Hz and 1270 Hz, representing the "0" and "1" states of binary data. This method is commonly called "Audio Frequency Shift Keying" (AFSK).

Since the product of AFSK signalling remains within the audio band, a standard SCA generator and receiver may be used, just like a background music service. There is little cost or complexity involved in starting this data service on SCAs since there is a great deal of AFSK-type equipment in use, and the data modulator and demodulator functions have been reduced to single integrated circuits. This technique is, however, relatively slow: the highest practical speed is 1,200 bits/second. This may be acceptable where the hardware must be "off the shelf" and there is no premium on SCA channel efficiency. A single-sideband AM subcarrier (discussed earlier) would, of course, increase channel efficiency.

Direct Data Modulation

Direct data modulation eliminates the intermediate steps that convert the data into audio signals, allowing the subcarrier to be directly modulated by the binary data. There are several types of direct data modulation which are discussed below.

In Frequency Shift Keying (FSK), the two binary states are represented by two different subcarrier frequencies. For example, a "0" could be 65 kHz and a "1" 69 kHz. FSK is the most popular form of direct data modulation at this writing, probably because of its robust performance, simplicity and low cost. Maximum speed of 4,800 to 9,600 bits/second are typical for current FSK systems using a nominal injection of 10 percent and a modulated bandwidth similar to that of background music services, i.e., about 16 kHz at a point about 25 dB below subcarrier injection level.[8]

All data FSK systems us filtering methods to control occupied bandwidth, since too-rapid frequency transitions could generate high-order sidebands that could interfere with adjacent subchannels and violate FCC Rules.[9] Filtering is accomplished in several ways, shown in Fig. 13:

- Where a pair of oscillators are alternately toggled by the data states, filtering must be done to the output spectrum to directly reduce occupied bandwidth.
- Where analog FM modulators are used, slew rate filtering of the switching signal tends to reduce high-order sidebands indirectly.
- In some cases, occupied bandwidth is reduced through the use of both pre- and post-modulation filtering.

The spectrum of the composite baseband of an actual FM station is shown in Fig. 14, carrying a 4,800 bits/second data signal. The spectrum was generated by a standard SCA generator driven by a filtered data signal. All reduction of the bandwidth was accomplished *prior* to modulation.

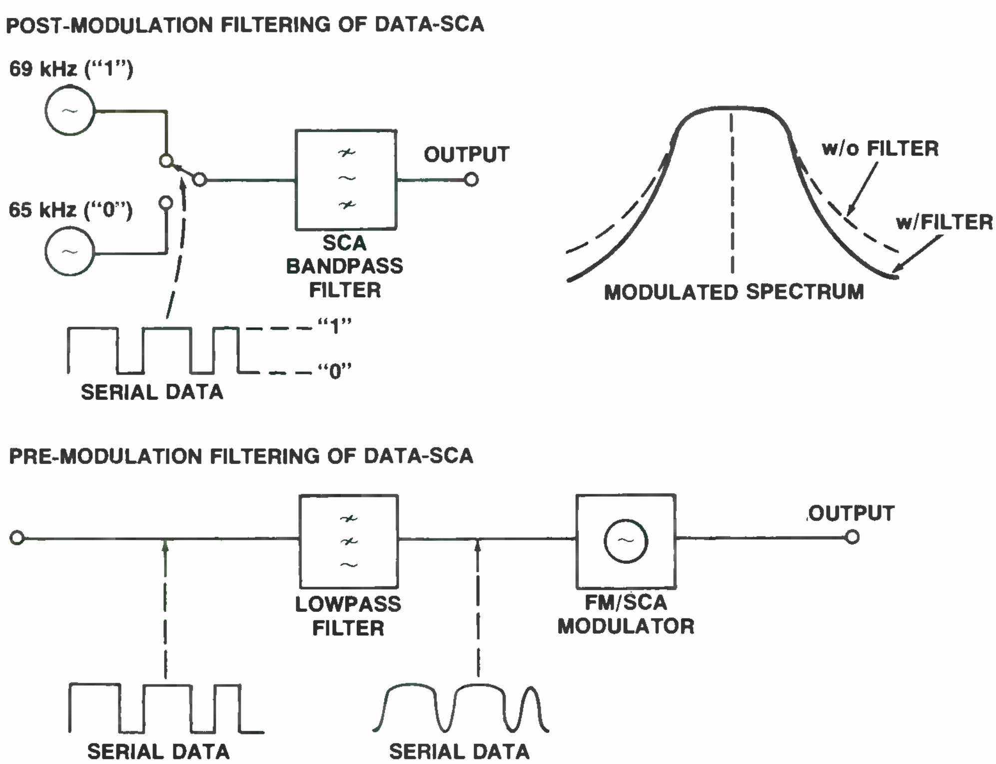


Fig. 13. Two basic methods for generating and controlling the bandwidth of data subcarriers used for SCA.

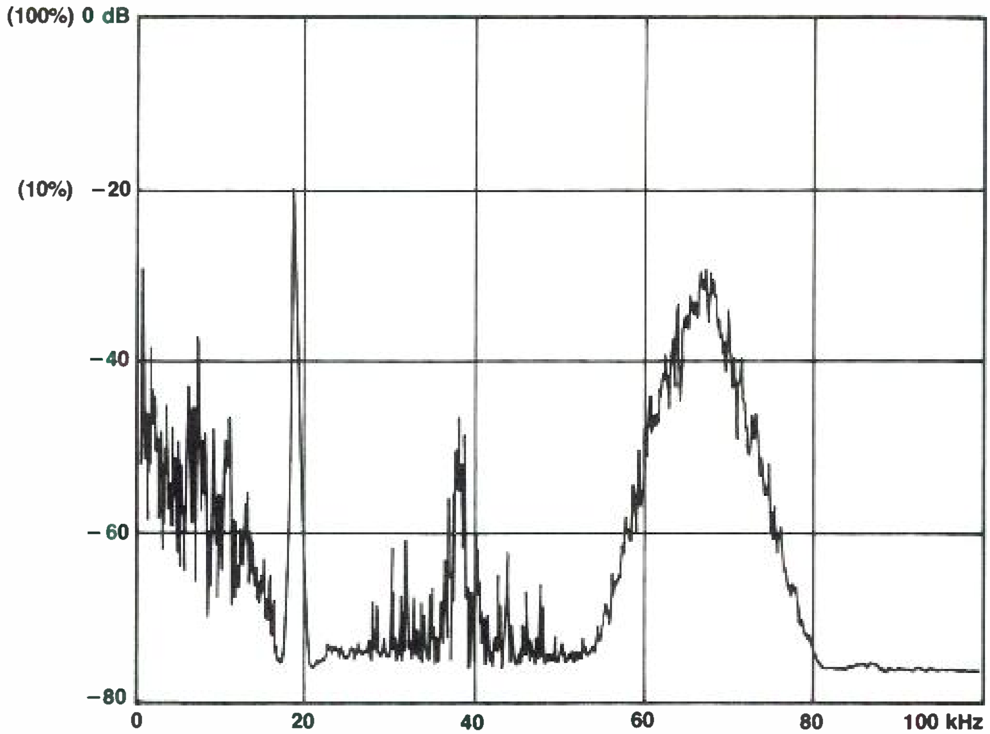


Fig. 14. Composite baseband of an FM station carrying a 4800 baud frequency-shift-keyed data subcarrier. Center frequency of the subcarrier is 67 kHz, with an injection of 9%. Station is carrying stereo programming.

Phase Shift Keying (PSK) is another technique for transmitting high-speed data. With PSK, one phase of the carrier represents one binary state and a second phase (usually 180 degrees apart) is used for the second state. To decode such modulation, a phase reference is required. Deriving a stable reference from the average phase of the incoming signal is quite difficult. But, since an external phase reference is permissible, the stereophonic pilot, if available, is well suited for this purpose. Pilot-tone-locked PSK systems hold considerable potential for high speed data transmission on SCA, with 56,000 bits/second practical. However, most of the 53 to 99 kHz bandwidth might be required and receiver/decoders would be much more expensive than those for moderate speed FSK systems.

Data Multiplexing

One of the problems for broadcasters is not the *speed* of data SCA subchannels, but data multiplexing. Multiplexing carries the same definition for data as it does in the analog domain: several channels of information share the same carrier. SCA subchannels operate independently of the main channel program because of Frequency Division Multiplexing (FDM). However, the use of this spectrum is not 100 percent efficient because *guard bands* are necessary between the channels: analog filters have less-than-ideal cutoff characteristics. In the data domain, on the other hand, a stream of bits may be divided with near-ideal precision by time division multiplexing (TDM) or statistical multiplexing (which can apportion time and data dynamically according to each channel's momentary need).

At this writing, no publicly-accessible data protocol has been adopted by the SCA broadcast industry. It is, therefore, difficult for broadcasters to establish services which can share independent users, even though the data speeds of the equipment available to them exceed the needs of most single users. This will tend to limit data services to users who can pay for an entire subchannel, even if their need does not require the channel's full capacity.

MONOPHONIC, STEREOPHONIC & SCA PERFORMANCE

While monophonic, stereophonic and various SCA services are conveyed by the same FM carrier, performance varies considerably due to the relative carrier deviation of each and the particular frequency span occupied within the composite baseband. The performance of a channel may be expressed in terms of demodulated signal-to-noise ratio for a given radio frequency carrier level, or, conversely, coverage range for a given signal to noise ratio. The purpose of this section is to summarize and illustrate these relations but not present the lengthy derivations for the equations.

The performance of a conventional FM receiver in the presence of random (thermal) noise is commonly judged on the basis of the variation of the output signal to noise power ratio as a function of the carrier to noise power ratio contained within the receiver bandwidth, usually determined by the IF bandwidth. An estimate of the IF bandwidth required for transmission of (monophonic) broadcast FM is given by[10]

$$\beta_{IF} = 2(F_{dev} + 2f_m) = 210 \text{ kHz} \qquad [1]$$

Where:

F_{dev} = the peak frequency deviation (75 kHz), and

f_m = the highest baseband modulating frequency (15 kHz).

This β_{IF} figure is rounded to 200 kHz for later use.

The equivalent input noise power in a bandwidth β launched into a noise-free (cool) receiver is

$$rB = 4 \times 10^{-21}\, \beta f_N = 8 \times 10^{-16}\, f_N \text{ Watt} \qquad [2]$$

Where:

f_N = the receiver noise figure, and

β = nominal noise bandwidth in Hz.

The carrier power is

$$C = (V\mu v^2/R) \times 10^{-2} \text{ Watt} \qquad [3]$$

Where:

$V\mu v$ = the rms input voltage in microvolts

R = the receiver r.f. input resistance.

For R = 75 ohms, the carrier to noise ratio in decibels is

$$CNR = 10\log_{10}(C/rB) = 12.2 + 20\log_{10}V\mu v - 10\log_{10}f_N \qquad [4]$$

For example, for a 10 dB noise figure, the carrier to noise ratio is 0 dB for an input voltage of 0.77 V over 75 ohms (1.55 μv over 300 ohms).

For a half-wave dipole antenna matched to a 75 ohm receiver, the input voltage is related to the field strength as

$$V\mu v = 48.5\, E\mu v/f_{MHz} \qquad [5]$$

Where:

$E\mu v$ = the field strength in μV

f_{MHz} = the carrier frequency in MHz.[11]

Thus, for 98 MHz (middle of the FM band)

$$V\mu v = E\mu v/2 \qquad [6]$$

With the FCC F(50,50) field strength curves, it is possible to relate $E\mu v$ to the distance to the transmitter, effective radiated power, and antenna heights. Ultimately, it is possible to predict the carrier to noise ratio for a given broadcast system. For a 200 kHz nominal bandwidth the CNR in dB can be shown to be

$$CNR = Eo + P + G - F + 20\log_{10}(h/30 \times 98/f) + 6 \qquad [7]$$

Where:

Eo = field strength for 1 kW ERP @ 1000′ HAAT, in dB$\mu V/m$

P = ERP in dBkW

G = receiver antenna gain relative to dipole, in dB

h = receiving antenna height above ground, in feet

F = receiver noise figure in dB

f = frequency in MHz.

For example, if $F=10$, $P=10$, $G=0$, $h=30$, and $f=98$, then the $CNR = Eo + 6$.

Next, the signal to noise ratio is established by the carrier to noise ratio

$$SNR = 40 + 10\log_{10}CNR \text{ dB} \qquad [8]$$

This expression assumes a peak deviation of 75 kHz, a de-emphasis corner frequency of 2.1 kHz (75Sec) and a nominal intermediate frequency noise bandwidth of 200 kHz.[12]

From the above, a carrier to noise ratio of 0 dB would result in a signal to noise ratio of 40 dB. However, standard FM demodulators typically exhibit a "threshold" at about 12 dB CNR, below which the noise floor changes from a "hiss" to a raspy kind of noise with noticeable "clicks" jutting above the noise. For this reason, the monophonic channel actually deteriorates below a signal-to-noise ratio of about 52 dB.

A well-known effect of FM demodulation is that the noise voltage in a narrow band (for example, 1 Hz) increases directly with frequency. This has important implications for systems which employ subchannels since both their frequency span and the specific bottom and top frequencies are necessary in predicting relative signal-to-noise ratios. The stereophonic (L – R) subchannel, as shown in Fig. 2, extends nominally from

23 kHz to 53 kHz and occupies twice the bandwidth of the main (L + R) channel. When the subchannel noise is combined with the main channel noise during stereo decoding, the resulting noise floor is 23.1 dB higher than a monophonic system.[13]

The signal to noise ratio of an SCA subcarrier, relative to the main channel, is a function of a number of variables and is

$$Ns/No = (2fc^2/m^2D^2)\,\frac{f_1^3(b_1 - \arctan\, b_1)}{f_0^3(b_0 - \arctan\, b_0)} \qquad [9]$$

Where:

F_{dev} = the peak frequency deviation (75)
fc = subcarrier frequency,
m = injection level as a percent of 75 kHz
D = deviation of FM subcarrier
f_0 = −3 dB freq. of de-emph. in main audio ch. (2.12 kHz)
f_1 = −3 dB freq. of de-emph. in SCA audio ch. (1.06 kHz)
b_0 = main channel audio output bandwidth (15 kHz)/f_0
b_1 = SCA channel audio output bandwidth (5 kHz)/f_1

An example of the excess noise in the 5 kHz SCA audio channel, compared to the 15 kHz main channel is given below, if $m = 0.1$ (10%). Note that the noise level of a 67 kHz subcarrier with 5 kHz deviation and a 92 kHz subcarrier with 7 kHz deviation are nearly equal. Comparative tests by the author have indicated that susceptibility to impulse noise, e.g., gasoline engine ignition radiation, is greater at the higher subcarrier frequency even though gaussian noise levels are approximately the same. This may be due to a higher noise threshold in the 92 kHz demodulator (due to a wider predetection bandwidth than the 67 kHz demodulator) or to differences in the spectral distribution of impulse noise compared to gaussian noise in the IF (composite) demodulator.[14]

fc (SCA freq., kHz)	67	92	92
D (deviation, kHz)	5	5	7
Ns/No (above main, dB)	36.4	39.2	36.2

Fig. 15 graphs the relative subcarrier noise level against increasing frequency of a subcarrier, compared to the main (monophonic) channel for the same receive conditions. Shown are two peak deviations (5 kHz and 7 kHz) for FM SCA, and SSBSCAM SCA with 10 percent peak envelope injection. The graph segment representing 7 kHz deviation is dashed below 70 kHz since excessive interference to the L − R subchannel is possible in this frequency range when stereo is transmitted.

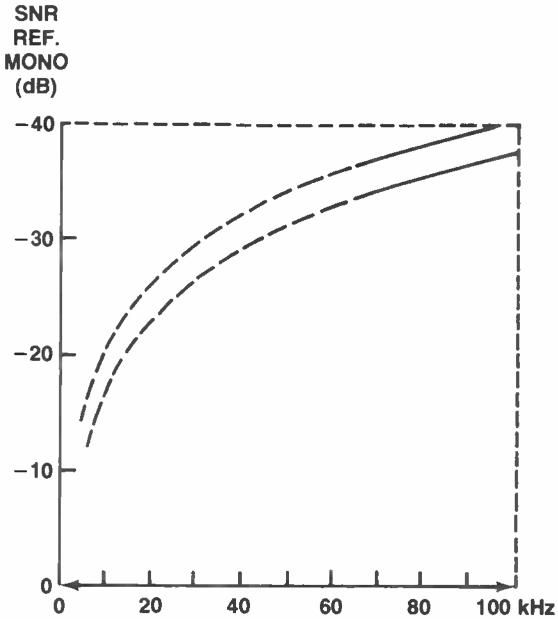


Fig. 15. SNR of FM/SCA audio (5 kHz bandwidth) versus subcarrier frequency, compared to monophonic reception, referred to 5 kHz and 7 kHz deviation.

Single Sideband AM Subcarriers

The conditions under which standard SCA subcarriers are operated are quite close to those existing with narrow band frequency modulation. For example, when a maximum audio bandwidth f_m of 5 kHz and peak deviation f_d of 5 kHz are used, the modulation index is one.

$$i = f_d/f_m = 5/5 = 1$$

After frequency demodulation, the signal-to-noise power ratio S/N will be[15]

$$S/N = \frac{3 \cdot i^2}{C/N} = \frac{3 \cdot 1}{C/N}$$

where C/N is the carrier to noise power ratio in the bandwidth occupied by the modulated signal. The noise power bandwidth β_i is difficult to predict for two reasons. First, the (usually) inexpensive SCA pre-detection filter has sloping skirt selectivity, thus departing from the ideal rectangular shape assumed in the calculation. Second, the actual bandwidth may vary, depending on the amount of audio distortion to be tolerated. The extremes of choice may be

for $\beta_i \leqslant 1$, $B_n = 2f_m = 10$ kHz
for $\beta_i \geqslant 1$, $B_n = 2(f_d + f_m) = 20$ kHz

A noise power bandwidth of 15 kHz is a compromise value that happens to represent the measured performance in some common FM SCA receivers.

In single sideband systems, one is simply frequency shifting the modulation passband to the baseband, so that the recovered signal-to-noise ratio is

$$S/N = C/N$$

where the noise power bandwidth is equivalent to the audio bandwidth (assumed to be 5 kHz).

From the above, it can be seen that while the FM system improves S/N, there is also three times the noise power. Thus, the 3 in the numerator is cancelled by a C/N in the denominator which is three times larger. At a modulation index of one, the FM improvement is approximately 10log (1) or 0 dB for an ideal detector. The use of preemphasis and de-emphasis improves the FM system S/N without significantly changing the subcarrier's occupied bandwidth. The choice of 150 microseconds adds an estimated 8.8 dB improvement in rms signal-to-noise ratio, but the author has measured typical improvements of approximately 4 dB in production samples.[16]

Since single sideband modulation is spectrally efficient, even when compared to narrow band FM, it is practical to divide the 53-99 kHz range into a number of independent subchannels. In the one system, five subchannels each having a 5 kHz baseband are operated at 57, 66.5, 76, 85.5 and 95 kHz. These are shown in Fig. 11.

Twenty percent peak injection is the maximum permitted by the FCC Rules when stereo is transmitted, therefore, each subchannel would receive four percent injection (20/5 = 4). It is reasonable to assume that slightly higher injection could be permitted for each subchannel due to the random interleaving of the total subchannel injection. If five percent were employed, then, the S/N would be lower than that for a channel using ten percent peak injection.

$$\text{change in } S/N = 20\log(10/5) = -6 \text{ dB}$$

Assuming an approximated improvement of 4 dB for a standard FM subchannel, and a reduction of 6 dB for one of five single sideband subchannels, there is an estimated difference in S/N, assuming the same center frequency of both subchannels and flat noise density over this spectrum.

$$S/N_{FM} - S/N_{SSD} = 4 - (-6) \cong 10 \text{ dB}$$

The power density from AM demodulators remains uniform across the baseband (unlike FM demodulators, in which noise density rises in proportion to frequency), therefore, preemphasis/deemphasis is not important. The use of a complementary noise reduction system (compandor) can significantly improve the signal-to-noise ratio of either system.

Using the above estimates, the performance of the (uncompandored) single sideband and (deemphasized) FM subchannels is compared to main channel and stereo in Fig. 16. The coverage radii are based on the FCC F(50,50) field strength predictions for an FM station having 10 kW ERP, 1000 feet height above average terrain, and a dipole receiving antenna at 30 feet above ground level into receiver having a 10 dB noise figure.

SCA-to-Stereo Interference Considerations

Introduction of new signals into the modulating baseband of an FM station requires attention to its possible interactions with existing baseband signals, as well as its chosen performance characteristics. Several recent findings regarding the potential for perceptible interference (crosstalk) are summarized here.

FM stereo receivers having stereo decoders with diode switching bridges driven by a high-level 38 kHz sinusoid (usually identified by discrete component construction and tuned transformer stages) were responsible for the original "birdie" or beat-note interference in early SCA/Stereo broadcast systems. For historical perspective, the sources of this interference were

- nonlinearity in the diode bridge due to the conduction characteristics of the diodes and the sinusoidal drive signal;
- imbalance in the center-tapped transformer driving the diode bridge and second harmonics of the 38 kHz sinusoid at this point.

The result caused a second-order intermodulation (mixing) of twice the 38 kHz switching frequency minus the instantaneous frequency of the SCA subcarrier as

$$9 \text{ kHz} = 76 \text{ kHz } [38 \text{ kHz} \times 2] - 67 \text{ kHz}$$

The product is not 10 kHz as was often believed. Today, the small amount of beat note interference that results from a 67 kHz subcarrier is 10 kHz, but due to an entirely different mechanism.

Over the past decade these receivers have been entirely replaced by sets with integrated circuit decoders, which have virtually no internally-generated beat-note problem. As a result, the interference to stereo service from SCA operation has dropped significantly.

Contemporary receivers create very little beat-note interference from SCA operation because of the type of stereo decoding done within the integrated circuit. The fact that the circuit contains

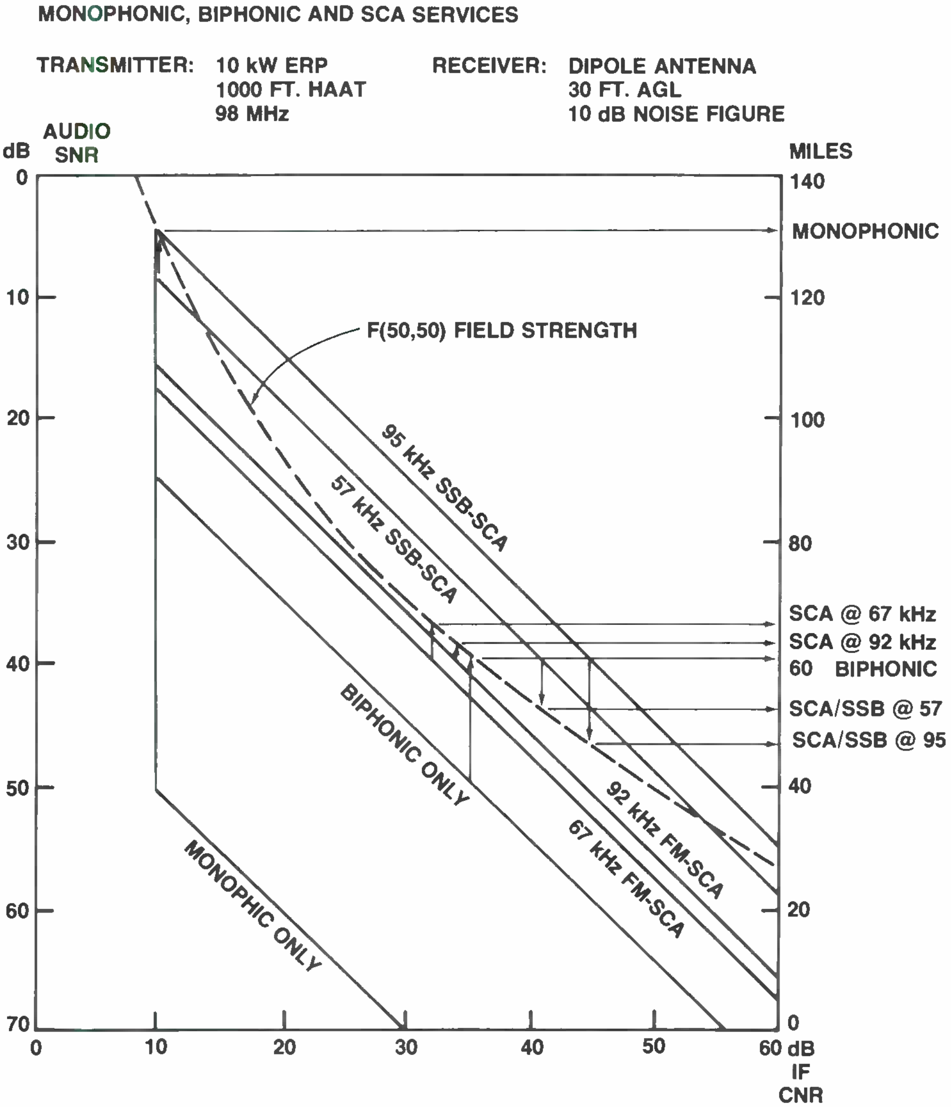


Fig. 16. Predicted service distances for 50 dB monophonic and biphonic SNR, and 40 dB SCA SNR. To use, find intersection of desired SNR with service (diagonal) line; next move up (or down) to F(50,50) dashed line; then read across to mileage scale.

a phase-locked loop (PLL) has little to do with this improvement. Modern decoders derive their 38 kHz signal by a digital technique, which is well-suited to circuit integration. This results in a 38 kHz square wave of fast rise time and balanced duty cycle (equal time in high and low states). The output circuit is usually a differential transistor pair configured as a balanced demodulator. When driven by the 38 kHz square wave (which has virtually no even-order harmonics), *very* little mixing is possible which can result in the product shown in the above equation.

If SCA beat-notes created within stereo receivers was the only source of interference, there would be virtually no problem with SCA operation. However, experience shows that a minor source of interference remains when certain SCA frequencies are used. The source of this interference is *external* to the receiver: multipath distortion of the radio frequency signal itself.

Large obstructions such as mountains, hills, and tall buildings can reflect VHF broadcast waves well enough to create the simultaneous reception of a direct and one or more reflected-path signals. Upon demodulation the baseband signals will include a combination of linear, second and third order distortion products, as shown in Fig. 16. Higher order distortions are negligible when the distortion is small.[17] Hence, these products take the form

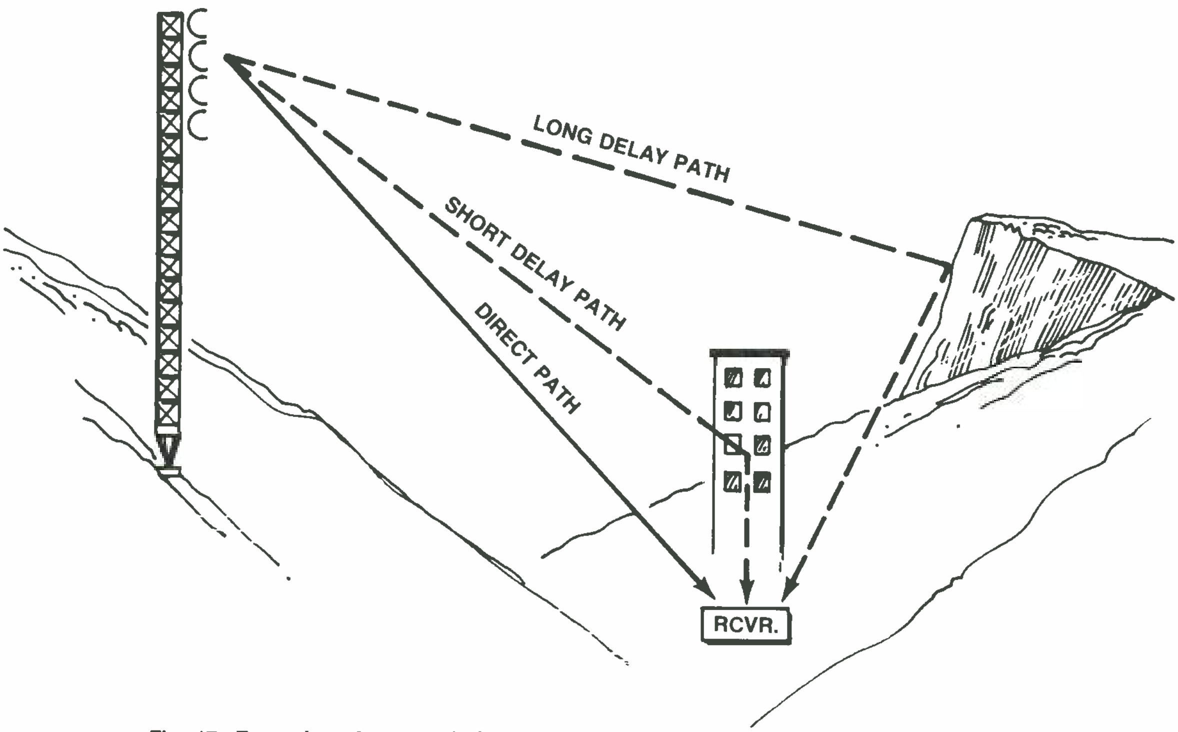


Fig. 17. Examples of reflected signals arriving at an FM receiver, along with a direct path RF signal. Distortion of the demodulated FM signal is related to both the direct/reflected signal ratio and delay time of the reflected signal; distortion increases as the signal ratio approaches 1:1 and as the secondary path delay time rises.

$2A$, $2B$, $2C$, etc. (where A, B, and C are the fundamentals)

$A+B$, $A-B$, $A+C$, $A-C$, $B+C$, $B-C$, ...*etc.*

$2A+B$, $2A-B$, $2B+A$, $2B-A$, ...*etc.*

Second order distortion (frequency sum and difference) is the most troublesome, because new product frequencies are created which may fall within the main and stereophonic channels where they can become audible. This is illustrated in Fig. 17, which shows the baseband of an actual FM station. The only modulation present in the transmitter is the stereo pilot and unmodulated SCA subcarriers at 67 kHz and 92 kHz, plus a small amount of noise from the audio chain which can be seen from zero to 15 kHz and centered around 38 kHz in the L – R subchannel.

Several second order products are visible in the spectrum graph. This distortion takes the form:

$f_{SCA1} \pm f_{SCA2} = f_{IM\ product}$, or

$f_{SCA} \pm \text{pilot} = f_{IM\ product}$

In the case of 67 and 92 kHz subcarriers, significant products are

$(J) \quad 92-67 = 25$ kHz
$(K) \quad 67-19 = 48$ kHz
$(L) \quad 92-19 = 73$ kHz
$(M) \quad 67+19 = 86$ kHz

While these IM products are not directly audible, some may be once they are demodulated by a stereo decoder:

$25-38 = 13$ kHz
$38-48 = 10$ kHz
$111-114 = 3$ kHz

(The last product could be demodulated by IC stereo decoders using a 38 kHz square wave switching signal. However, a 111 kHz product of significant amplitude is quite rare.)

Some other general findings about multipath effects are[18]

- Multipath distortion is almost inversely proportional to the DU (desired-to-undesired path signal ratio) if the ratio is greater than about 10 dB.
- Baseband signal distortion increases almost proportionally to the delay time up to about 10 μS, then increases with delay time at a slower rate.

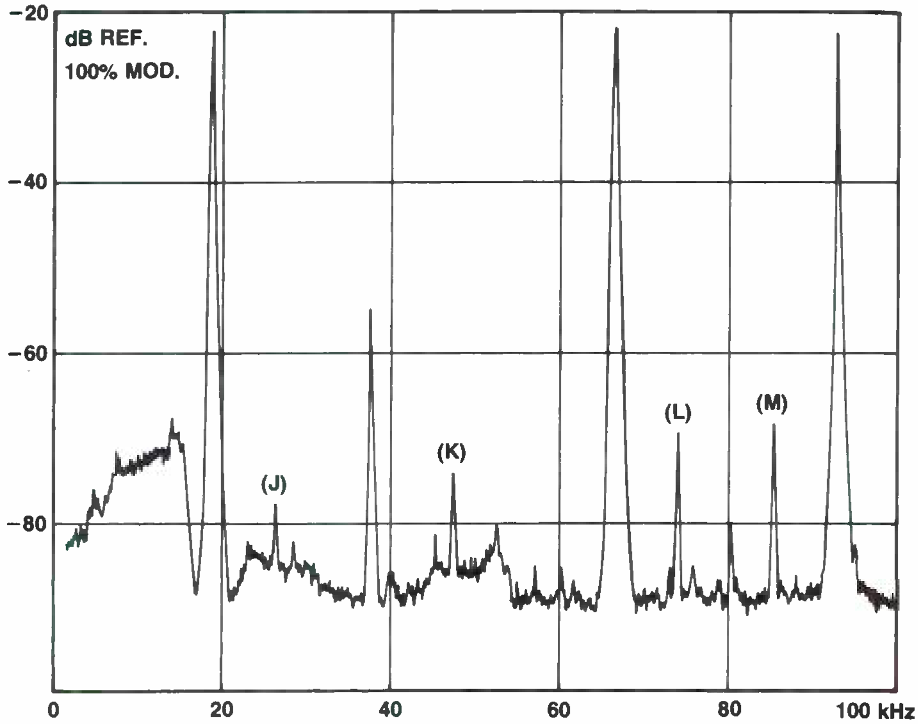


Fig. 18. Composite baseband of an FM transmitter system with pilot, 67 kHz, and 92 kHz SCAs at 9 percent injection. IM products are identified by letters and are described in the text.

- A high DU ratio required to suppress the beat-note interference to stereo program reception from an SCA subcarrier (about 20 dB at a delay time of around 5 μS and about 30 dB at around 20 μS).

It should be emphasized that the perceptibility this form of beat-note interference depends on the loudness and consistency of main channel programming, i.e., program audio in most broadcast FM stations is capable of masking the beat-note. Generally, only stations which do light amounts of audio processing and broadcast programming with wide dynamic range are even aware of an occasional case of interference.

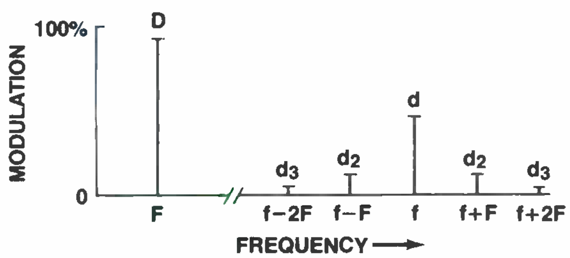


Fig. 19. Spectrum of a subcarrier "f" shown with second and third order distortion products resulting from low frequency modulation "F".

The *type* of SCA modulation or the subcarrier frequency may significantly affect the perceptibility of any beat-note. For example, properly-encoded high-speed data subcarriers using FSK or PSK spread the carrier energy in such a way that it sounds like faint, band-limited noise which is most easily masked. Furthermore, higher subcarrier frequencies will shift the beat-note image out of the audible range: experience has shown that 92 to 95 kHz subcarriers have little problem in this regard.

Crosstalk From Main Channel Program Into SCA Subchannels

Crosstalk remains a knotty problem for audio SCA services, despite the development of low-distortion receivers for SCA use. The cause is the same as SCA-to-stereo beat-notes: multipath distortion.

In the reference case illustrated in Fig. 19, the baseband signal consists of one low frequency sinewave of frequency *F*, e.g. 1kHz, which modulates the main carrier by a deviation *D* (95 percent or 71.25 kHz for broadcast FM), and a sinewave subcarrier at a variable frequency *f* Hz which modulates the main carrier with a very small deviation *d*, having a modulation index less than 0.3.

Second order distortion will cause sidebands at frequencies of $f-F$ and $f+F$ with an output injection $d_2(f-F)$ and $d_2(f+F)$. The sum of the sideband amplitudes is

$$d_2(f) = d_2(f-F) + d_2(f+F)$$

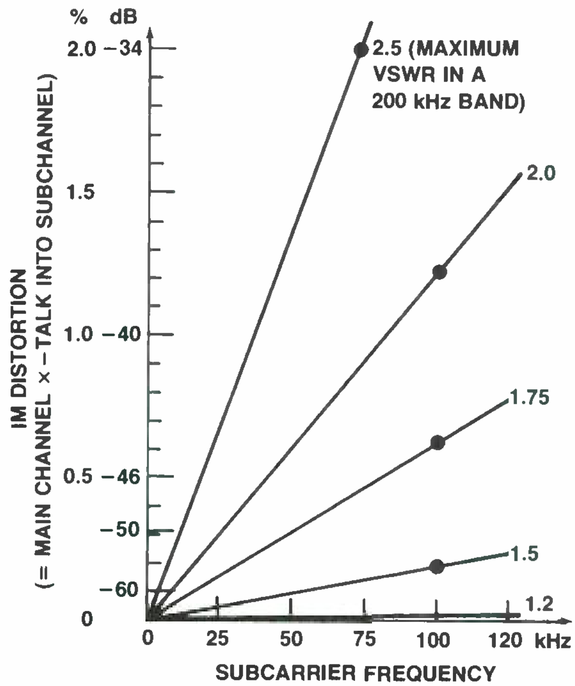
The distortion is defined as the ratio of the sum of the sidebands to the amplitude of the subcarrier d_1 is

$$\delta_2(f) = d_2(f)/d_1(f)$$

This distortion is easily measured as a function of the subcarrier frequency f with a spectrum analyzer. If the sidebands are equal, they may in the extreme either amplitude modulate or phase modulate the subcarrier. In general, there is a combination of amplitude and phase modulation.

Third order distortion is similarly defined by the sidebands created at frequencies $f-2F$ and $f+2F$.

While FM SCA detectors usually include amplitude limiting, they are intended to convert any angular modulation, whether program audio or not, into an output signal. Thus, any phase non-linearity in the RF system or multipath will generate second and higher order sidebands around the subcarrier, causing main-to-SCA crosstalk at audio modulating frequencies.



NOTE: ANTENNA IS MATCHED TO TRANSMISSION LINE AT THE CENTER FREQUENCY.

Fig. 20. Distortion due to limited antenna bandwidth vs frequency of subcarrier with max VSWR in 200 kHz band as parameter.

Fig. 20 shows the relationship between main-to-SCA crosstalk, subcarrier frequency, and maximum antenna VSWR in a 200 kHz band. It is evident that higher frequency subcarriers require somewhat larger, but not very much larger bandwidth. It has also been determined that antenna matching must be improved as the transmission line becomes longer. This is especially important for higher frequency subcarriers since the phase error is compounded rapidly with an increase in subcarrier frequency. Distortion is proportional to the subcarrier frequency. Thus, for equal distortion, doubling the subcarrier frequency requires halving the reflection coefficient.[19]

FOOTNOTES AND REFERENCES

1. Edwin H. Armstrong, "A Method Of Reducing Disturbances In Radio Signalling By a System of Frequency Modulation", Proceedings of the I.R.E., May 1936, Vol. 24, No. 5, pp. 689-740.
2. Electronic Industries Association, "Report of the National Quadraphonic Radio Committee to the Federal Communications Commission", Vol. II, Chapter 1, pp. 21-29.
3. Federal Communications Commission, "Rules and Regulations", Vol. III, Part 73.310.
4. NAB/Group W/NPR (National Association of Broadcasters, Westinghouse Broadcasting and Cable, Inc., National Public Radio), "Increased FM Deviation, Additional Subcarriers and FM Broadcasting: A Technical Report," August 30, 1983, pp. 5-6.
5. Federal Communications Commission BC Docket Number 82-536, First Report and Order (April 7, 1983), Second Report and Order (March 29, 1984).
6. John Kean, "Laboratory & Field Tests of Several FM/SCA Frequencies", National Public Radio Engineering Report, October 15, 1981, pp. 13-14.
7. Technical Manual for Moseley SCG-8 SCA generator.
8. Eric Small and Eric Stoll, "Data SCA: Some Real World Experiences", Modulation Sciences, Inc., pamphlet, August, 1984.
9. Harry Anderson and Ronald Crane, "A Technique For Digital Information Broadcasting Using SCA Channels", IEEE Transactions on Broadcasting, Vol. BC-27, No. 4, pp. 65-70.

10. *Reference Data for Radio Engineers, 6th Edition.* Indianapolis: Howard W. Sams, Inc., 1975. pp. 23-9 – 23-12.

11. Report of the National Quadraphonic Radio Committee to the Federal Communications Commission, Volume II-I, Chapter I, pp. 31-40.

12. Ibid., Supplement to Volume II, Chapter 1, Section 1.4, p. xviii.

13. Ibid., pg. 36.

14. Ibid., John Kean, pg. 24.

15. *Reference Data for Radio Engineers, 6th Edition.* Indianapolis: Howard W. Sams, Inc., 1975. pp. 23-4 – 23-12.

16. Ibid., John Kean, pg. 24.

17. Ibid., NQRC Report, Section by James Gibson, RCA, Chapter III, pg. 3/7.

18. Mitsuo Ohara, Distortion and Crosstalk Caused by Multipath Propogation in Frequency-Modulation Sound Broadcasting, IEEE Transactions on Broadcasting, Vol. BC-26, No. 3, Sept. 1980, pp. 76-81.

19. Ibid., NQRC & J. Gibson, pg. 3/11.

3.5

Television Transmitters

W.A. Resch
W.A. Resch & Associates
Quincy Illinois
C.E. Daugherty
M.R. Griesbaum
D.W. Myers
A.A. Spielbauer
Broadcast Division
Harris Corporation
Quincy, Illinois

INTRODUCTION

Many improvements have been made in TV transmitters in the last decade (1974 to 1984). These improvements, along with a tutorial description of the principal subsystems, will be highlighted in this chapter. Transmitter improvements have generally been market driven as opposed to being mandated by the Federal Communication Commission. Competition among broadcasters seeking a technical edge and among manufacturers to deliver a better transmitter at a lower cost directs the industry forward.

The advances can be divided into four broad technologies, UHF, VHF, measurements, and multichannel sound. UHF technology affects transmitters on channels 14 to 69, 470–806 MHz. In this frequency range the most important advance is the improvement in electrical plant efficiency. To the broadcaster, efficiency improvements means saving millions of dollars in energy costs each year. VHF covers channels 2 to 13, 54–216 MHz. Two areas of progress are the universal acceptance of IF modulation and the elimination of tubes in low-level stages and bias supplies. As a result transmitters are more reliable, require less routine maintenance, and deliver consistently better sound and pictures. Another VHF development is the adoption of circular polarized antennas with maximum radiated power in both the horizontal and vertical planes. To satisfy the need for double the RF power, that is equal power in both planes, higher power tubes and transmitters were developed.

The efforts to select a multichannel sound system was led by the Electronic Industry Association. The result is a compatible system with enhanced service and performance. Sophisticated signal measuring instruments have aided manufacturers and broadcasters in achieving nearly a twofold improvement in performance characteristics over the last decade.

VISUAL EXCITER

The visual exciter contains the necessary processing circuits to accept the studio video output and convert it to a fully modulated NTSC visual television signal. The functions of the visual exciter include IF and RF signal generation, sideband filtering, group delay equalization, linearity precorrection, and modulation control. These functions are grouped into circuits which are represented by the block diagram shown in Fig. 1. As a minimum, the TV exciter accepts a standard NTSC video signal, converts it to RF, and corrects for processing and power amplifier distortions. In other words, the exciter contains the necessary circuits for the transmitter to be transparent in the TV transmission system.

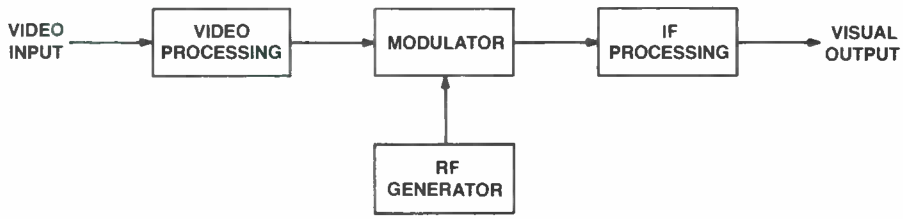


Fig. 1. Simplified visual exciter block diagram.

Basic Primer

The fundamentals of video and the associated waveforms are a foundation for understanding the purpose and operation of the overall transmission system.

Nowhere else is this more important than in understanding the operation of the visual exciter. Before introducing video processing, this section provides familiarization with units of measure, nomenclature, and signals.

The Picture

A picture is developed using a raster scanning technique and modulating the Z axis for the necessary gray shades. Each picture is defined as a frame that is created by interlacing two fields. Interlacing of two fields, each containing 262.5 horizontal lines, effectively doubles the frequency of field occurrence to 60 fields/second and eliminates visible flicker (Fig. 2).

Measurement Units

The television, or composite video waveform, consists of horizontal sync, vertical sync, video information, and test signals. This waveform and its various components are specified by the Institute of Electrical and Electronic Engineers and are defined in IRE units ranging from −40 to +100. One hundred IRE units equates to 0.714 volts peak-to-peak with +100 and +7.5 corresponding to white and black video levels respectively. Timing measurements are usually in the microsecond range.

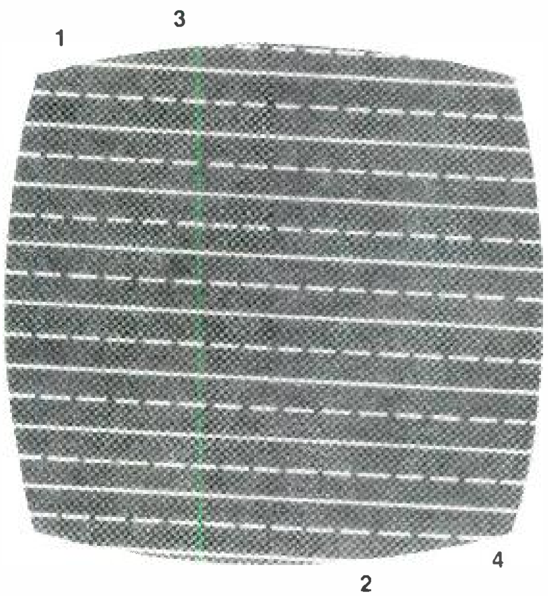


Fig. 2. One complete frame interlaced with field (1 to 2) and field 2 (3 to 4). The scanning sequence is horizontally left to right, and vertically top to bottom.

Frames and Fields

In the NTSC system, one frame of information creates a complete picture. Each frame contains two interlaced fields with a total of 525 lines. Each field contains 262.5 lines, 21 of which are blanked (The Vertical Interval) for beam repositioning to the top of the screen (Fig. 3).

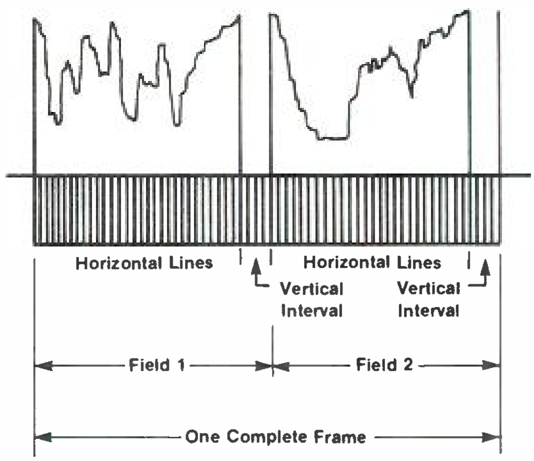


Fig. 3. The basic combination of fields and vertical intervals.

Horizontal Interval

The horizontal portion of the composite video waveform consists of horizontal lines, each containing video information, and sync pulses which provide blanking, retrace time, and other information. A detailed picture of horizontal line sync pulses, various IRE levels, and timing information are valuable in the video world (Fig. 4). Descriptions of various events within the horizontal line provide insight into the many purposes of the waveform.

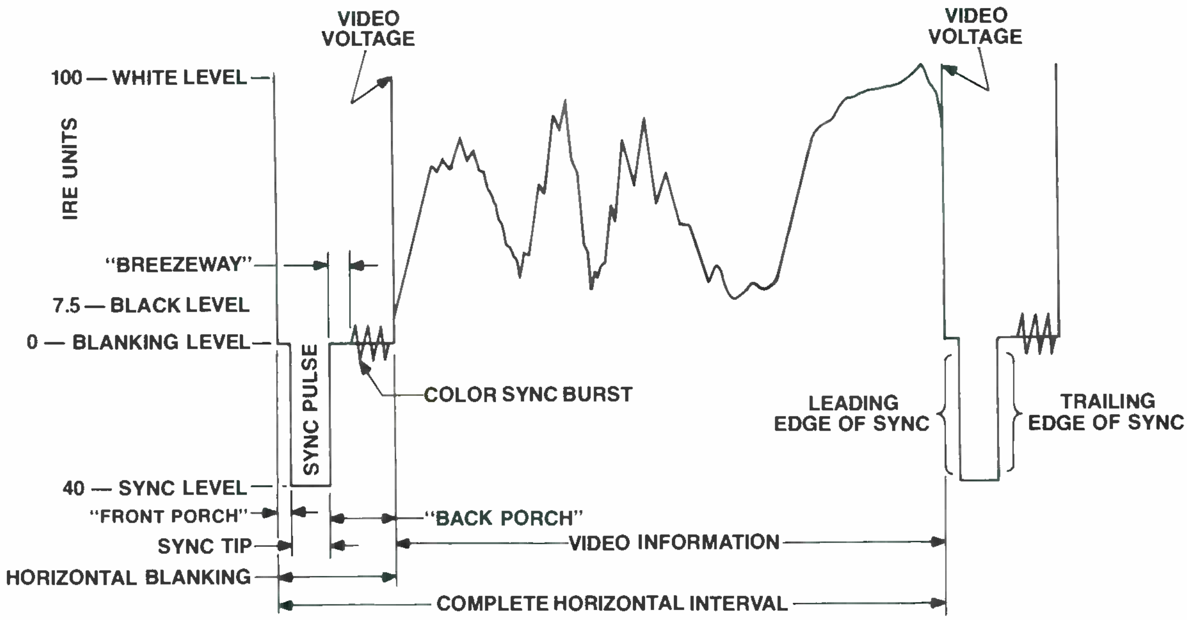


Fig. 4. Horizontal line.

The "front porch" prevents video voltages from prematurely triggering the sync circuits by isolating transients from the sync pulse. The horizontal sweep oscillator is reset by the leading edge of the sync pulse, while the tip of the pulse is a reference point for dc restoring in receiving sets. The "back porch" provides time for the blanked beam to return to the left side of the screen. During color transmission, a 3.58 MHz sine wave color sync burst is added to the "back porch" to frequency and phase lock the picture color information. Video information is contained between the end of the "back porch" and the beginning of the next "front porch" and is transmitted by modulating the carrier signal from black to white levels.

Vertical Interval

Prior to sweeping each field, the beam must be repositioned vertically. The repositioning time, called the vertical blanking interval, is composed of 21 horizontal lines which are not displayed. This portion of the composite video waveform is extremely important since it contains critical pulses, test signals, source identification codes, reference signals, captioning and teletext (Fig. 5).

The first six pulses, known as equalizing pulses, synchronize video information in fields 1 and 2. These pulses occur at twice the horizontal sync rate (period = 63.5 μS, 15.75 kHz) and assure that vertical triggering occurs at the same time for odd and even fields.

The serrated vertical (sync) pulses follow next. The second set of six equalizing pulses ensure field frequency regularity. The remainder of the 21 blanked horizontal lines contain various test and information signals, including teletext and captioning.

The tradition of using test signals during the vertical interval began with FCC Rules and Regu-

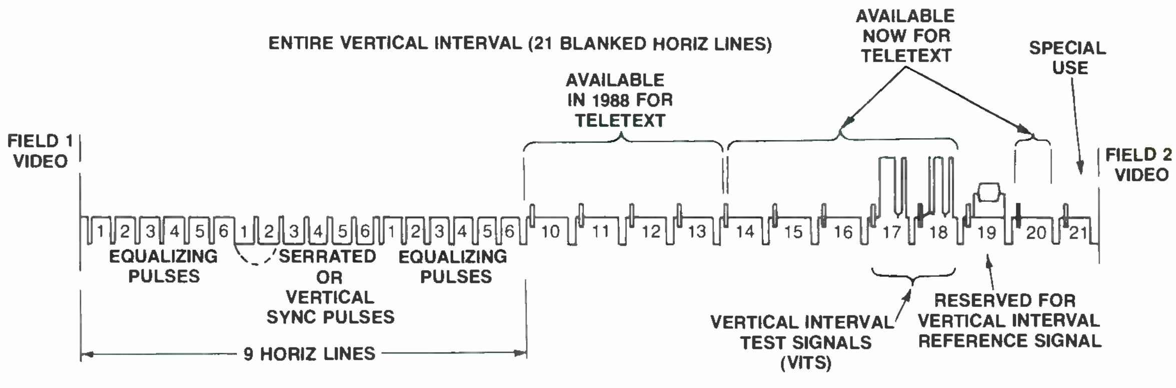


Fig. 5. Entire vertical interval (21 blanked horiz lines).

lations. Recently these test signal requirements have been relaxed. The test signals themselves continue to be in wide spread use.

Vertical Interval Test Signals (VITS)

Vertical Interval Test Signals (VITS) are used to evaluate various parameters of a broadcast system performance and are monitored daily. The VITS are transmitted during active operation to ensure continuous quality and accuracy in terms of color and distortion. In regard to transmitter performance, the quality of the VITS frequently determines if a problem warrants a trip to the transmitter site.

The type of test signal, as well as specific line location, is determined by individual networks, stations, and the FCC. Generally lines 17 and 18 are used for VITS: multiburst on field 1, line 17 Fig. 6; color bars on field 2, line 17 Fig. 7; composite radiated signal on fields 1 and 2, line 18 Fig. 7. In cases where the network determines the type and location of the VITS, several alternatives are available. These signals may be on line 17 or 18, field 1 or 2. Some frequently used VITS are highlighted with measurement applications emphasized in the following paragraphs.

The multiburst (Fig. 6), a VITS aimed directly at frequency-gain characteristics, also monitors setups and frequency selective compression. This

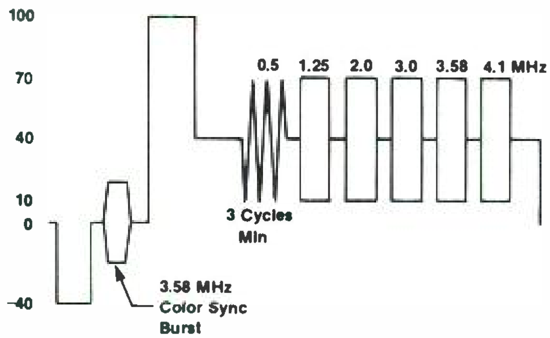


Fig. 6. Multiburst.

test pattern usually consists of a white bar (100 IRE) followed by six bursts of sine wave frequencies ranging from 0.5 MHz to 4.1 MHz. The six bursts must have equal amplitude for the correct response.

The Composite Radiated Signal (Fig. 7) contains several distinct signals, each having its own diagnostic capability. Differential gain and phase are evaluated through the modulated stairstep. This signal is composed of six equal amplitude steps, ranging from −20 to 100 IRE levels, each step is modulated with a 3.58 MHz sine wave burst on its dc level. Variation in sine wave amplitude is a measure of differential gain. The 2T

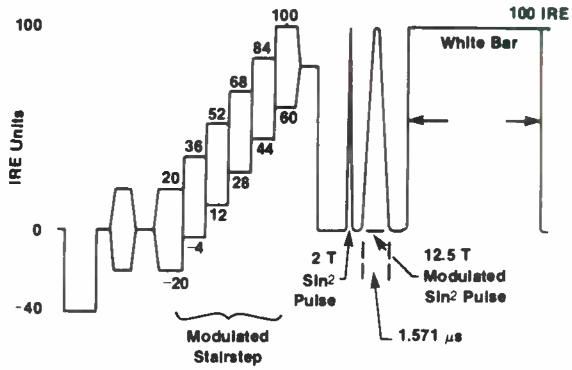


Fig. 7. Composite radiated signal.

sine-squared pulse is especially useful for determining frequency response and group envelope delay. The relative amplitude of the 2T pulse in relation to the white bar indicates frequency response while the symmetry of the 2T pulse indicates delay. The 12.5T modulated sine-squared pulse provides an indication of amplitude frequency response errors where the 2T pulse is less sensitive, namely from 3 to 4 MHz which are the chrominance frequencies. The detection of chrominance/luminance gain and delay, which cause color saturation and distortion problems, is easily accomplished with this pulse.

The color bar test signal (Fig. 8) is one of the most widely used test waveforms in color television broadcasting. The waveform consists of the three primary colors (red, green, blue), their complements (cyan, magenta, yellow), and a white reference bar. The color bars are arranged in order of luminance, with the brightest colors on the left and the darkest on the right. This signal provides a quick method of assessing color transmission quality. The confirmation of performance, either satisfactory or unacceptable, is the oscilloscope's main objective. Beyond a status indication a vectorscope must be used to evaluate the encoding and processing of color signals.

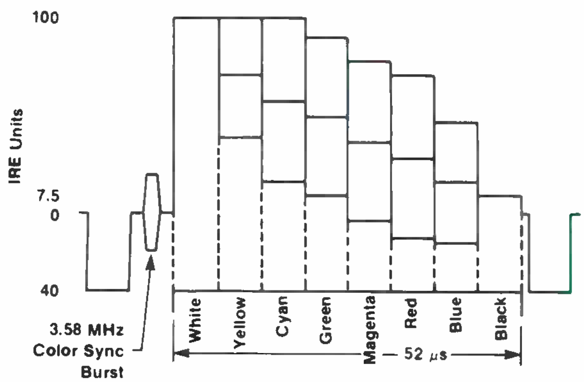


Fig. 8. Color bar test signal.

NTSC Color Transmission System

Two items were considered in adopting the NTSC system: 1) the pattern resolving power of the eye for patterns having only color contrast is considerably less than when there is also luminance contrast, and 2) the smaller an object is, the more difficult it is to determine its hue. This means that for the finer picture details, loss of chrominance resolution will not necessarily be noticed. The coding of chrominance signals also takes advantage of variations in the resolving power of the eye for colorimetric variations.

Initially, the light reflected from the scene being televised is optically separated by color selective filters. Three distinct colors (red, blue, and green) result, each activating a color pickup tube in the camera and converting the light into their respective signals. In an effort to transmit the minimum amount of information needed to reconstruct the actual scene, the colors are combined to give precise information about the brightness, saturation, and hue.

A portion of each color signal is matrixed to generate a luminance signal, which completely describes the brightness of the scene. A complete description of the scene requires precise information about each color, and three linear combinations of the color would satisfy all of the receiver requirements. Since one linear combination is transmitted with the luminance signal, the two remaining sets of information are transmitted with color difference signals. These signals, I and Q, are composed of linear combinations of each color and completely describe the scene. The two color difference signals each amplitude modulates a subcarrier of identical frequency but different phase, which are added to produce the chrominance signal. The resultant is an amplitude and phase modulated wave. In the simplified color diagram (Fig. 9), phase represents the characteristic angle (hue) on the vector diagram, and amplitude determines the degree of saturation.

At the receiver, a method of synchronous detection is used to separate the color signals which are applied to the control grids in a tricolor viewing tube. This is only possible if the phase oscillators are stable. A 3.58 MHz reference burst at the beginning of each line reestablishes the correct phase relationship. This method of color transmission links the chrominance directly to the phase modulation of the subcarrier.

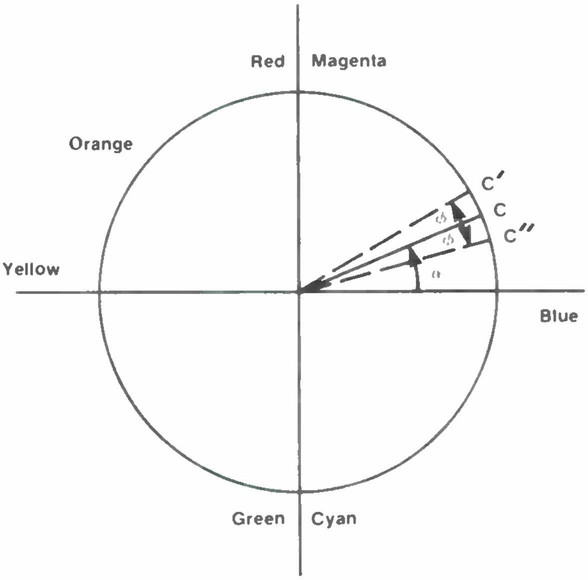


Fig. 9. Simplified color diagram.

Video Processing

The division between studio video signal processing and the transmitter video processing can be difficult to separate. Often times video automatic gain control, sync and burst regeneration, white limiting, cable length compensation, dc restoration and ac hum canceling circuits are required for a plant to operate satisfactorily. These problems usually arise from the distribution of the video signal rather than in the studio production or transmission processing. Traditionally video processing amplifiers, called proc amps, are used to clean up distribution distortions.

The video processing circuits commonly included in a transmitter are dc clamp, differential gain, and differential phase precorrectors. These circuits condition the signal prior to modulation.

Clamp Circuit

The clamp circuit or dc restoration circuit reestablishes the proper dc video signal level. This function is important because picture brightness information is contained in the dc video level. Since dc information can not be transmitted by a single sideband signal the TV signal is AM modulated about carrier.

In order to maintain accurate video amplitude information, the video signal is clamped to a static portion of the horizontal waveform. Clamping the video to a fixed reference voltage represented by the static portion of the waveform allows further amplitude dependent processing to be done prior to modulation.

Hum distortion can be observed in the picture when alternate light and dark bars on the picture appear to move slowly up the picture. The bars are caused by the sinusoidal effect of increasing and decreasing brightness at a low-frequency rate. The slow movement is caused by the vertical frame-rate being slightly slower than the line frequency. The number of bars indicate the frequency of distortion. One dark bar indicates 60 Hz distortion while two dark bars indicate 120

Hz distortion, etc. Care should be taken in attributing a picture distortion to one part of the transmitter. Hum bars, for instance, can also be caused by power supply filter failure or even produced in the monitor itself.

Fig. 10 is representation of a video signal with hum superimposed. This hum can be conducted interference induced by differences in ac grounding potentials between video distribution equipment. It could also be caused by inductive pick up of radiating ac magnetic fields emanating from an ac transformer or power supply choke.

The horizontal sync is used to trigger an electronic switch which inserts a dc reference potential to a portion of the static waveform either during sync or burst. Fig. 11 shows the same signal after being clamped. The horizontal line period is many times faster than the hum period so by forcing the video to a reference voltage during a static portion of the video, the line-to-line change is practically not discernible. By clamping in the transmitter, ac coupling of the video signal can be allowed in the signal distribution system. AC coupling is convenient since differences in ac and dc ground potentials can be ignored without causing picture distortions.

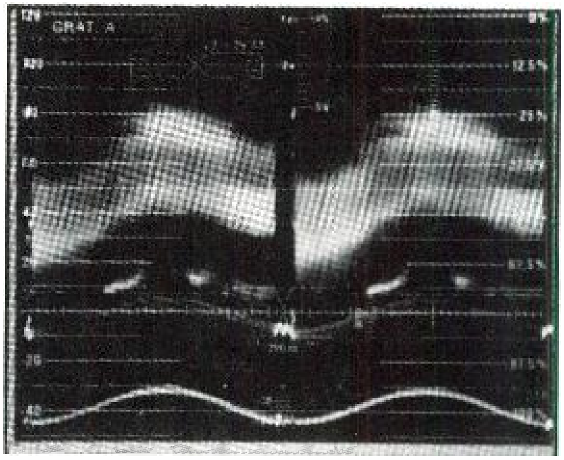
Video Precorrection

Errors in chrominance amplitude and phase amplifications that are level dependent are often corrected at video. A picture affected by nonlinear chroma gain shows a change in the saturation of the colors. A piecewise linear gain curve is normally fitted by using diodes which are adjusted to conduct at preset amplitude levels. Gain can be controlled to either increase or decrease as needed. This type of corrector is called differential gain.

Differential gain correctors are usually broadband in that gain errors are corrected equally with respect to frequency. These correctors assume that luminance and chrominance nonlinearities are equal.

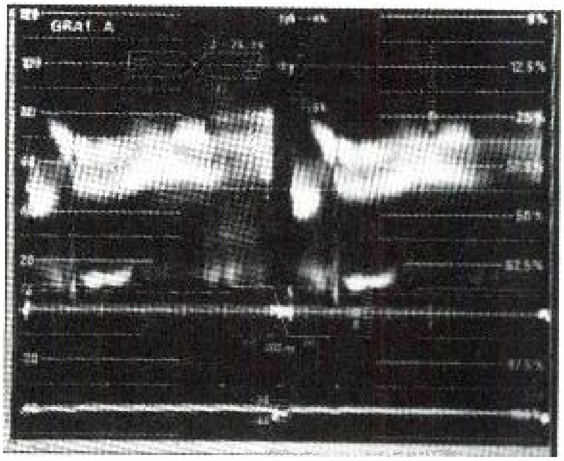
Luminance nonlinearity is the change in picture brightness. A gain corrector affecting only low-frequency luminance can be implemented by first separating the signal into frequency bands and then recombining after amplitude precorrection.

Chrominance phase nonlinearity, called differential phase, is the change in chrominance phase as the corresponding portion of the luminance signal amplitude is changed. The effect on the picture is a change in hue of colors. The signal is split into two paths. One path has a 90° phase shift at 3.58 MHz chroma subcarrier frequency. The other path contains the diode gain breaks. Recombining the inphase and quadrature signal allows phase variations of ±10° independently of gain errors.



1. Observe the signal with the waveform monitor DC RESTORER OFF and in SLOW clamp mode. Mains hum, if present, will be evident in a 2 field display.

Fig. 10. AC Hum, field display.



2. The FAST DC RESTORER of the waveform monitor eliminates hum on the display.

Fig. 11. Signal after clamping.

Modulator

FCC type accepted transmitters using IF modulation first became available in the early 1970s. Now it is a universal practice. The reason for its popularity is the ease at which precise IF signal processing can be implemented and the convenience of field adjustment. The modulator is usually a broadband, balanced diode mixer. It is configured for maximum rejection of the oscillator signal and biased so as to provide linear, low noise and low modulation capability. A schematic diagram of a typical balanced mixer is shown in Fig. 12. The video signal is offset to provide the proper modulation level. The video signal is used to control the attenuation of the

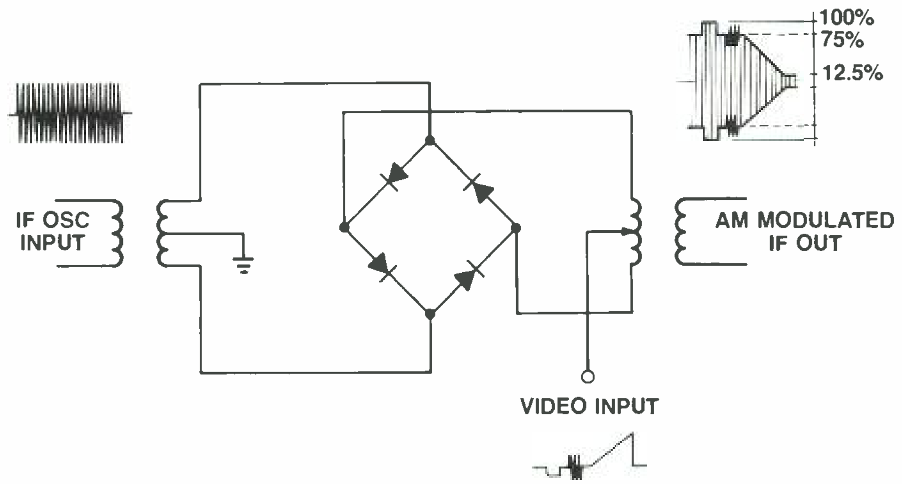


Fig. 12. Balanced mixer.

diodes. Peak of sync corresponds to maximum IF envelope output and white corresponds to minimum IF output.

The output of the modulator is a double sideband AM signal having the proper depth of modulation.

IF and Video Delay Compensators

Picture and waveform distortions can be caused by group delay errors in narrow band RF power amplifiers and tuned RF output systems. Group delay error is the nonuniform group delay over the bandwidth of the TV signal. In a bandpass filter, spectrum components on the edge of the passband are delayed more than components in the center of the passband. In general, the closer the amplitude roll-off frequency is to the visual passband, the higher the delay error.

Group delay impairments show up as color smear and halo effects on edges. On a waveform monitor the effects may be seen using the 2T and modulated 12.5T contained in the composite test signal shown in Fig. 7. Pulse responses may include exaggerated pre or post ringing on 2T and modulated 12.5T base line disturbance. The multipulse signal shown in Fig. 13 may also be used.

The multipulse signal consists of a gray flag (80 IRE), 2T pulse, and five sine-squared pulses modulated with five discrete frequencies (consisting of one 25T pulse and four 12.5T pulses). The multipulse may be used to measure linear distortion in TV systems, such as gain/frequency distortion (frequency response) and delay/frequency distortion (group delay).

The modulated pulses contain two spectra of information, the low frequency and the high frequency, as illustrated in Fig. 14. These modulated

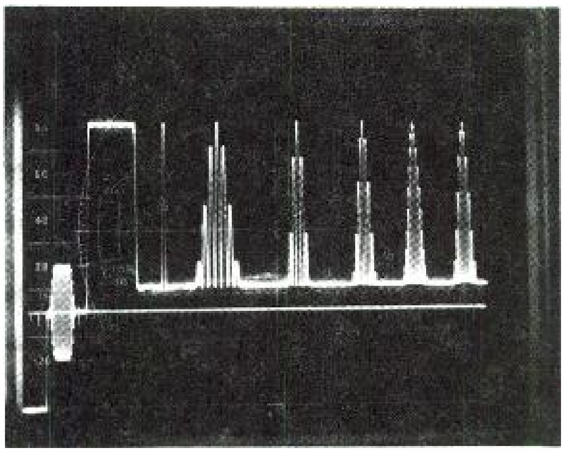
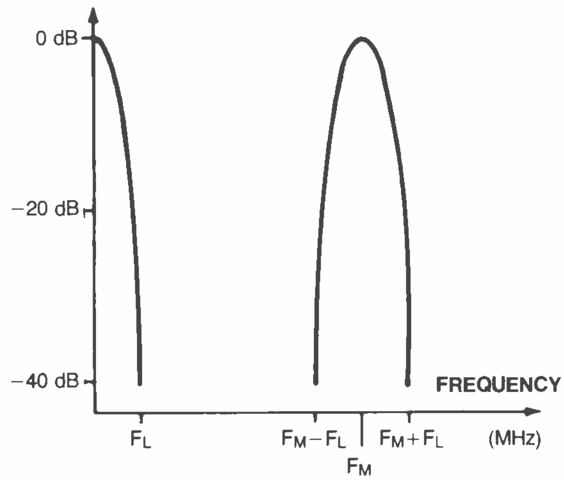


Fig. 13. Multipulse.



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

F_M = MODULATING FREQUENCY

Fig. 14. Energy spectrum for modulated pulses.

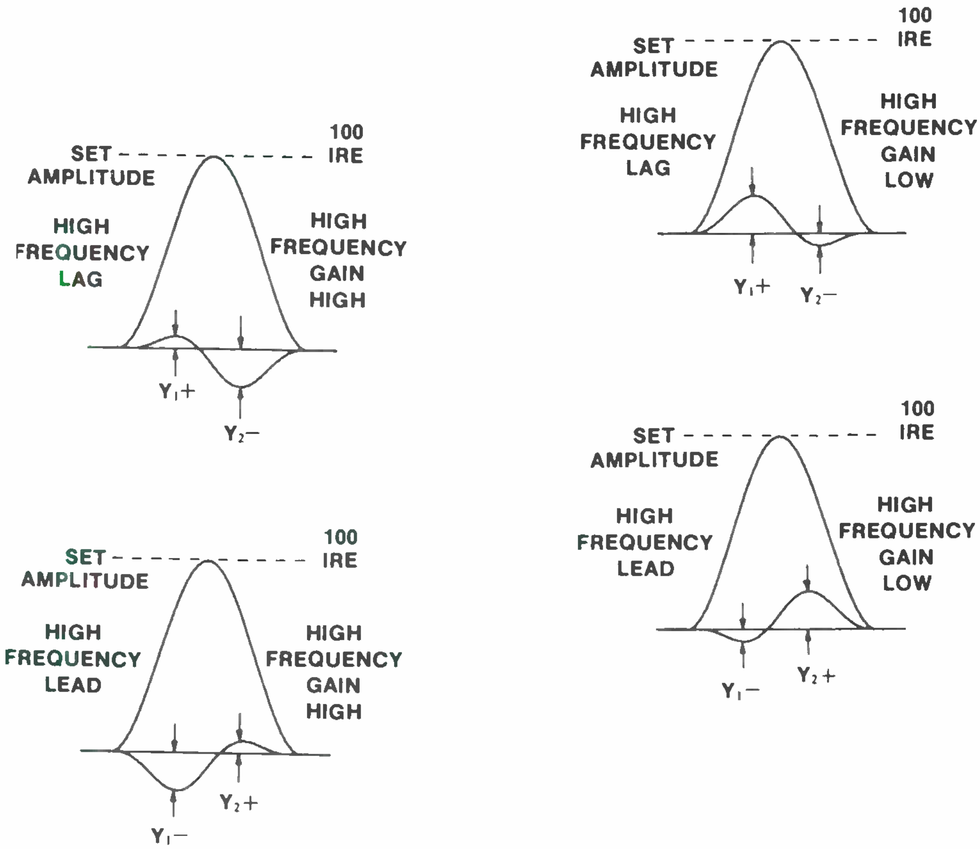


Fig. 15. Modulated sine-squared pulses with gain and phase errors.

pulses are useful for measuring group delay errors. If the low-frequency spectrum and the high-frequency spectrum are delayed equally, the results will be a symmetrical modulated pulse with the same shape as the input. The gain/frequency distortion will alter the base line flatness but will not change pulse symmetry. Delay errors will result in an asymmetrical pulse baseline. Combinations of delay and gain errors are shown in Fig. 15.

Low-frequency group delay and amplitude errors are referred to as short time waveform distortions. Fig. 17 shows typical distortions of a 2T pulse.

A method of quantifying these distortions is to use the waveform graticule shown in Fig. 16. This graticule was arrived at empirically and represents constant perceptible distortion levels. It shows that overshoots closer to the desired pulse are not as perceptible as ringing further away.

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

If the group delay error affects only the high frequency side of the passband, precorrection can be accomplished at video. This is the case for most visual-aural RF notch type diplexers and FCC receiver equalizer curve. In the case of the notch diplexer, a pair of cavities resonant at the aural frequency are inserted between two 90° hybrid couplers. Refer to Fig. 79. The presence of the tuned circuits near the upper edge of the visual passband can cause significant group delay error.

Another source of group delay is the tuned circuits of the high power tetrode or klystron am-

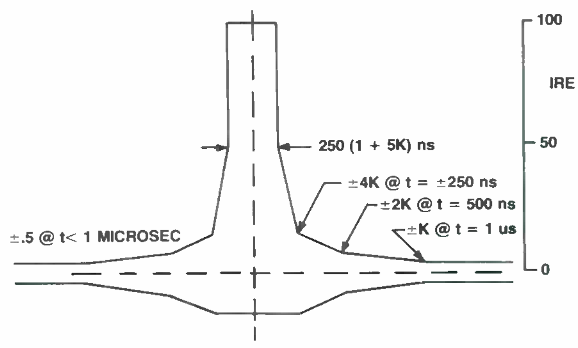
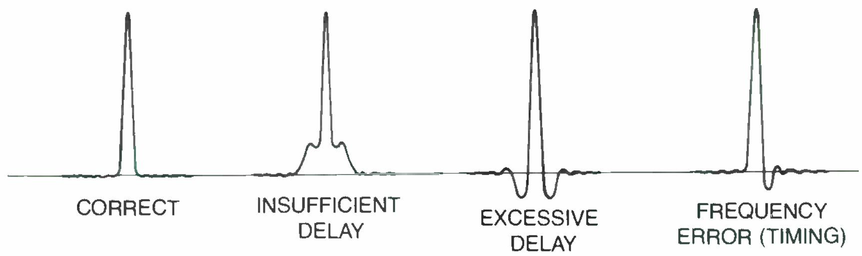


Fig. 16. 2T sine-squared pulse graticule.



TYPICAL 2T WAVESHAPES

Fig. 17. Typical distortion of a 2T pulse.

plifiers. Finally the FCC requires the TV transmitter to predistort group delay according to the curve shown in Fig. 18.

The delay shown in Fig. 18 is a characteristic delay curve the FCC requires all visual transmitters to have. The purpose of it is to compensate for the delay error caused by discrete LC aural notch filters in early TV receivers. The theory was that it would be far cheaper to group delay compensate one transmitter than every TV set. The notch filter in TV receivers is necessary to prevent the aural carrier from mixing with the detected video and chroma subcarrier signals and producing visible spurious beats.

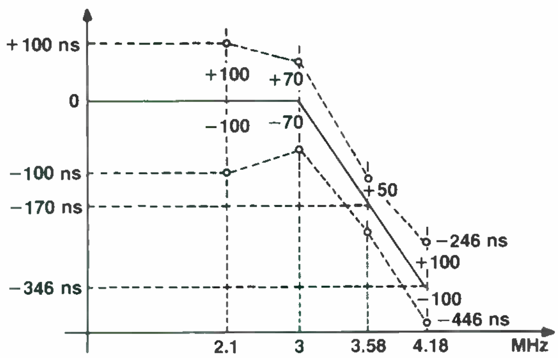


Fig. 18. Predistorted group delay curve.

Group delay predistortion is accomplished at video or IF. The techniques used are similar in concept. Both active and passive equalizers are employed. IF group delay correction is necessary to correct group delay errors associated with the lower sideband below visual carrier. Above visual carrier both IF and video correction is effective.

Passive Group Delay Equalizer

A common form of a passive video delay equalizer is shown in Fig. 19. It provides a flat frequency response and a nonlinear group delay which peaks at the resonant frequency of the circuit. This type of circuit is referred to as a passive all-pass network.

Without going through a detailed analysis of the circuit much can be understood by examining the all-pass at frequencies well below and above resonance. At low frequency the network can be approximated by a low-series inductive reactance due to L1. Here the output voltage leads in phase near zero. At high frequency well above resonance the circuit can be approximated by a series capacitive reactance due to C1. At high frequencies the output lags in phase near zero or more correctly 360°.

The output amplitude is a constant. The phase of the network is plotted in Fig. 20. The slope of the phase change is defined as group delay. The steeper the slope the higher the maximum group delay. A plot of the group delay is shown in Fig. 21.

Active Group Delay Equalizer

As in the passive type the network has a constant amplitude frequency response and a nonlinear phase response. The action of an active allpass network can be best explained using the simplified schematic of Fig. 22. If the voltage e_1 and phase were plotted as a function of frequency, it would trace out a circle with maximum

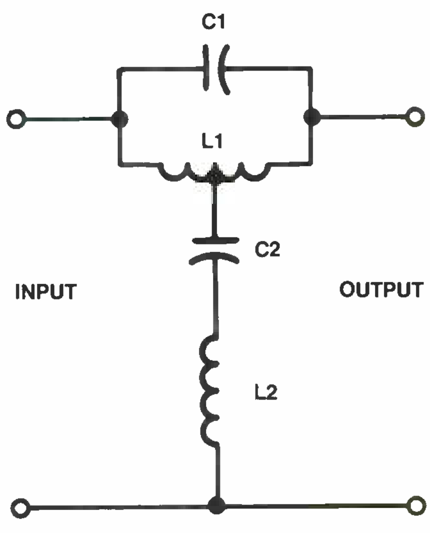


Fig. 19. All-pass network.

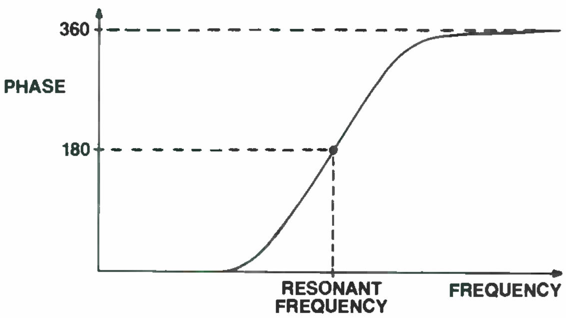


Fig. 20. All-pass phase.

amplitude and zero phase at resonance. Voltage e_2 on the other hand has a constant amplitude and phase. For the case when e_1 maximum is equal to twice e_2, the e_3 output of the summing amplifier A1 has the characteristics of an ideal all-pass network. The output amplitude is constant, the resonator tuning determines the frequency of maximum group delay and the Q determines the magnitude of delay.

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

Vestigial Sideband Filter

The FCC requires that the radiated TV signal suppress a major portion of the lower sideband. In addition, on the high frequency side the signal must be contained within 4.75 MHz of the visual carrier. Originally, the bandshaping was performed at the output of the transmitter. With the advent of IF modulation the filter has moved to low-power stages and at first was built using discrete inductors and capacitors. More recently transmitter manufacturers have selected solid state filters using Surface Acoustic Wave (SAW) technology.

The term Surface Acoustic Wave (SAW) refers to the propagation of elastic waves on the surface of a piezoelectric crystal. The propagation is roughly the speed of sound, and therefore called acoustic. A time varying voltage on a

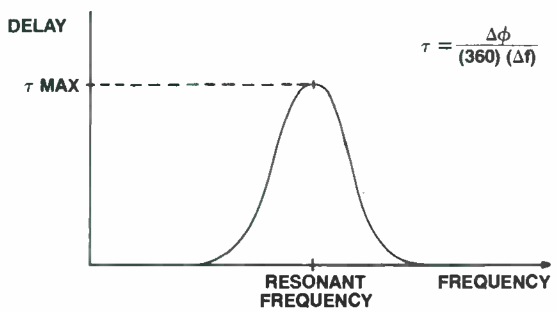


Fig. 21. All-pass group delay.

metalized transducer is used to induce a deformation on the surface and produces a electromechanical wave. The deformations produce local electric fields which travel along with the mechanical wave, and interact with metal electrodes, which convert the mechanical wave back to a time varying voltage. The length of the electrodes, the spacing and the number determine the wave shaping properties of the filter.

The electrodes act like tapped delay lines, as illustrated in Fig. 23. The electrodes are designed to amplitude scale and delay the signal. The output is selectively attenuated depending on the time delay and signal frequency relationship. This type of filter is called transversal because the attenuation is controlled by delay lines rather than resonators. The transversal filter was invented in the early forties and used coax cables as delay lines.

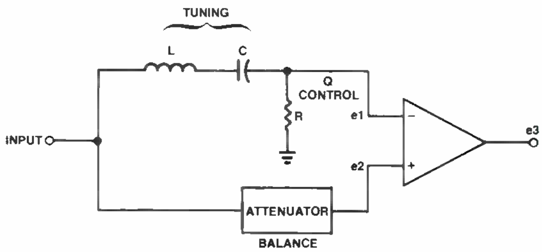


Fig. 22. Active all-pass network.

Today, surface acoustic wave transducers provide the same function. The wavelength of a 37 MHz IF acoustic signal is approximately .003 inches.

This small size allows the filter to be very small and compact. Because the transducers are only on the surface, photographic masks can be used to accurately control the physical dimensions of the transducers. The photographic mask lends

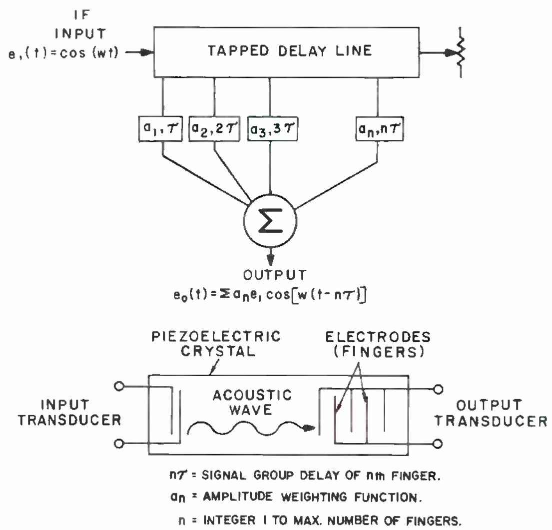


Fig. 23. Saw filter.

itself to modern manufacturing techniques and insures a reproducible filter with a permanent attenuation and group delay.

A characteristic of the transversal filter is that group delay can be set independently of the amplitude characteristic. This is not true of discrete LC filters where basic physical laws couple group delay and bandwidth. The independent nature of group delay and bandwidth allow the manufacturer to provide ideal pulse response without the need to use group delay equalizers. The FCC group delay characteristics shown in Fig. 18 can also be built in.

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

Common distortion products associated with SAW filters are due to spurious signal reflections and direct feedthrough. These distortions cause time echoes displaced either before or after the desired responses. In the frequency domain these echoes show up as ripples in the passband. Ripples are related to spurious time echoes as shown in Fig. 24. A given time echo will contribute uniquely to the amplitude, phase and group delay characteristics by the addition of a ripple component in the passband. The amplitude of the ripple is proportional to the echo level, and its period is the reciprocal of the time displacement.

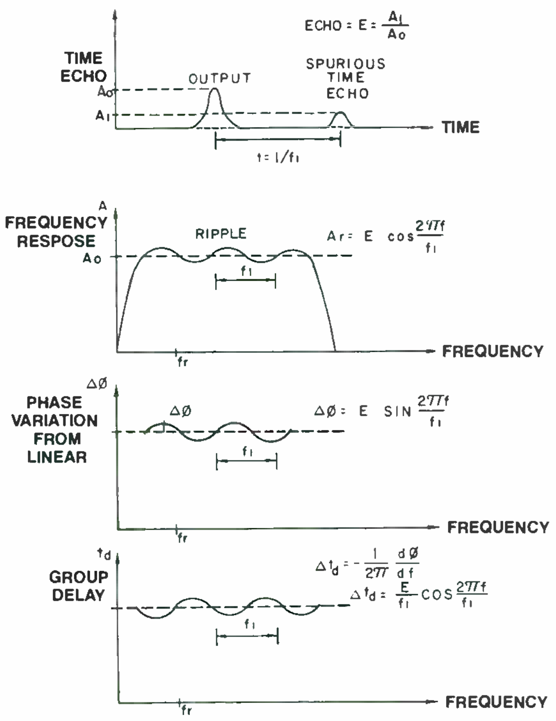


Fig. 24. Response, phase, and delay ripples caused by time echoes.

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

A better test is to measure group delay distortion using the pulse waveforms in Figs. 7 and 13. The 2T preshoot, overshoot, and dissymetry can be used to gauge low-frequency group delay error. Modulated Pulse baseline S curve is a result of group delay error at the modulated carrier and the peak excursions of the S curve can be used to access high frequency group delay.

IF Linearity Precorrection

It was pointed out in the section on Video Precorrection that amplitude and phase nonlinearities are sometimes corrected at video. The alternative is to provide correction at IF. There are advantages to correcting at IF in that since most distortions are caused in the high power RF amplifiers after vestigial sideband filtering, a corrector placed after the vestigial sideband filter can more accurately predistort the modulated signal in a manner consistent with the sideband energy. This is especially true of chroma distortions. 3.58 MHz chroma has only a single sideband information at RF and thus has 6 dB less energy than the luminance signals. The RF power spectral density is altered by the vestigial sideband nature of the TV signal. An ideal diode detector frequency response of the NTSC modulated signal is plotted in Fig. 25. Here it is seen that beginning at .75 MHz the video begins to fall to -6 dB. For video signals lower than .75 MHz, the RF is double sideband and has twice the peak RF voltage.

Intermodulation products are caused in high power amplifiers by nonlinear amplitude and/or phase transfer characteristics. As the power output increases towards saturation amplitude compression begins and in general the phase begins to lag as well due to electron crowding in the

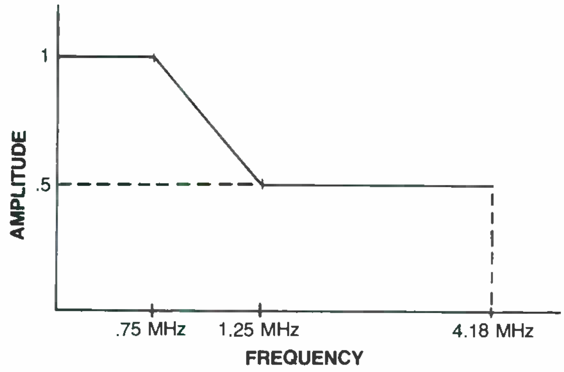


Fig. 25. Frequency response using ideal diode detector.

amplifier device. The nonlinear amplifications give rise to mixing products which occur at sum and difference frequencies around the visual carrier. The amplitude compression distortions produce spectral components which are in phase with the modulation signal. Correctors for compression distortion are usually similar in concept to video differential gain correctors. The linearity correctors generally use biased diodes to modify gain as described earlier in the video processing section. Implementation, however, is generally more difficult because for given diode bias the peak-to-peak IF envelope amplitude requirement is twice that of video. In addition, since the susceptibility to reactance drift varies with frequency, the design and manufacturing tolerances for IF correctors are several orders of magnitude more stringent than video correctors.

A basic gain expansion circuit is shown in Fig. 26. The signal is normally attenuated a fixed

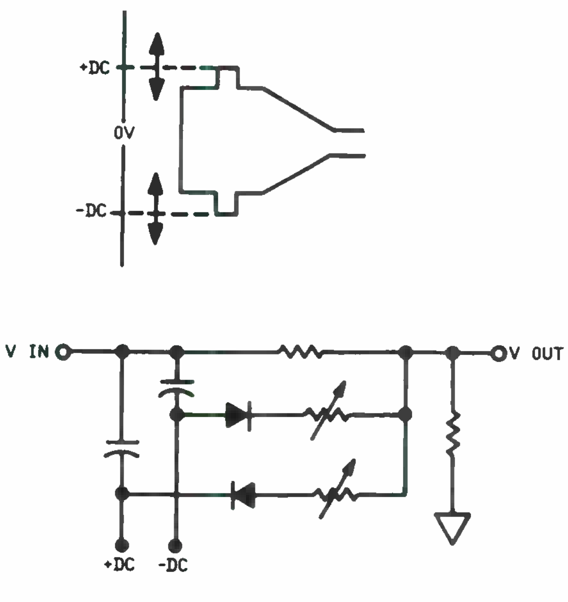


Fig. 26. Basic gain expansion circuit.

amount by using a resistive L-pad. The diodes are normally reverse biased by equal, but opposite polarity, dc voltages. Reducing the dc voltage amplitude permits the diodes to conduct on the signal peaks. This inserts additional resistance in parallel with the series arm of the L-pad thereby decreasing the attenuation. Varying the resistance in series with the diodes provides for a variable gain expansion.

The phase distortions produce Incidental Carrier Phase Modulations (ICPM) which are bessel type spectral components in quadrature with the modulation signal. For a given level dependent phase shift the amplitude of the spectral component will increase directly with the video frequency. The spectral amplitude frequency relationship is consistent with phase modulations. In other words fast video amplitude changes such as a step or pulse will cause larger phase spectral components than slow changes. Receivers make this condition worse by not detecting the lower sidebands below .75 MHz. The receiver then responds to the sidebands as if they were amplitude single sidebands and produce spikes. The faster the rise time on the signal the more high frequency energy is present resulting in edge distortions.

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

Audio impairment is produced by ICPM in receivers employing intercarrier detection. Intercarrier receivers use an AM or synchronous detector to produce a 4.5 MHz aural IF signal. Any phase modulation present on the visual carrier is then transferred to the aural IF carrier. In the L + R or monaural baseband audio the increasing amplitude frequency effect of ICPM is nullified by de-emphasis to some degree. With the advent of multichannel sound, however, the distortion is more pronounced at the subcarrier and pilot frequencies. To counteract the effects of ICPM and other noise sources on subcarriers an elaborate audio companding system has been adopted which greatly reduces the potential interference. Although the audio companding process reduces the effects of ICPM, ICPM correction is an important aid in delivering clear, low-noise audio to intercarrier receivers.

ICPM correctors can be grouped into two types, ones using a phase modulator and the other operating on the signal directly. The phase modulator type uses video to vary the IF or master oscillator phase with the opposite phase characteristics of the nonlinear amplifiers. A block

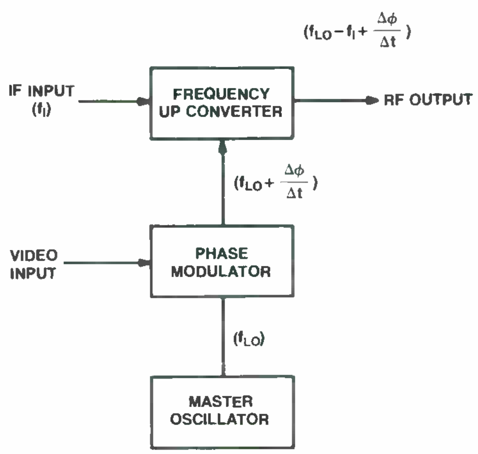


Fig. 27. Master oscillator phase modulator block diagram.

diagram of a master oscillator phase modulator is shown in Fig. 27.

ICPM correctors operating directly on the IF signal can be implemented several ways. Direct correction at IF is similar in concept to baseband phase correction. In both cases the video is split into two paths which are in phase quadrature. In the IF corrector the entire channel is in quadrature, whereas in the video corrector only the chroma band is in quadrature (Fig. 28).

One implementation is to modify the quadrature signal gain function with level dependant diode breaks. This can be done using the same techniques as in the linearity corrector.

RF Generator

Transmitters employing IF modulation generate the following frequencies; visual IF, modulated aural IF, and a common master oscillator for translating visual and aural IF to final carrier frequencies. The oscillator schemes have been implemented with either digital synthesizer techniques or crystal oscillators. An advantage of the synthesizer is that only one crystal which may operate at a single standard frequency for all TV channels is required. The multiple crystal approach, however, involves simpler circuitry.

The two commonly used IF frequencies are 37 MHz and 45.75 MHz. There are many reasons for selecting one IF frequency or the other. One advantage of 37 MHz is that the temperature drift sensitivity of most IF components such as the SAW filter is related directly to carrier frequency. Thus the lower IF has a 12 percent less drift sensitivity than components at 45.75 MHz. On the other hand, 45.75 MHz is a common demodulator IF which can be useful for IF troubleshooting, and temperature drift may be minimized at either IF by maintaining the filter at a stable temperature.

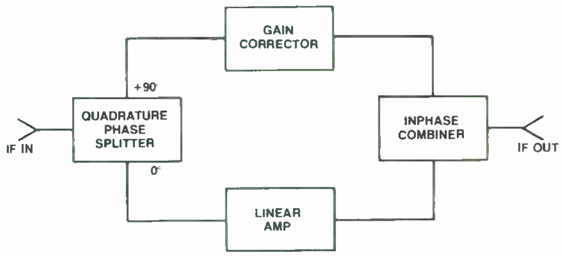


Fig. 28. Direct IF ICPM corrector.

The important performance characteristics of an oscillator is its low-phase noise, low sensitivity to frequency drift with time and temperature, and resistance to mechanically induced phase and frequency shifts called microphonics.

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

AURAL EXCITER

BTSC Multichannel Sound System

The combination of dbx companding and the Zenith Multichannel Television Sound Transmission System (referred to as the BTSC—Broadcast Television Systems Committee—System) provides high-fidelity stereo performance for all viewers within the present range of TV signals (grade A and B areas).

The introduction of TV stereo and SAP into the spectrum requires that the person or persons maintaining the entire aural transmission chain from audio input to transmitted RF fed to the antenna be aware of the BTSC system. The following paragraphs offer an overall view of the BTSC system.

The transmission spectrum of the BTSC system is graphically illustrated in Fig. 29. A total system block diagram is presented in Fig. 30.

Except for the stereo subcarrier center frequency, the system is similar to that used for transmitting stereo FM in the United States with the pilot frequency being the video horizontal-scanning frequency of 15.734 kHz ($1f_H$).

To keep the signal compatible with mono television sets, the L + R signal is transmitted in the same spectrum space previously occupied by the mono audio signal. Stereo information is encoded in the L − R channel which will only be decoded by stereo compatible television sets while mono sets will ignore the L − R subcarrier. The stereo set uses the L − R signal to reconstruct the original left and right audio signal by adding and subtracting the L + R and L − R signals.

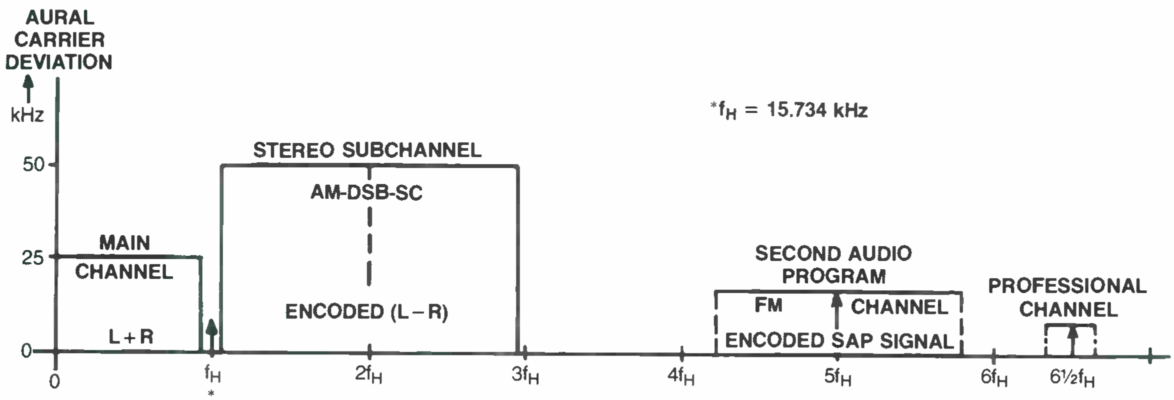


Fig. 29. BTSC system frequency spectrum.

The main channel aural carrier modulation consists of an (L+R) audio signal with a pre-emphasis of 75 microseconds. The (L−R) audio signal is subjected to compression according to a companding system that complements the expansion in the receiver. The compressed L−R signal causes double sideband, suppressed carrier amplitude modulation of a subcarrier at 31.468 kHz ($2f_H$). The audio bandlimits of both pre-emphasized L+R and of the encoded L−R are 50 Hz and 15 kHz.

The main channel (L+R) peak deviation is 25 kHz. The stereophonic subchannel (L−R) peak deviation is a maximum of 50 kHz. When L and R are statistically independent, the peak deviation of the main channel and the stereophonic subchannel combined is also 50 kHz when full interleaving exists. When L and R signals are not statistically independent or when (L+R) and (L−R) signals do not have matching pre-emphasis characteristics (as is the case when (L−R) is compressed in the BTSC system) the combined deviation of the main channel and stereophonic subchannel is also constrained to a maximum of 50 kHz peak deviation.

A CW pilot carrier of frequency $1f_H$ is transmitted with the main carrier deviation of 5 kHz.

The subcarrier for the Separate Audio Program (SAP) channel has a frequency of $5f_H$ (78.670 kHz) and is frequency locked to $5f_H$ in the absence of modulation. The SAP audio signal is subjected to compression identical to that of the L−R signal. The resulting SAP modulating signal is bandlimited to 10 kHz and frequency modulates the SAP subcarrier to a peak deviation of 10 kHz. The RF deviation or injection of this subcarrier is 15 kHz.

The professional subchannel has a subcarrier located at approximately $6.5f_H$ and modulates the aural carrier by 3 kHz peak deviation.

Total peak deviation allowed for the aural baseband is 73 kHz.

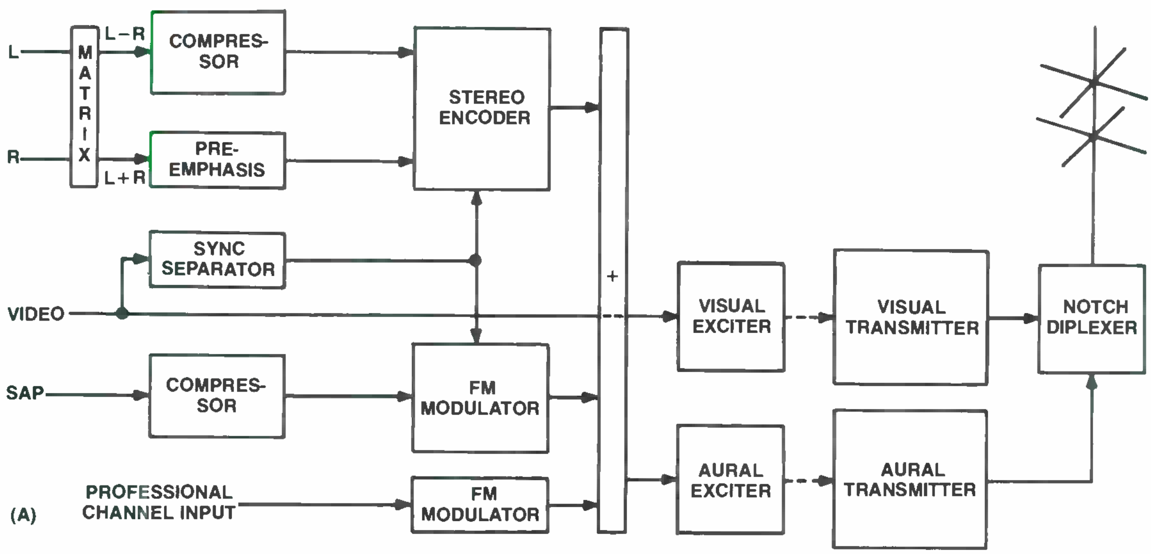


Fig. 30. Block diagram of BTSC system.

Stereo and SAP Noise and the Need for Companding

FM transmission systems like the BTSC system have a receiver parabolic noise characteristic, which means that they have increasing noise at higher frequencies. Because the stereo subchannel is at a higher frequency than the L + R signal (the L + R signal contains only frequencies up to 15 kHz, while the modulated subchannel bandwidth extends from about 17 kHz to 46 kHz), the L − R signal contains much more noise when demodulated.

To improve the signal to noise, the L − R modulation level can be increased, but this technique is limited. Increasing the modulation level too much will produce crosstalk from one channel to the other. In the BTSC system, 6 dB more L − R modulation is allowed than for L + R. The result is that the subcarrier adds approximately 15 dB of noise to stereo reception as compared with no companding, even under ideal reception condition.

To make matters worse, when transmission and/or reception conditions are impaired (transmitter ICPM incidental carrier phase modulation, multipath, etc.), buzz or hum can be introduced into the audio. This would further degrade the stereo signal-to-noise ratio as compared to mono if no companding system were used, especially in grade B reception areas. The SAP channel's subcarrier is located at a frequency of 78.67 kHz, so more noise is introduced because of the FM parabolic noise curve. Also the SAP subcarrier is an FM subcarrier which makes it additionally subject to buzz beat, an intermodulation of the picture with the audio that causes a particularly obnoxious type of non-harmonically related distortion.

Companding

Companding is defined as a process in which compression is followed by expansion. For noise reduction, the signal is compressed before the noise exposure (before the transmitter) and expansion afterwards (in the TV receiver). In the BTSC system, the dbx TV system of companding is used and will be explained in the paragraphs that follow.

The dbx TV noise-reduction system was designed to aid the BTSC multichannel sound TV transmission system in delivering a clean, noise-free audio signal into the home. Specifically the system was designed to:

- Provide significant amount of noise reduction even in poor reception areas (grade B or worse).
- Preserve input-signal dynamic range without headroom loss or transient overshoots.
- Prevent the stereo subcarrier from interfering with the overall transmitted power levels (AM-interleave effects).
- Reduce buzz due to ICPM.

The system was designed to compand only the L − R and the SAP channels. The L + R signal is transmitted without companding to ensure compatibility with the mono TV sets still in use. The same companding process is used for both the L − R and SAP channels.

Masking

All audio noise reduction systems work on the principles of masking. This principle can be simply stated as follows:

> *If a desired program signal (music or speech) is loud enough and broad enough in its spectral content, then the ear's attention will be captured by this signal rather than by the noise of the transmission medium.*

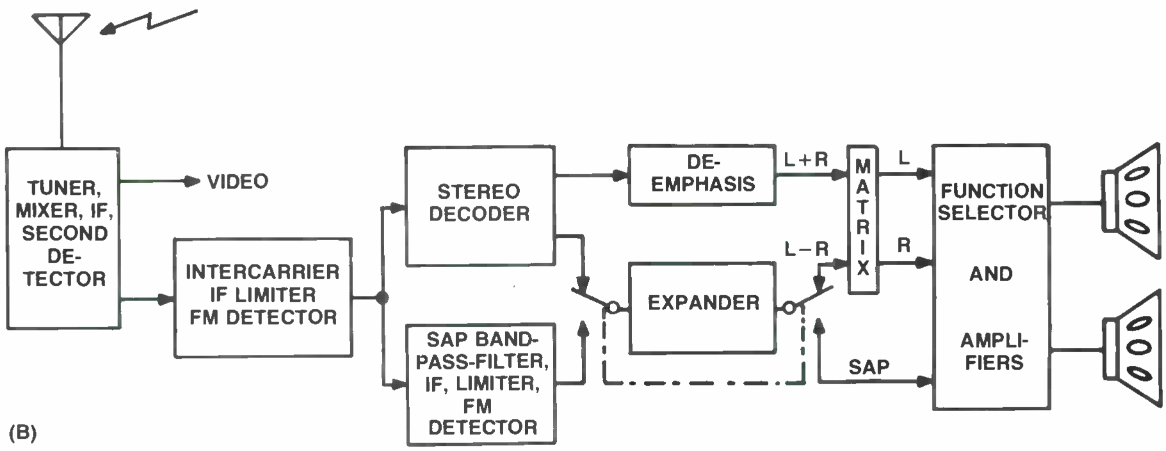


Fig. 30. (continued)

The noise-reduction system must encode (compress) the audio signal so that it will consistently mask the noise of the channel during transmission and then decode (expand) the transmitted signal to recover the original audio. And in the decoding process, all the audible noise should be eliminated. This requires that:

- The level of the transmitted audio is high relative to the background noise.
- The audio spectrum must be conducive to masking.

The background-noise spectrum of the BTSC stereo subcarrier is parabolic, having a 3-dB/octave rising characteristic, while the SAP subcarrier has a 9-dB/octave rising characteristic. Proper masking of the noise in the presence of signal takes place only if the transmitted signal spectrum contains substantial high frequency content, especially in the case of the SAP channel. If the program material itself could be relied upon to have sufficient high frequencies, then the compander would only have to keep amplitude levels high through the transmission channel. However, most program material has its dominant energy at low frequencies.

The simplest way to transform such program material into a signal conducive to masking is to apply fixed pre-emphasis to the signal before transmission. This pre-emphasis changes the spectrum of the average program material (containing mostly low frequencies) to be more evenly balanced between highs and lows. In the TV receiver, the program signal is de-emphasized, which restores correct tonal balance and reduces the audibility of the high frequency hiss at the same time.

In dbx TV, three processes are used to achieve the desired noise masking; pre-emphasis, spectral companding and wideband amplitude companding.

Pre-emphasis

In dbx TV two pre-emphasis networks are used. One is essentially the same as 75 microseconds (actually 72.7 microseconds). The other is 390 microseconds, but the rising frequency response created by this pre-emphasis is curtailed at 30 microseconds. The frequency response curve of the complete pre-emphasis has a very steep section between about 2 kHz and 5.5 kHz, which helps dbx TV overcome the large amounts of noise present in grade B reception. In the TV receiver, a corresponding de-emphasis restores the correct tonal balance to the program material and reduces the audibility of the hiss picked up in transmission.

Unfortunately, audio program material is inconsistent in its spectral balance and its level. With fixed pre-emphasis alone, two problems remain. First, some audio signals contain predominately high frequencies and the pre-emphasis would boost them too much, causing overmodulation. This is called lack of headroom—insufficient room for the peaks of the program to be transmitted cleanly. Second, some audio signals would be too low in level and too lacking in high frequencies to properly mask the channel noise, even with the strong pre-emphasis used in dbx TV.

Spectral Companding

In order to help make the audio more consistent in its spectral balance before transmission, a second stage, in which the pre-emphasis adapts its characteristics to suit the signal, is used and is called spectral companding. The spectral compressor (in the encoder) monitors the spectral balance of the input signal and varies the high frequency pre-emphasis accordingly. When very little high frequency information is present, the spectral compressor provides large high frequency pre-emphasis. When strong high frequencies are present, the spectral compressor actually provides de-emphasis, thereby reducing the potential for high frequency overload. The resulting encoded signal is therefore dynamically adjusted to consistently contain a substantial proportion of high frequencies before transmission, providing masking of the channel noise.

Spectral compression uses a variable pre-emphasis/de-emphasis stage and varies the gain of a voltage controlled amplifier (VCA) embedded in a frequency selective network. When the VCA gain is low, no signal appears at point B in Fig. 31. The only transfer from the input to the output must go from point C to the output, which passes through the de-emphasis network. This attenuates the high frequencies.

On the other hand, when the VCA gain is high, the response between point A and the output is essentially flat (like an op amp with a closed feedback loop). The signal at point C is small compared to the signal at point B (due to the high VCA gain), so the output signal is essentially the same as that at point A, which is the input signal after pre-emphasis. This provides gain to the high frequencies.

At intermediate VCA gain settings, the response varies smoothly between these two extremes, with flat occurring at unity (0-dB) gain.

The VCA gain is controlled by an rms level detector which senses the amount of high frequency energy at the output of the compressor. The bandpass filter center frequency is about 10 kHz.

When high levels of high-frequency energy are detected, the VCA gain is high, and de-emphasis

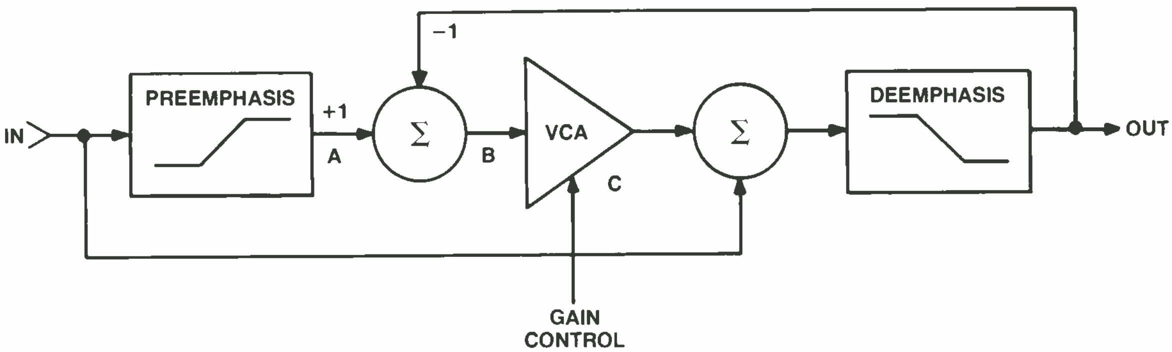


Fig. 31. Variable preemphasis/deemphasis network.

results, reducing high frequencies. When low levels of high-frequency energy are detected, the VCA gain is low, and pre-emphasis results, boosting high frequencies. In effect, the range of spectral variation at the output is reduced, "compressing" the spectral dynamic range.

During reception by a TV set, the spectral expander (in the decoder) will restore the high frequencies to their proper amplitude.

By using the spectral compressor, two simultaneous requirements are met:

- The system is extremely forgiving of high background noise environments because the spectral shaping of the input signal is adjusted according to the needs of the input signal to provide high masking at all times.
- Headroom is maintained throughout the frequency range because extreme pre-emphasis is used only when it is really needed.

Wideband Amplitude Companding

Neither the spectral compander nor the fixed pre-/de-emphasis will help reduce noise when the signal is very low in level, especially if the signal has little high frequency content. This is where the wideband compander is used to adjust the level of all frequencies simultaneously to keep the signal level in the transmission channel high at all times.

The wideband compressor reduces the dynamic range of input signals by a factor of 2:1 in dB. Not only are small signals raised in level but large signals are reduced. The output level tends toward about 14% modulation which has three benefits:

- The signal is consistently above the noise floor.
- The signal is consistently below 100% modulation, reducing AM interleaving effects.
- Headroom is maintained for transient peaks to overshoot the nominal level at the compressor output without causing overmodulation.

The wideband amplitude compressor works by controlling the gain of a VCA in response to an rms level detector, which senses the low and mid frequency energy level at the output of the compressor.

Transient Protection

A clipper or limiter is incorporated into the noise reduction compressor and is set to operate at 100% modulation to prevent large transients from causing excessive modulation of the transmission system.

To keep the effects of clipper operation from being audible, the amount of clipping is kept small. This is accomplished by:

- Setting the unaffected level point of the wideband compressor to substantially below the point that would cause 100% modulation in the transmission channel.
- Allowing the wideband and spectral compressors, which precede the clipper, to operate quickly enough to let only brief overloads reach the clipper.
- Placing the static pre-emphasis before the clipper, further reducing the transient overload duration.

Aural Transmitter Requirements

To ensure that the transmitter is not the limiting factor in stereo reproduction, it is desired that the transmitter be as transparent to the incoming signal as possible.

Baseband audio (stereo, SAP and profession channel) include components out to 105 kHz. Emphasis is placed on phase linearity, low distortion, minimization of any amplitude ripples, or roll-off over the audio passband to achieve optimum stereo separation and to minimize cross talk between the stereo channel and the SAP channel.

All errors in phase linearity or amplitude response within the aural transmitter tend to be cumulative. Separation degradation can be caused by amplitude roll-off or departure from phase linearity.

For stereo, modulated oscillator performance requires flat modulation sensitivity vs frequency characteristics up to 47 kHz (with a desirable flatness out to 120 kHz to allow for SAP and Pro channels).

Intermodulation and harmonic distortion products which, in monaural operation lie above 15 kHz, now lie in the stereo channel or the SAP channel and will degrade stereo separation or crosstalk into the SAP channel. In addition, IM products generated by stereo subchannel will now lie in the main channel.

All modern transmitters are IF modulated. To insure these low-level stages do not contribute any group delay or amplitude roll-off, wideband amplifiers should be used. As the signal is up converted to the RF aural carrier frequency, the FM of the local oscillator signal should be checked. The residual level of FM produced by the LO should be 10 dB better than the modulated oscillator noise. Synthesized LO sources should be checked for reference frequency spurs which may show up as FM noise.

All RF amplifiers should be viewed from the point of obtaining a flat symmetric amplitude response and a minimized group delay across the passband. Since FM modulation and demodulation is a nonlinear process there is not a one-to-one correspondence between RF amplitude/phase response and baseband stereo separation and crosstalk. Measurements indicate that a 2.5 MHz, 3 dB bandwidth is sufficient to achieve excellent stereo and SAP performance. No longer can amplifiers be tuned just by maximizing efficiency without compromising stereo performance or crosstalk, unless the stage is designed to operate that way. The amplifiers must be swept and frequency analyzed to insure the passband is sufficiently wide to pass the stereo signal and that its passband shape is symmetric about the carrier frequency.

Klystron amplifiers have previously been adjusted with all cavities, excluding the penultimate, synchronously tuned on the aural carrier. The cavities must be stagger-tuned to achieve the required 2 MHz 3 dB bandwidth.

The notch diplexer is possibly the most critical element in the aural chain with regard to stereo separation. If correct tuning of the notch diplexer itself is not properly maintained, stereo separation will be degraded and crosstalk between the mono, stereo and SAP channels will occur. Hybrid diplexers do not present any degradations to multichannel sound.

In summary, these major changes were incorporated and must be maintained in the aural portion of the transmitter:

- Wideband low distortion audio stages with high S/N are employed to insure negligible distortion in the baseband signal.
- Low distortion wideband modulated oscillators with improved noise performance and PLL techniques are used so that the quality of TV multichannel sound will approach that of the FM broadcast service.
- The bandwidth of IF and RF stages is wide. Power bandwidth tradeoffs may need to be made to achieve optimum stereo performance. Where feasible, the TV station might consider 20% aural power to increase its stereo coverage area. This may compensate partially for the degradation in S/N due to the increased bandwidth of the stereo signal.

Visual Transmitter Requirements

ICPM

There is no defined level of ICPM for a given stereo performance level since the signal to buzz ratio is highly dependent on the picture spectral components. Refer to the EIA Recommended Practices for the current recommendations on ICPM limits.

Video components at 4.5 MHz should be minimized prior to being injected into the transmitter.

Video sidebands translated to the intercarrier appear at nf_H in the audio passband. If the 4.5 MHz intercarrier is maintained precisely, the buzz caused by the ICPM transferred to the intercarrier will be minimized.

Overmodulation causes the intercarrier to disappear and thus the local oscillator in the TV receiver has nothing to which to lock. This produces buzz in the receiver. This may be prevented by the use of a luminance clipper.

Antenna reflections cause multipath effects in the receiver and a loss in stereo separation. Therefore, it is extremely important to minimize the aural VSWR across the passband.

Monitoring TV Multichannel Sound

Assurance of having a high quality aural signal is perhaps the most important reason for having a monitor. An aural monitor provides a means of maintaining constant loudness from various program sources. Periodic maintenance and proof of performance are other important reasons to have a monitor that demodulates the RF signal and separates the components in the composite signal for analysis.

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

- Demodulates the composite signal from the aural carrier.

- Separates the components in the composite signal for measurements.
- Capable of off-air monitoring.
- Covers all VHF and UHF channels.
- Suitable for professional performance use.
- Contains a precision expander.

UHF KLYSTRONS

The klystron is a device for amplifying signals at microwave radio frequencies. The high-velocity electron beam emitted from the cathode passes through the anode and into the RF interaction region, as shown in Fig. 32. An external magnetic field is employed to prevent the beam from spreading as it passes through the tube. At the other end of the tube, the electron beam impinges on the collector electrode, which dissipates the beam energy and returns the electron current to the beam power supply.

The RF interaction region, where the amplification occurs, contains resonant cavities and field-free drift spaces. The first resonant cavity encountered by an electron in the beam (the input cavity) is excited by the microwave signal to be amplified, and an alternative voltage of signal frequency is developed across the gap.

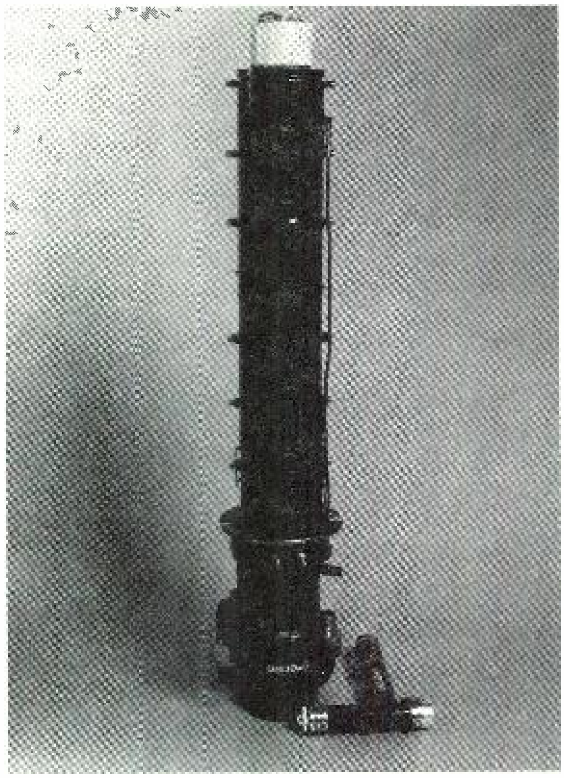


Fig. 33. Varian 110 kW VKP 7853 integral cavity klystron and output coupler.

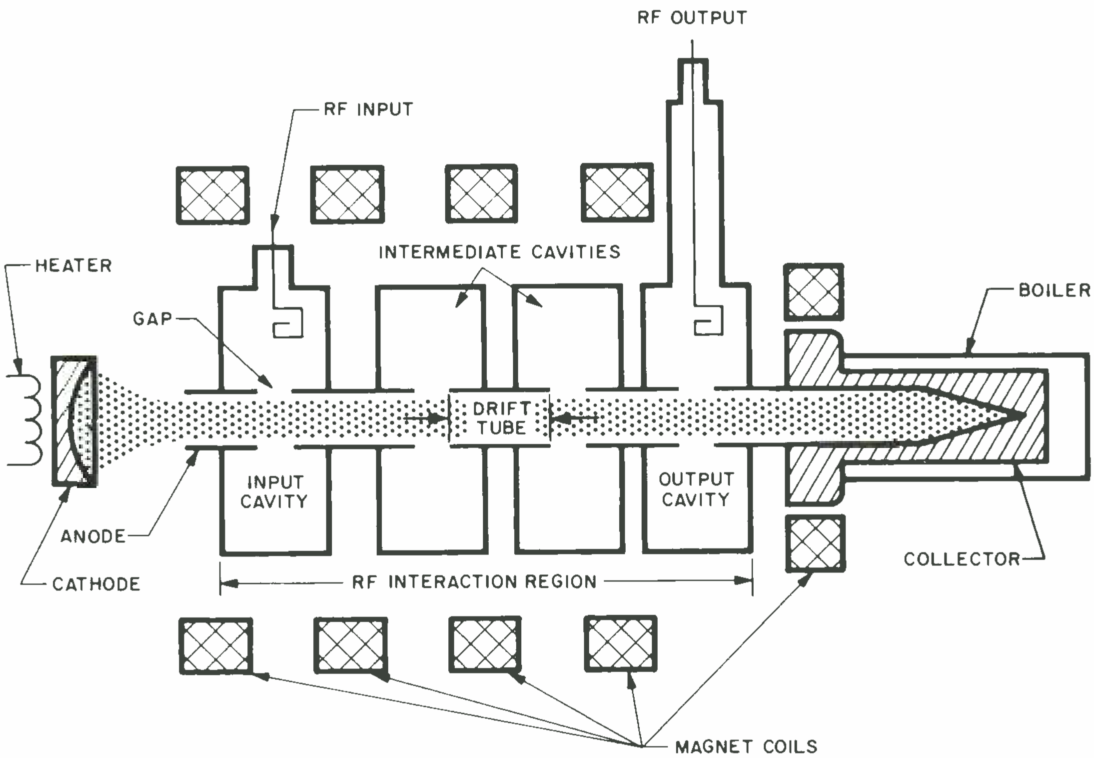


Fig. 32. Principal elements of a klystron.

Since electrons approach the input-cavity gap with equal velocities and emerge with different velocities, which are a function of the microwave signal, the electron beam is said to be *velocity modulated*. As the electrons travel down the drift tube, bunching develops, and thus the density of electrons passing a given point varies cyclically with time.

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

The cavities can be included in the vacuum envelope as shown in Fig. 33 or can be mounted external to the klystron as shown in Fig. 34.

All cathodes have optimum ranges of operating temperature. The operating temperature of the cathode must be high enough to prevent variations in the heater power from affecting the electron emission current (beam current) in the klystron. However, the temperature of the emitting surface must not be higher than necessary, since excessive temperature can shorten emission life. In Fig. 35, Ef_2 refers to filament voltage and Eb is the beam voltage.

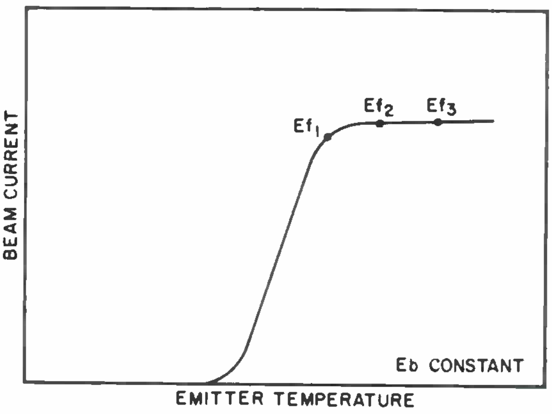


Fig. 35. Beam current variation with emitter temperature.

Perveance and the Mod Anode

Perveance is a function of the geometry of the cathode-anode structure. The modulating anode voltage controls beam current which can be calculated using the following equation:

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

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

Fig. 36 shows the relationship between beam current and voltage described in the above equation. Two examples for using the graph are given. In example A, if a modulating anode of 4000 volts with respect to the cathode beam voltage produces a beam current of 0.6 amperes, the in-

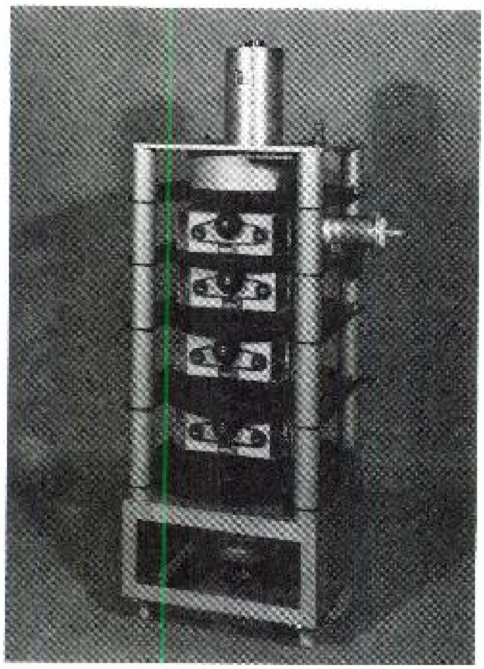


Fig. 34. EEV 58 kW K4251 external cavity klystron assembly.

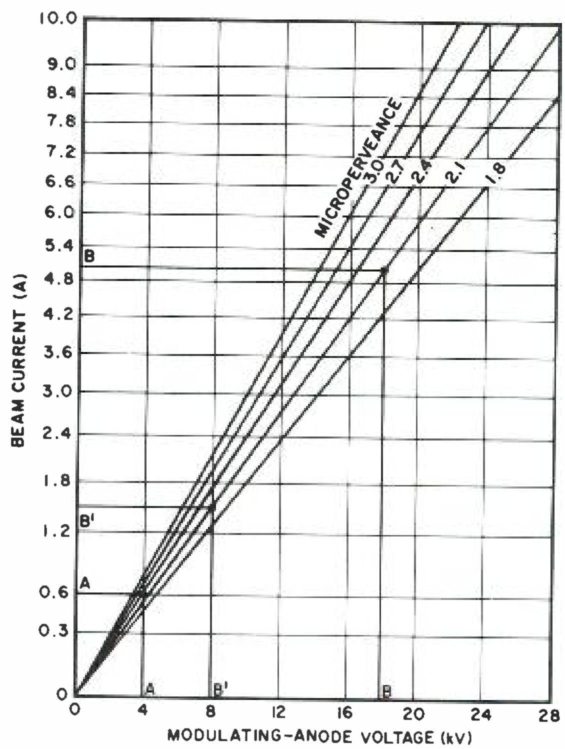


Fig. 36. Beam current variation with modulating-anode voltage.

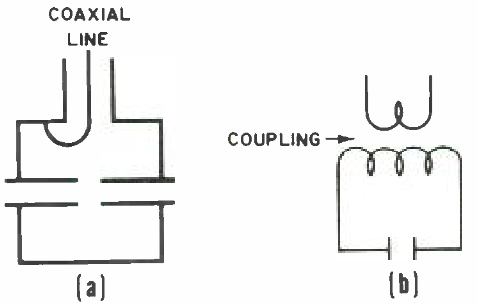


Fig. 37. Loop coupling and equivalent circuit.

tersection point lies on the 2.4 microperveance line, and the perveance is expressed as $2.4 \times 10^{-6} A/V^{3/2}$, or 2.4 micropervs. Operating condition B illustrates a rather practical television transmitter situation in which a common beam supply of 18 kilovolts is used to power both the visual and aural klystrons. At 18 kV, the visual tube operates at a beam current of approximately 5.0 amperes if the modulating anode is connected (through an isolating resistance) to the body of the tube, and the perveance is $2.1 \times 10^{-6} A/V^{3/2}$, or 2.1 micropervs. Since the aural output power required is much less, the dc input power can be reduced from that required to operate the visual tube. Points B′ indicate that if the modulating anode is supplied with only 8 kV (through a voltage divider) then the intersection with the 2.1 microperv line yields a beam current of only 1.5 amperes, thus accomplishing the necessary reduction of input power for aural service.

Magnetic Field

Electromagnetic coils are placed around the klystron to develop a magnetic field along the axis of the RF circuit which controls the size of the electron beam and keeps it aligned with the drift tubes.

Cavity Tuning

The resonant frequency of each of the cavities of a klystron can be adjusted to the operating frequency of the transmitter. This can be done in two ways:

- The inductance can be changed by changing the volume of the cavity. (External cavities)
- The capacitance of the drift-tube gaps can be changed. (Integral cavities)

Cavity/Transmission Line Coupling

Fig. 37 illustrates magnetic-loop coupling, where the RF energy is fed through a coaxial line with its center conductor inserted into the klystron cavity. The end of the center conductor is formed into a loop. This forms a simple one-turn transformer which couples RF energy into or out of the cavity through a coaxial transmission line.

Effect of RF Drive Power on RF Output Power

Fig. 38 shows RF output power as a function of RF drive power applied to the tube. From this curve, we see that when the RF drive power level is low, the RF output power is low. As the level of RF drive power increases, RF output power increases until an optimum point is reached. Beyond this point, further increases in RF drive power result in less RF output power. Because of these effects, two zones and one point have been labeled on the curve. In the zone labeled "Underdriven", RF output power increases when the RF input power is increased. The point labeled "Optimum" represents the maximum RF output power obtainable. Klystrons are said to be saturated at this point, since any further increase in RF drive only decreases the RF output power. The zone formed at the right side of saturation is labeled "Overdriven". To obtain maximum RF output power from a klystron, sufficient RF drive power must be applied to the tube to reach the point of saturation on the curve. Operating at RF drive levels beyond the saturation point will only overdrive the klystron, decrease RF output power, and increase the amount of beam interception at the drift tubes (body current). In TV service, klystrons are *always* operated within the underdriven zone of Fig. 38 unless beam current pulsing is used, which will be discussed later.

Fig. 39 shows how RF output power changes with various levels of RF drive power applied to a klystron under different tuning conditions. The RF power at Point A represents the drive saturation point for a synchronously-tuned tube. Point B shows a new point of saturation that is reached

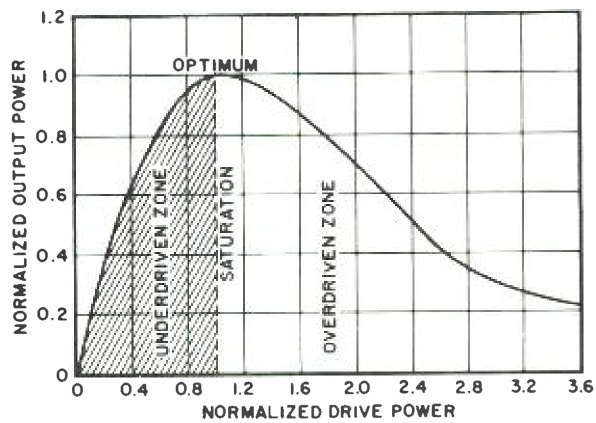


Fig. 38. RF output power as a function of RF drive power.

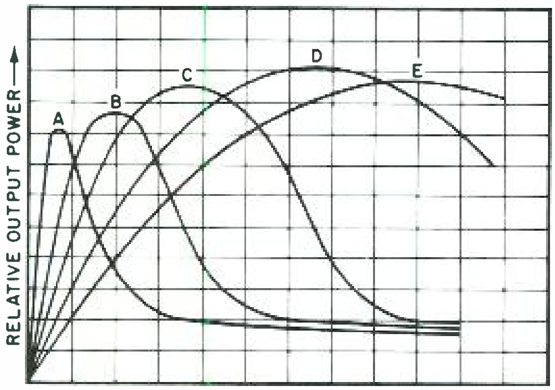


Fig. 39. Output power variation with drive power under different tuning conditions.

by tuning the penultimate (next to the last) cavity to a somewhat higher frequency. By tuning the penultimate cavity still further, Point C is reached. There is a point, Point D, where increasing the penultimate cavity frequency no longer increases RF output power; instead, it reduces the output power, Curve E.

Power Calibration

At UHF frequencies power is best measured by the calorimetric method. Dummy load water flow, input and output temperature can be accurately measured. Average power is then calculated:

P_{ave} = .264 × ΔT × GPM
ΔT = Change in water temperature from input to output of load in degrees C.
GPM = Gallons per minute of water flow through the load.

NOTE: For systems using special coolants other than pure water, an appropriate correction factor must be used.

Klystron Efficiency

Peak efficiency is calculated as follows:

dc input =
average beam current × beam voltage

Peak efficiency =

$$\frac{\textit{RF}\text{ power out (peak of sync for visual)}}{\textit{dc}\text{ input}}$$

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

Some typical efficiencies are illustrated in Fig. 40.

For NTSC visual service with video set at blanking (no setup), 75% modulation, the peak power equals 1.68 × average power, or P_{ave} = .595 × Peak.

Proper measurement of sync to blanking ratio is essential to establish an accurate calibration of power. *Note:* Attempting to operate a klystron at more than its rated power by only a few percent can result in a number of difficulties (ie. sync type oscillation, nonlinearity, etc.) as well as wasting electrical energy.

Tuning

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

- Synchronous tuning for maximum gain.
- High efficiency tuning for aural services.
- Broadband aural for stereo service.
- Visual service with fixed beam current.
- Visual service with pulsed mod anode voltage.
- High efficiency visual tuning for integral cavity klystrons employing the variable visual coupler.

It is best to consult the transmitter manufacturer for specific data regarding tuning with the accessories and mode of operation required.

External Cavity Klystron Tuning Considerations

Some additional adjustments are provided on external cavity klystrons. The input coupling loop is adjusted for minimum reflected power. Individual cavities may be externally loaded to lower Q and increase bandwidth.

Magnet coils for each cavity are sometimes controlled separately to optimize performance and minimize body current.

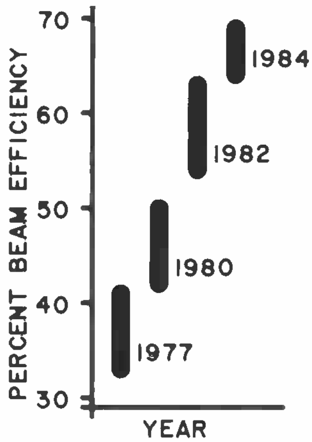


Fig. 40. Improvements of efficiency in UHF TV transmitters.

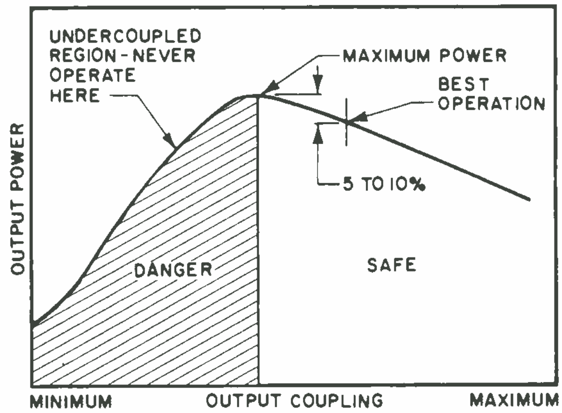


Fig. 41. Adjustment of output coupling control.

It is essential to operate with the output cavity slightly over coupled. Fig. 41 shows the relationship of output power to proper coupling loop adjustment.

Fig. 42 shows a typical response and cavity placement of an external cavity klystron tuned for visual service.

Mod Anode Pulsing

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

Operating the klystron at saturation will improve efficiency, but requires more linearity and phase compensation. By using a mod anode pulser to switch to a higher beam current during sync and back to a lesser current during video the average beam current is reduced. Refer to Fig. 43 block diagram of pulser.

Fig. 44 shows the horizontal line timing, modulation, and power output. Observe that sync is 8% of the duty cycle and line video is the remaining 92% of the transmitter signal.

Klystrons are high gain devices. In some applications as little as one watt is all that is required to drive a five-cavity klystron to 55 kW, a gain of 47 dB. Tuning for high efficiency and mod anode pulsing can reduce this gain by 13 dB or require 20 watts of drive.

Consider the following example of a typical 55 kW integral tube.

In static operation the tube is supplied with enough beam current to saturate at 115% power.

In pulsed operation only enough beam is supplied to saturate at 100% power during sync. The lower tip of burst causes 87.5% AM modulation.

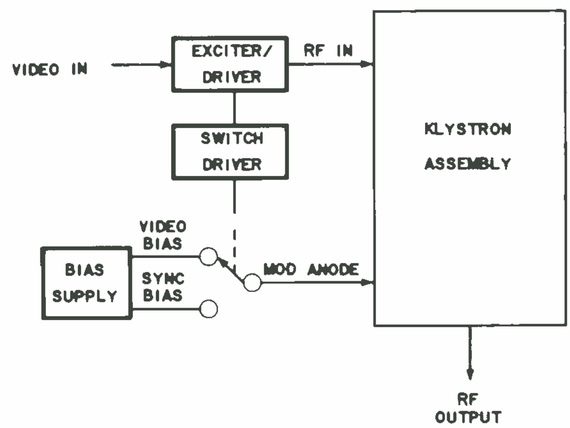


Fig. 43. Pulser block diagram.

Removing the lower sideband as in the case with vestigial filtering, the tip of burst represents 81.25% modulation. In order to prevent severe burst and back porch distortion approximately 6% modulation headroom is maintained. In other words the theoretical minimum beam current is increased to a practical level of about 87% corresponding to 76.6% of sync peak power. Refer to Fig. 45.

For Static Operation, Beam I is fixed with 24 kV 6.2A.

$$\text{dc Input} = 24 \times 6.2 = 148.8 \text{ kW}$$

$$\text{Peak Efficiency} = \frac{55}{148.8} = 37\%$$

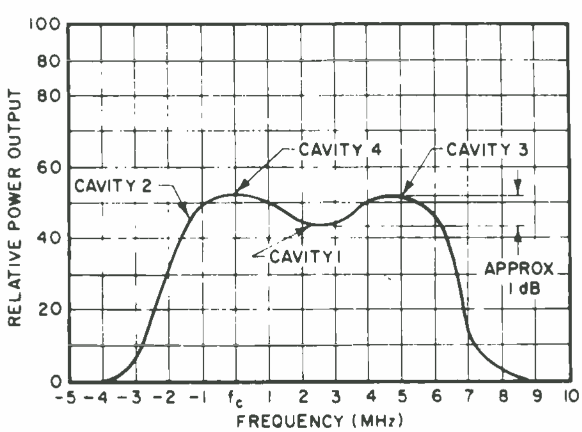


Fig. 42. Normal 4-cavity klystron tuning.

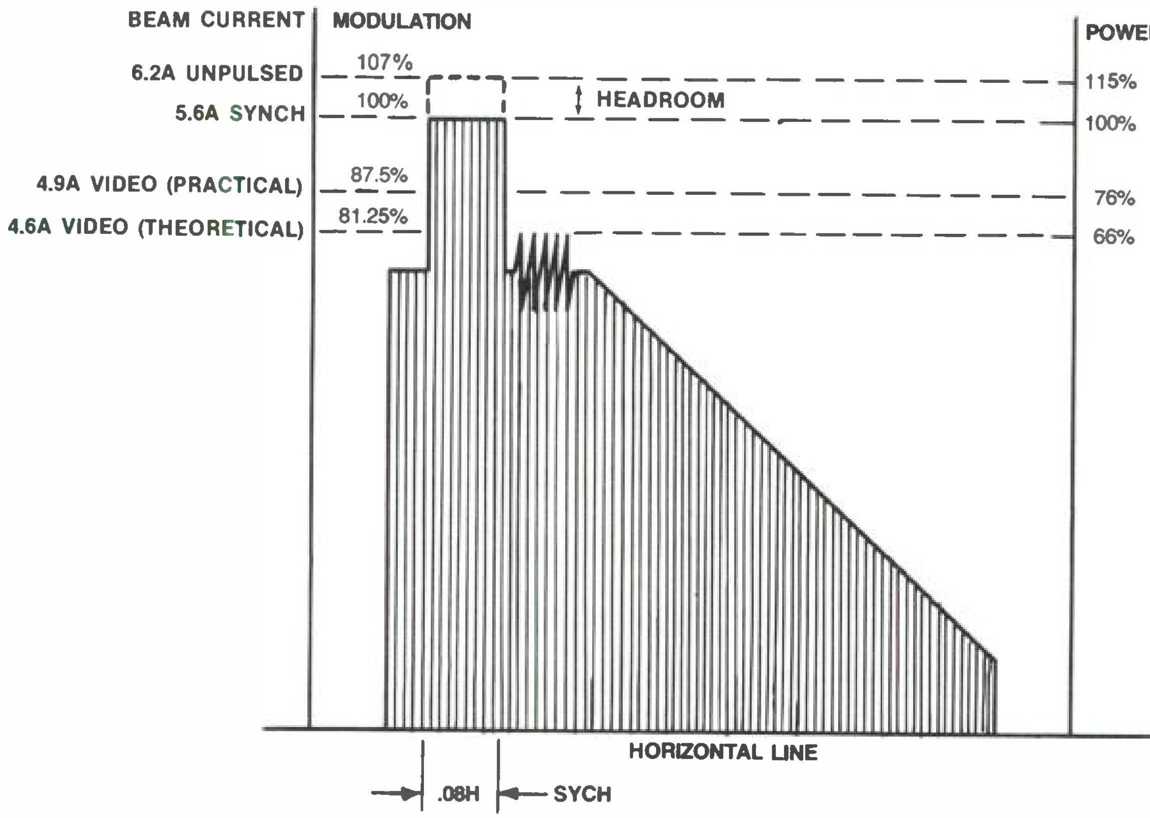


Fig. 44. RF envelope vs. beam current and power output.

For Pulsed Operation, DC Input is calculated as follows:

$$\text{DC Input} = \text{Beam Voltage} \times ((\text{Beam } I_{\text{sync}} \times \text{duty cycle}) + (\text{Beam } I_{\text{video}} \times \text{duty cycle})) = 24\ ((5.6 \times .08) + (4.9 \times .92)) = 24 \times 4.95 = 119 \text{ kW}$$

$$\text{Peak Efficiency} = \frac{55}{119} = 46\%$$

Both external and integral cavity klystrons have been successfully pulsed.

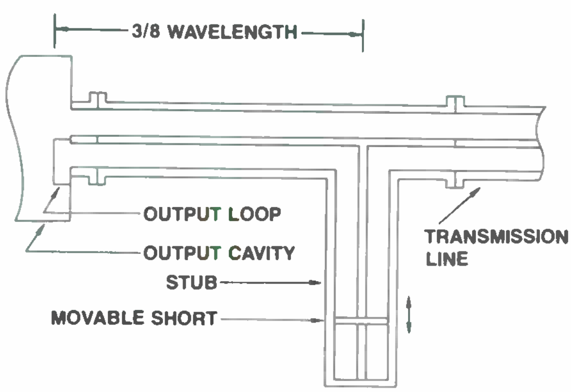


Fig. 45. Construction of Varian variable visual coupler.

Pulsing

During sync the klystron is operated at saturation which along with the change in beam current causes additional phase shift of the signal.

A phase modulation stage in the IF of the exciter is keyed by sync and adjusted to precorrect for incidental phase distortions caused in the klystron during pulsing.

Video signals are operating very close to saturation at burst and an additional range of differential gain and differential phase correction may be needed.

Modern exciters can fully precorrect these conditions. Refer to Figs. 48 and 49.

The practical limit of reduction of the video beam current is the point at which tip of burst and back porch modulation begins to be suppressed.

Variable Coupler for Integral Cavity Tubes

The variable load coupler has a shorted transmission line stub which is tuned beyond a quarter wavelength to present a capacitive susceptance load to the output transmission line. This capacitance plus the transmitter load reflects back through an approximate 3/8 wavelength to present a substantially resistive load to the output coupling loop of the final klystron cavity.

The tuning procedure calls for the shorted stub to be pulled out by a quarter turn. This presents a higher impedance to the output cavity and raises the cavity Q. The new saturated power is then found by varying the drive power while observing power output on the transmitter power meter and the detected output waveform from an envelope detector. If saturated power increases, the procedure is repeated.

Even though the visual coupler presents substantially a resistive load to the output cavity, Q changes and slight reactance effects require that passband response be slightly retuned every few steps.

This incremental procedure is repeated until the beam efficiency decreases or until a tendency for sync tip oscillations from return beam electrons is observed.

Sync pulse oscillations appear as ringing on the sync pulse and are believed to be caused by secondary electron feedback enhanced by the reverse gain of the klystron cavities.

The S Klystron in Visual

The VKP-7550S "S" klystrons attain 52 to 54% beam efficiency under CW or unpulsed TV conditions. The S klystron can be sync pulsed to produce 63 to 69% beam efficiency in a TV transmitter.

Achieving high efficiency with the VKP-7550S series klystrons requires up to 50 watts of linear drive power, large amplitude linearity correction and high levels of modulating anode pulsing.

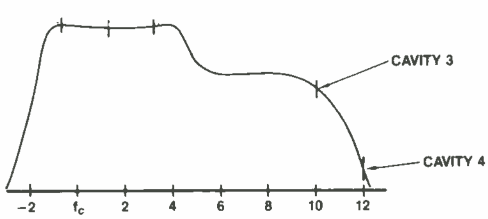
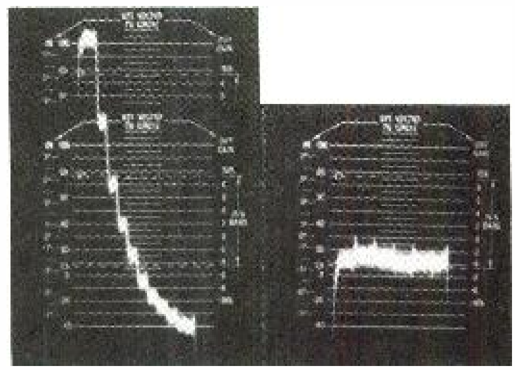


Fig. 46. High efficiency tuning of a 5-cavity Varian S series klystron.

Fig. 46 shows typical response and cavity placement of a five-cavity, Varian S series 60 kW klystron tuned for high efficiency. Figs. 47, 48, and 49 are actual visual performance waveforms showing large exciter precorrections necessary to achieve transparent output characteristics while achieving high efficiency.

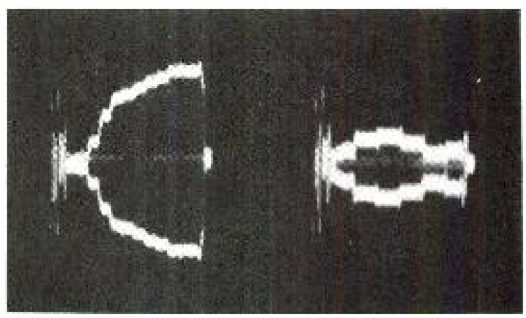
DIFFERENTIAL GAIN AT 70% EFFICIENCY, 60.6 kW



PRE-CORRECTED OUTPUT SIGNAL

Fig. 47. Differential gain at 69% efficiency, 60 kW.

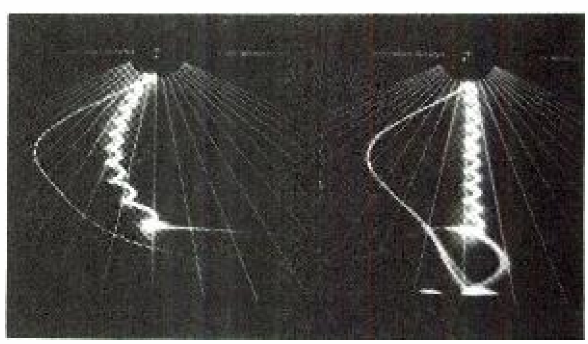
DIFFERENTIAL PHASE AT 70% EFFICIENCY, 60.6 kW



PRE-CORRECTED OUTPUT SIGNAL

Fig. 48. Differential phase at 69% efficiency, 60 kW.

INCIDENTAL PHASE AT 70% EFFICIENCY, 60.6 kW



PRE-CORRECTED OUTPUT SIGNAL

Fig. 49. Incidental phase at 69% efficiency, 60 kW.

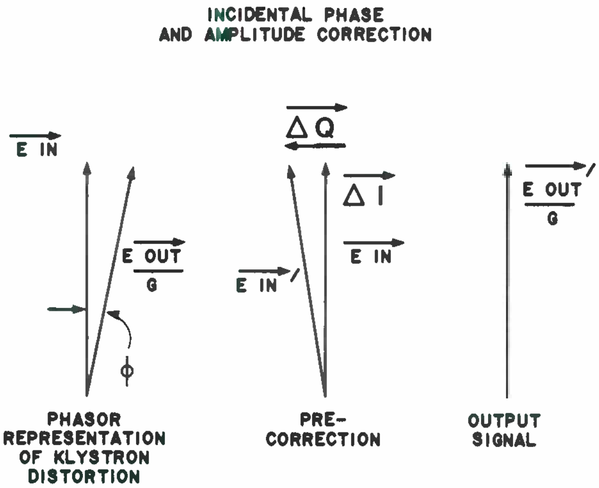


Fig. 50. Incidental phase and amplitude correction.

Tuning is as follows:

1st cavity	at −1 MHz,
2nd cavity	at +2 MHz,
3rd cavity	at +10 MHz,
4th cavity	at +12 MHz, and
5th cavity	at visual carrier.

Precorrected Signal driving klystron shows 36% differential gain. Output Signal shows 3% differential gain. Vertical scale is 2%/div.

Precorrected signal driving klystron shows 13.5° differential phase. Output signal shows 1.5% differential phase. Vertical scale is 1°/div.

Precorrected signal driving klystron shows −6° over video and +20° at sync.

Output signal shows +2° incidental phase.

Incidental Phase and Linearity Correction

With increasing RF drive level, klystron amplifiers exhibit a phase change in addition to an amplitude compression. This phase change is called incidental phase modulation (ICPM). Very little can be done to reduce ICPM by klystron tuning, selection of magnet current, etc. Unfortunately, incidental phase distortion increases rapidly near saturation of the klystron. For maximum efficiency, one must operate the tube at saturation at sync tip. In a pulsed transmitter, the color burst and black picture content are near saturation as well.

To combat this problem, ICPM correctors have been developed. These correctors generally operate at the exciter intermediate frequency and introduce a correction equal and opposite to the distortion produced by the klystron.

Fig. 50 illustrates first order klystron nonlinearities and the operation of the quadrature and linearity correctors. The desired TV signal is represented by an instantaneous phasor diagram as E_{in}. The output of the transmitter would ideally be an amplified version of E_{in}.

An instantaneous phasor diagram illustrates how first order klystron non-linearities are corrected by ΔQ, and an inphase linearity correction, ΔI.

However, the transmitter output E_{out} (divided by the nominal RF gain G is phase shifted by a phase error ϕ and compressed in amplitude. The signal in the exciter is precorrected to E_{in} by an amplitude expansion ΔI and a correction in quadrature ΔQ. When the resultant signal E_{in} is amplified, the output signal E_{out} will be replica of the desired TV signal.

TRIODES, TETRODES, AND TELEVISION TRANSMITTERS

Triodes and tetrodes are used in the RF power stages of television transmitters. High power levels for VHF television transmitters are usually obtained by a single ended vacuum tube.

Much written material about the vacuum tube is still available but most of it emphasizes receiving tube theory. This section will highlight some of the facts on high power gridded transmitter tubes. The parts of the power tube will be reviewed.

General

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

A Triode is a three element tube consisting of the following:

- A heated cathode which emits electrons.
- A control grid which modulats the electron stream in accordance with the dc and ac voltages impresssd between the control grid and cathode.
- A plate which accepts the electron stream. The plate is positive with respect to the cathode.
- Input signals, dc and ac are applied between control grid and cathode.
- Output signals, dc and ac flow to the tube output load impedance by way of the plate and cathode.

A tetrode is a four element tube which has a screen grid added between the control grid and plates.

- The screen grid acts as an electrostatic shield which helps to isolate the plate output signal from the control grid input signal.
- The screen grid is operated at a positive dc potential with respect to the cathode. This potential is much lower than the dc plate voltage.
- The screen grid is usually operated at ac (RF) ground potential.
- The tetrode has higher gain than the triode.
- The plate voltage has minimum effect on the plate current as long as the instantaneous plate voltage is greater than the screen grid dc voltage.
    1. The plate voltage is normally allowed to swing close to but not below the value of dc screen grid voltage.
- If the tetrode screen grid is not maintained at ac (RF) ground potential the tube can take on the characteristics of a triode.

The tetrode input bias and signal are applied between the control grid and cathode and are similar in performance to the triode.

The dc plate current flows from cathode to plate.

The output ac (RF) signal current path in the tetrode is *from the plate, through the load impedance, and returns through the screen grid.*

The Cathode

A power grid tube will have high peak and average current levels. Therefore, the cathode must emit many electrons. It must be hot and requires large heater power with low voltage and high currents. DC current is often used to reduce ac noise levels.

It is usually directly heated to provide quicker warm up with lower heater power. It will requires some type of cooling to remove excess cathode heat from the tube filament/cathode contacts and the tube socket.

The Grids

The voltages (dc, ac, and RF) of all grids are measured with respect to the cathode.

The grids can have current which are positive when the grid accepts electrons from the electron stream or negative when the grid emits electrons.

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

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

The control grid has a negative dc bias and an ac signal which drives the grid positive.

The instantaneous grid current can be negative when the grid is swinging maximum negative, zero when the grid is only slightly negative, or positive when the grid is positive.

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

Both grids have instantaneous positive, zero, or negative currents that follow the voltage swing. The average of these currents is the dc grid current.

Since both dc voltage and current are present in both grids, the grids can dissipate power. The control grid also has ac (RF) voltages and currents present so that ac power dissipation in grids must be observed.

The dc plate power, E_b and I_b, that is applied to the tube is either converted to ac (RF) output power or dissipated as heat. This heat is developed in the plate and must be removed by air or water cooling.

Below 30 MHz, RF amplifiers use lumped components to resonate the tube input and outputs. At VHF and UHF frequencies several problems make the use of lumped components impractical.

The Transmission Line Cavity

As frequency increases, lumped component resonant circuits get smaller and smaller to reduce L and C, larger in diameter to reduce skin effect, closer to the tube to reduce the effects of stray L, and there is great difficulty in predicting exactly what values of R, L, and C a component or circuit may have.

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

One solution to the above problems is the resonant transmission line cavity amplifier. In this type of amplifier the tube becomes part of a resonant transmission line. The elements of these tubes are arranged to look like concentric coax-

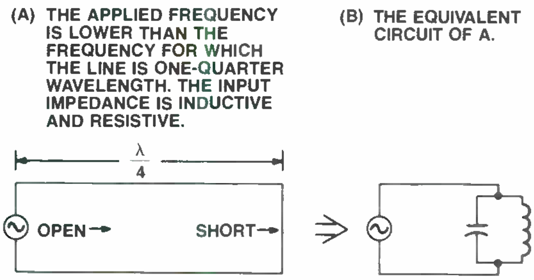


Fig. 51 Shorted quarter wavelength line.

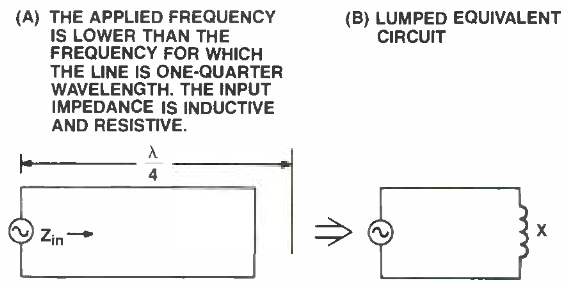


Fig. 52. Shorted line less than one-quarter wavelength.

ial transmission lines. The design of these tubes stresses low interelectrode capacity and low-distributed inductance. The stray interelectrode and distributed capacity and inductance of the tube becomes part of the resonant transmission line. The resonant transmission line is physically larger than the equivalent lumped constant $L-C$ resonant circuit operating in the same frequency. This larger physical size aids in solving the high power operation problems of skin effect losses, prevention of arcs and shorts, and reliable and predictable operation. A commonly used transmission line cavity amplifier uses a quarter wavelength transmission line as its resonant element.

The Shortened Quarter Wavelength Line

A shorted quarter wavelength transmission line has a high (almost open), purely resistive input impedance. Electrically it looks like a parallel resonant circuit (refer to Fig. 51). If the applied frequency is changed slightly so that the shorted line is no longer one quarter wavelength long, the input impedance drops and no longer remains purely resistive.

The Shortened Transmission Line less than a Quarter Wavelength Long

When operated at a frequency below that for which the shorted line is one quarter wavelength long, the physical length of the line at the new lower frequency will be less than one quarter wavelength. The impedance will be lower and the line will look inductive. Refer to Fig. 52.

The Shortened Quarter Wavelength Cavity

In Fig. 53, shorted transmission lines are used to resonate the inputs and outputs of this amplifier. Notice that the length of the lines are less than a quarter wave length but the tubes shunt input and output capacity and its series lead inductance will electrically lengthen and resonate the transmission lines. The input is shown inductively coupled, but it could just as easily have been capacitively coupled to the cathode. The input could also have a lumped constant resonant circuit or a transmission line resonant circuit since its power level is low. The output coupling is capacitive, but it also could have been inductive.

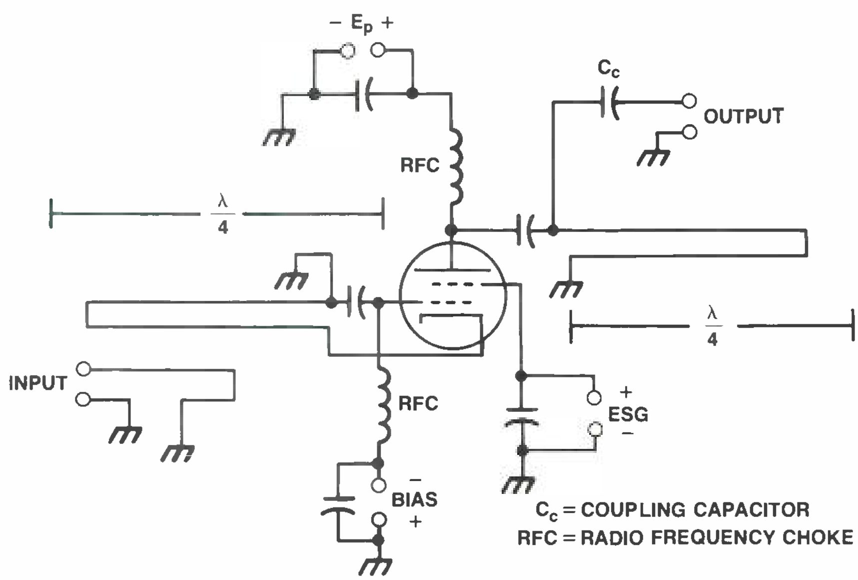


Fig. 53. A shorted quarter wavelength transmission line amplifier.

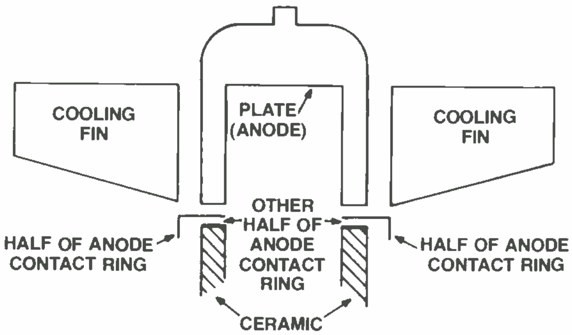


Fig. 54. Cutaway view of the anode structure.

The construction of the tube lends itself to transmission line tuning techniques.

Coaxial Line Feature of a Tetrode RF Power Amplifier Tube

The Anode (Plate)

The plate resembles a copper cup with the plate contact ring welded to the mouth and the cooling fins silver soldered or welded to the outside of the cup (refer to Fig. 54 and Fig. 55).

The contact ring is bonded to the base ceramic spacer through a strain isolation ring. This ceramic spacer is the same ceramic that is shown above the screen contact ring in Fig. 56.

The Screen Structure

The screen grid consists of many vertical supports fastened to a metal base cone. The other end of the metal base cone fastens to the screen contact ring. The inductance of the individual vertical supports is reduced by building the screen grid of many of them in parallel. The vertical supports are held rigid by horizontal rings welded to them and a metal cap on the top of the assembly. The screen contact ring, metal base, and metal base cone also functions to reduce lead inductance and RF resistance due to skin effect (refer to Fig. 56).

A cutaway view of the plate circuit and the screen circuit in Fig. 57 shows a concentric construction that resembles a coaxial transmission line.

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

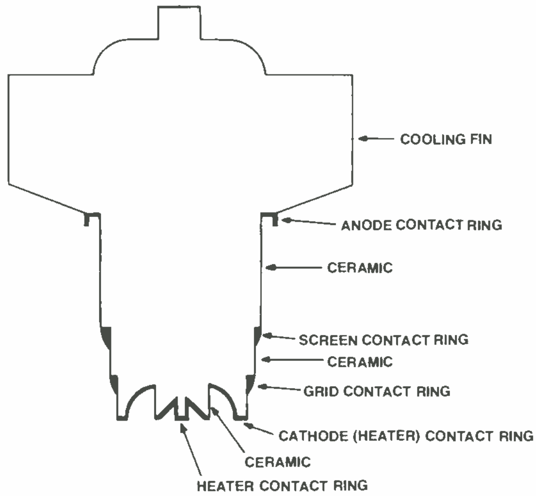


Fig. 55. Cutaway view of the exterior of a RF tetrode.

screen contact ring, up the cone, and up the the screen bypass capacitor, then through the screen grid to return to the imaginary generator. The screen grid has RF current returning to it but due to its low impedance, the screen grid is at RF ground potential. The RF current generator appears to be feeding an open ended transmission line consisting of the anode (plate) assembly, and the screen assembly. The RF voltage developed by the anode is due to the plate impedance (*Zp*) presented to the anode by the resonant circuit and its load.

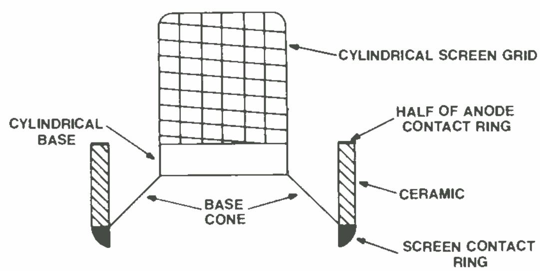


Fig. 56. The screen grid assembly.

The control grid assembly and the cathode assembly are also cylindrically constructed and concentric. The control grid assembly is constructed similarly to the screen grid and is slightly smaller than it. Fig. 58 shows the screen grid, control grid, and the cathode assemblies as they are placed in the tube.

Fig. 58 also shows the current path of an RF generator feeding a signal into the grid/cathode assembly resembles a transmission line terminated by the RF resistance of the tube's electron stream. The outer contact ring for the cathode heater is the inner conductor of this transmission line.

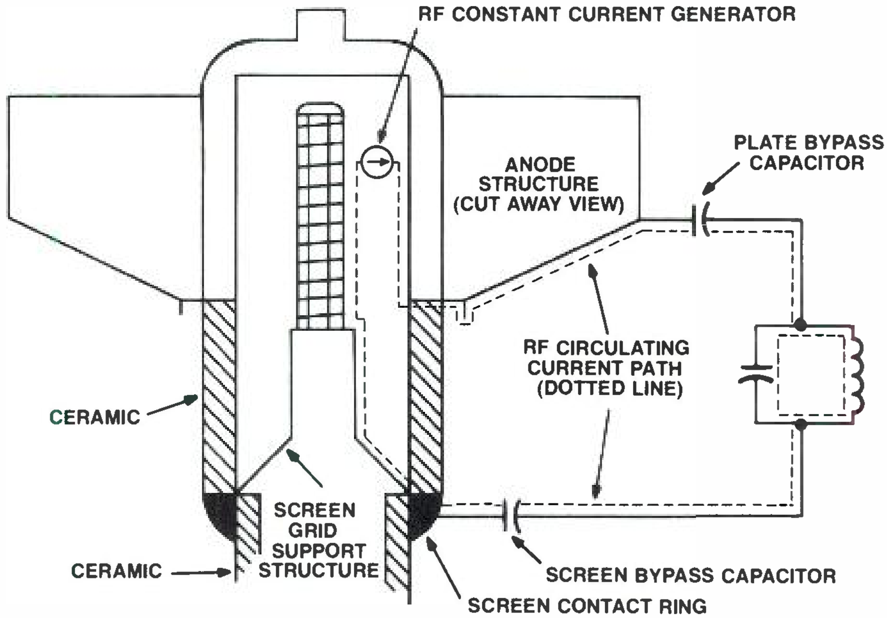


Fig. 57. Showing the plate and screen assembly and RF circulating current path (dotted line).

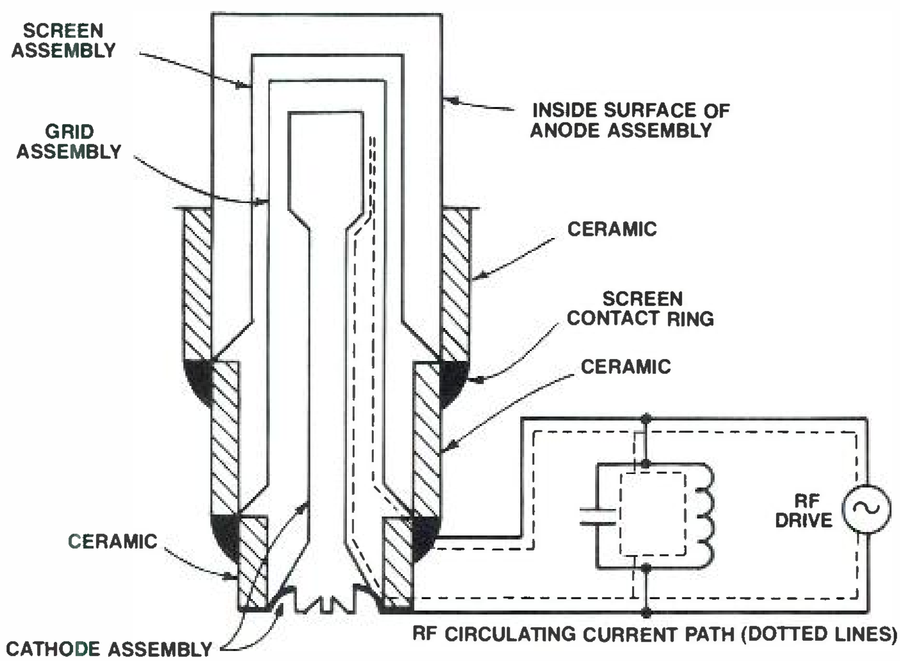


Fig. 58. Showing details of assembly of grids and cathode components and the simplified RF input circuit (bias circuit not shown).

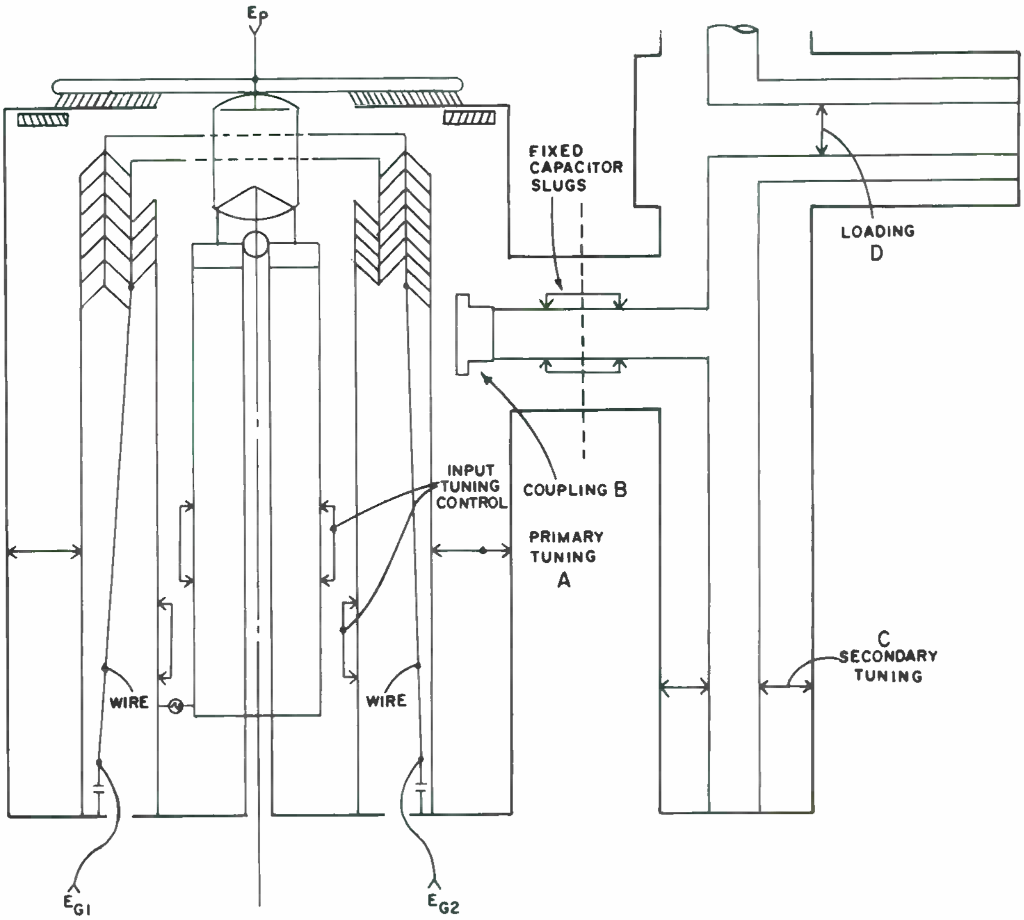


Fig. 59. High band coaxial transmission line cavity amplifier for TV use.

Double-Tuned Tube-Type Power Amplifiers Used in TV Transmitter Applications

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

As an example, the double-tuned overcoupled amplifier shown in Fig. 60 should not be thought of as an amplifier and a socket into which a tube is placed. The tube is an integral part of the cavity. The internal electrical properties of the tube will determine the amplifier gain and power han-

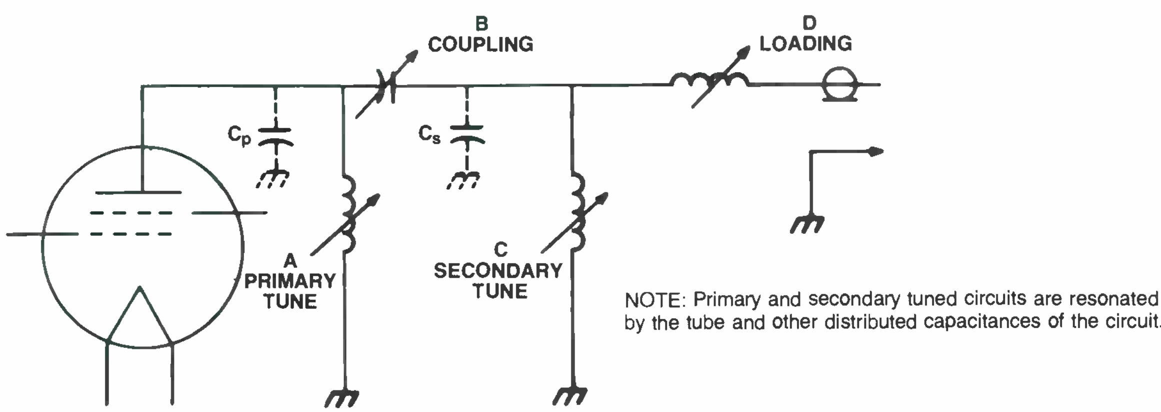
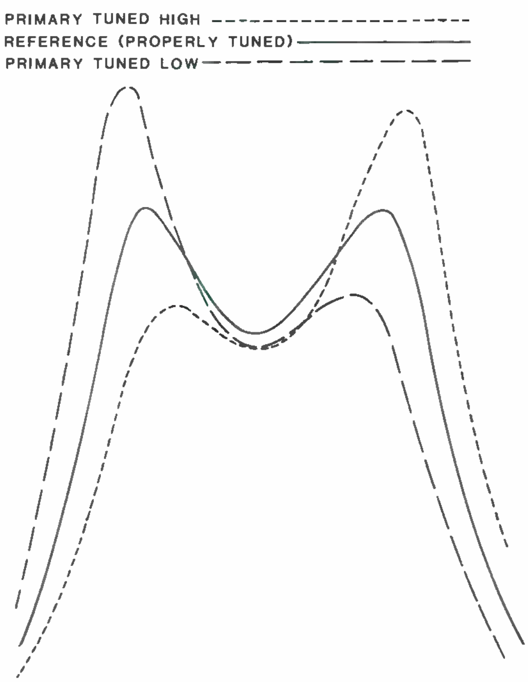


Fig. 60. Double-tuned power amplifier.



NOTE: As the primary tuning is rocked the response tilts and shifts.

Fig. 61. Primary tuning.

dling capabilities. It will also dictate the dimensions and functions of the circuitry.

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

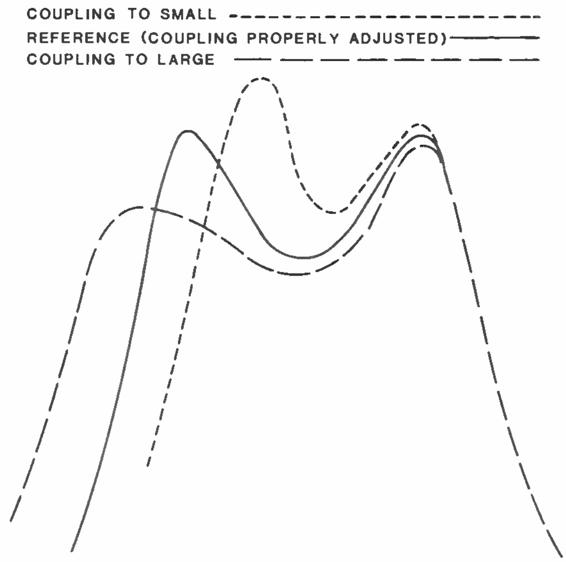
Effects of Tuning an Overcoupled P.A.

Plate Tune (Primary tune): Resonates the plate circuit and tends to tilt the response and slide it up and down the bandpass (refer to *A* on Fig. 60, also Fig. 61 and Fig. 59).

Coupling: Sets the bandwidth of the PA. Increased coupling increases bandwidth and lowers the PA efficiency. When the coupling is adjusted, it can tilt the response and change the center of the bandpass necessitating the readjustment of the plate tune control (refer to *B* on Fig. 60, also Fig. 62 and Fig. 59).

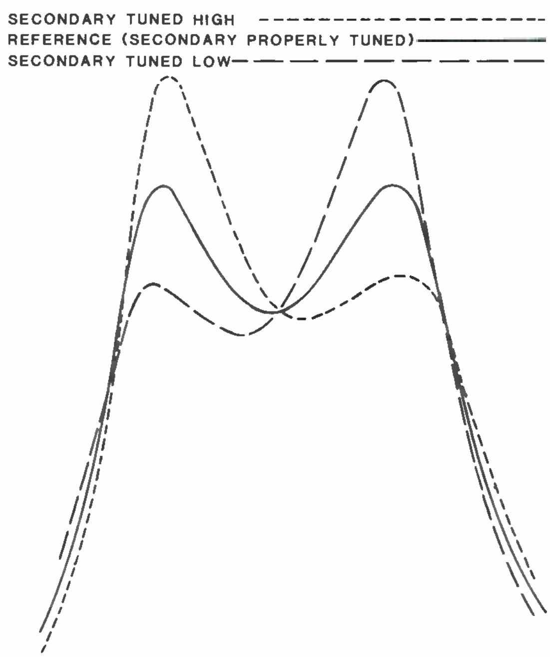
Secondary Tune: Resonates the secondary cavity and will tilt the response if adjusted. Generally it will not slide the response up and down the bandpass as will the primary tune (refer to *C* on Fig. 60, also Fig 63 and Fig. 59).

Loading (Secondary load or output load): This control determines the depth of saddle or value of ripple in the response. Heavier loading (maximum *C* or minimum *L*) creates a haystacked response and light loading creates excessive ripple. Adjustment of the loading control usually tilts the response and changes bandwidth. This necessitates readjustment of secondary tune and



NOTE: When coupling is adjusted, the upper end of the response remains constant and the upper end of the bandpass moves thus changing the bandwidth. Also the response is tilted and the center of the bandpass is shifted. This makes necessary the adjustment of the primary and secondary tuning to center and flatten the bandpass.

Fig. 62. Coupling plate loading.



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

Fig. 63. Secondary tuning.

coupling. In some cavities, primary tune may also have to be readjusted. The coupling control will also effect the value of ripple, but its greatest effect will be on the amplifiers bandwidth (refer to *D* on Fig. 60, also Fig. 64 and Fig. 59).

NOTE: As loading is decreased, the amount of ripple is increased (greater depth of saddle) along with a tilt of response. The response tilt can be corrected by adjusting the secondary tuning. Likewise, as loading is increased, the value of ripple decreases. Again the response tilt is corrected by adjustment of secondary tuning.

When the loading control is changed and the response is corrected by the adjustment of coupling, secondary tune, and primary tune, the plate impedance of the amplifier is changed.

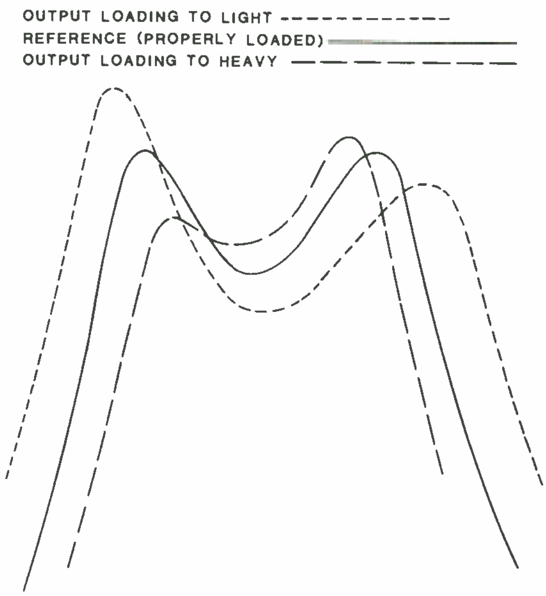


Fig. 64. Loading adjustments.

Heavy Loading	Smaller Ripple	Lower Plate Impedance	Higher I_p and Positive I_g Lower Gain	Lower Positive I_{sg}
Light Loading	Larger Ripple (Saddle) (Saddle)	Higher Plate Impedance	Lower Ip and Positive Ig Higher Gain	Higher Positive I_{sg}

Tuning for Power

Sweep the entire transmitter into the station load at 100% power. If tuning of any part or all of the transmitter is in doubt, check the following. Watch the response, reflected power (VSWR), I_p, I_{sg}, and I_g. If total response is not right, check each stage, starting at the exciter, to be sure each stage is flat and has the correct bandpass. The final amplifier should be the narrowest of all stages.

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

1. Check for proper sync level at the transmitter output and adjust at the visual exciter linearity corrector as necessary.
2. Watch for excessive plate dissipation. Plate dissipation $= (E_p) \times (I_p) -$ (average rf power out)

   Where: E_p = dc Plate Voltage
   I_p = dc Plate Current
3. Check for excessive screen current. The screen current will increase rapidly at higher power if the PA is lightly loaded. Very light loading can cause sync compression.
4. Watch for excessive grid current if PA is loaded too heavy. It can cause sync compression.
5. Check the amplifiers output efficiency. It can also be a guide to proper operation of the stage. The efficiency at VHF typically will range from 41% to 45%. At UHF the range is somewhat lower.

$$\text{Efficiency} = \frac{P_o}{E_p I_p}$$

Where: P_o = Average RF output power

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

Fig. 65. Nonsymmetrical response caused by the amplifiers outside the transmitters bandpass. Solid line is transmitter overall response. Dotted line is the pat of PA response outside of the transmitters bandpass.

6. It is assumed that the RF output power is measured by an accurately calibrated rf wattmeter or by a calorimeter. If the wattmeter is not accurate, the amplifier dissipation and efficiency may be in doubt and the tube life will be shortened.

If the wattmeter reads low, 100% output power will be difficult to obtain. The amplifier might be trying to produce more than its full rated power. The symptoms would be high I_p, high dissipation, and sync compression. These are the same symptoms that improper tuning could yield.

If the wattmeter reads high, power will be easy to obtain and efficiency will appear high.

Plate dissipation might appear low and in reality be high. This will become apparent by a blackened anode.

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

The power measured on a calorimeter or on the wattmeter is average power. To find peak power use this formula:

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

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

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

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

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

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

1. Sweep the transmitter and decrease the PA loading.
2. This will cause the response to tilt. Correct this tilt by adjusting the secondary tuning.
3. Decreasing the loading will also cause the bandwidth (and saddle) to increase. This can be counteracted by decreasing coupling.
4. Changing coupling will cause the response to tilt. This can be corrected by adjusting the primary and/or secondary tuning.

The above procedure may have caused the PA response to slide up or down out of the bandpass. It will show up as a nonsymmetrical bandpass (See Fig. 65). It can be corrected by adjusting both primary and secondary tune simultaneously to center the response.

5. The transmitters overall response will have the same bandpass but will have slightly more ripple. The ripple content in the transmitters overall response should still be within the ¼ to 1¼ dB limits.

CAUTION: When changing loading, coupling, and primary and secondary tuning, it is possible to get the PA bandwidth too wide. The excessive bandwidth of the PA will be masked by driver narrow response. This excessive bandwidth will cause PA overdissipation. Refer to Fig. 67. Correct bandwidth is shown in Fig. 66.

Once again transmit a black picture into the station load. Switch the Vestigial Sideband Filter and the linearity corrector in and check sync level at the transmitters output. I_p and I_g should be lower and I_{sg} should be higher. The amplifier efficiency should fall between 41% to 45%.

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

External Air Systems for Transmitter Sites

Most electronic equipment which requires forced air cooling has the required blower or fan already designed into it. However, with equipment such as transmitters dissipating relatively large amounts of power, the air which has already passed through the equipment and has picked up heat must be removed from the immediate vicinity of the equipment. In order to prevent the hot air from being recirculated and thus providing a hotter environment than that which the equipment was designed for. In addition to the air removal requirement, provisions must be made for sufficient makeup air to replace that which has been circulated through the equipment and removed.

If the equipment user is to provide adequately for hot air removal and fresh air makeup, the following information from the equipment manufacturer must be known:

- Maximum and minimum environmental conditions which the equipment may operate in.

  Altitude: __________
  Max Temp: __________
  Min Temp: __________

- Total air through the transmitter

  CFM __________

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

  P (Inches H_2O) __________

- Air temperature rise through the transmitter: __________
- Air exhaust area: __________

The following facts must be known about the conditions at the site or transmitter location:

- Maximum temperature: __________ (In the transmitter room.)
- Altitude above sea level: __________

First, the removal of heated exhaust air away from the transmitter will be dealt with. Equipment layouts usually provide for heated air to exit from the top surface of the cabinet. The size and location of this exhaust area is usually shown on a manufacturer supplied outline drawing. Most broadcast equipment internal air systems are designed to be operated into free space (0 inches of water back pressure) so any unpowered exhaust must have minimum loss. A good practice usually is no more than +0.1 inches H_2O pressure in the duct close to the exhaust area of the transmitter.

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

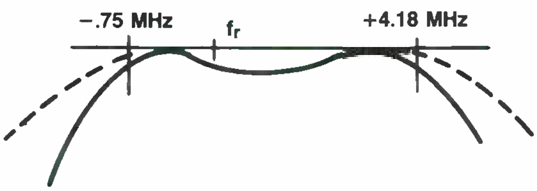
The following is an example of an exhaust duct design illustrating key cooling concepts:

From Manufacturers Specifications:

Air flow through transmitter = 325 CFM

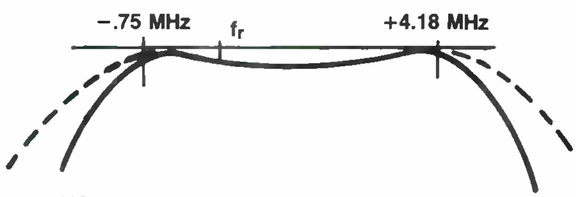
Air exhaust area = 3.4 ft²

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



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

Fig. 66. Proper bandwidth.



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

Fig. 67. Improper PA bandwidth.

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

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

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

$$V = \frac{325 \text{ CFM}}{.545 \text{ ft}^2} = \text{596 ft/min air velocity in a 10-inch diameter pipe.}$$

Where: D = Duct Diameter
A = Exhaust Duct Cross Sectional Area
V = Exhaust Duct Air Velocity

In a 10-inch diameter pipe air friction chart gives .06 drop per 100 feet of pipe with a flow of 325 CFM. Refer to friction loss in pipes chart Fig. 68. Assuming there is 20 feet of straight pipe to a roof, with two 90 elbows to turn the pipe down for weather protection, the total loss of the exhaust system may be estimated as follows:

Figure 71 shows loss is .06 inch/100 square feet.
Since 20 feet is .2 of 100 feet,
.2 × .06 = .012 drop in 20 feet of pipe.

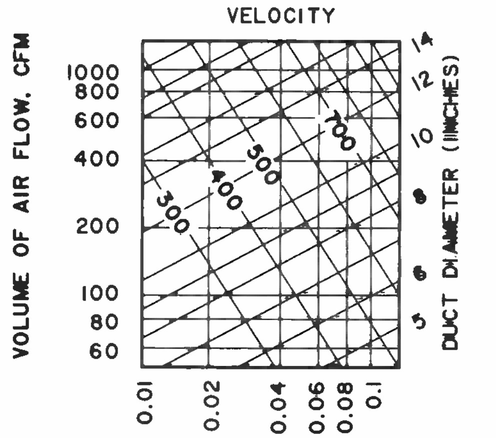


Fig. 68. Friction loss in pipes.

Considering the static pressure drop of two 90° elbows, to give a 180° bend in the pipe, it is found that a 10-inch elbow at 900 feet per minute gives less than .01 pressure drop, and we have only 600 feet per minute in our system. See pressure loss in a typical 5 gore round 90° elbow (Fig. 69). Adding the drops together, the result is as follows:

.012 + .01 + .01 = .032 total pressure drop.

Therefore, no exhaust fan is necessary.

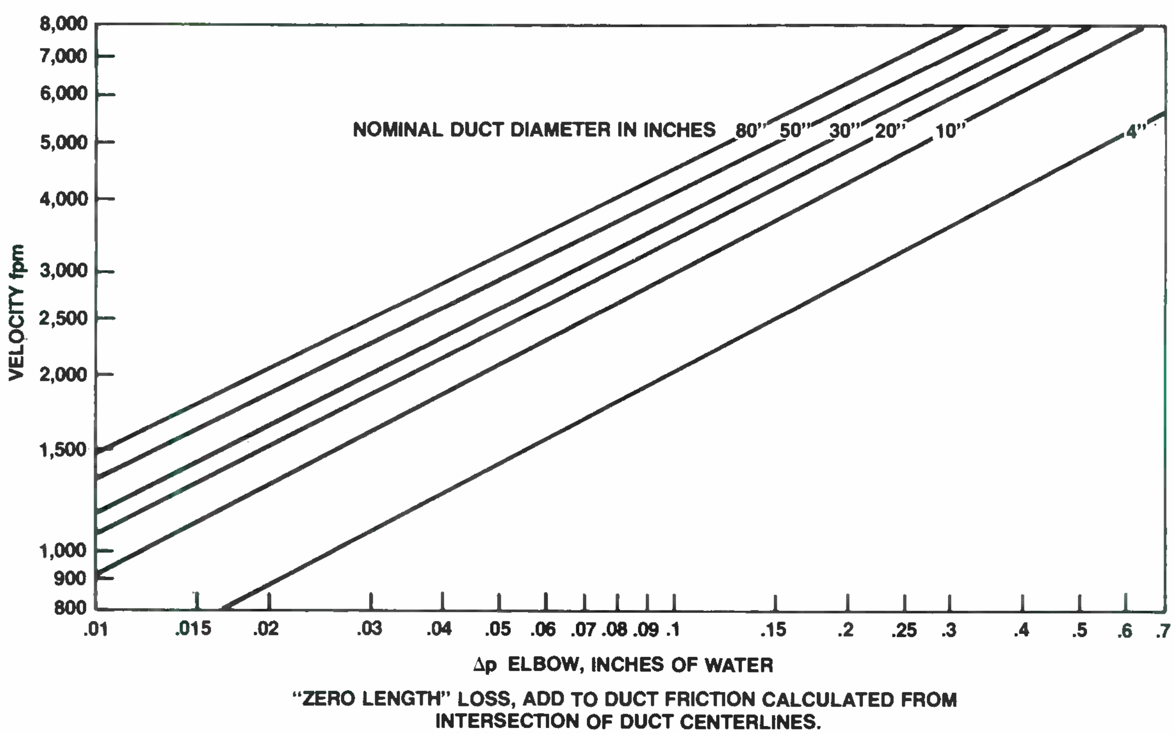


Fig. 69. Friction loss in 90° elbow.

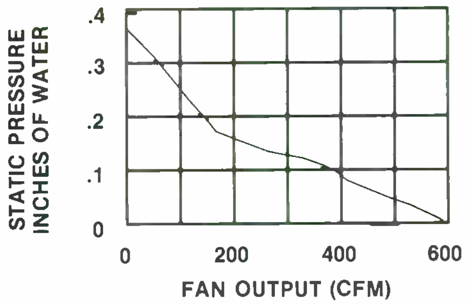


Fig. 70. Fan performance curve.

If the installer has a problem in exhausting the transmitter in this simple fashion, for example a roof exit is not available, and is required to add 2 additional 90° elbows and a straight run of 10-inch pipe 100 feet long. The pressure drop in the exhaust system will be as follows:

Friction loss in 20 ft of 10″ dia. pipe = .012
4 elbows × .01″ H_2O (est.) = .04″ H_2O.
Friction loss/100 ft. of 10″ dia. pipe = .06″ H_2O

0.112 H_2O
total loss in system.

Because the transmitter manufacturer recommends no more than a 0.1 inch H_2O loss in an exhaust system, good engineering practice indicates that an exhaust fan is required in this configuration. Searching through vendor catalogs for a fan results in the selection of a vaneaxial Caravel fan. Note that the performance curve on this fan (Fig. 70), indicates that it will deliver 325 to 330 CFM into 0.1 inch H_2O back pressure. This is sufficient to handle all of the air coming from the transmitter and overcome all of the estimated duct losses in this configuration.

It is also necessary to supply at least 325 CFM of air to replace the air circulated through the transmitter and exhausted to the outside of the building.

Because the transmitter only requires 325 CFM of air, it can probably be assumed that there would be enough leaks around windows and doors to supply this amount of air. However, if the installer wishes to insure an adequate amount of replacement air, a blower or fan may be installed in the wall of the building to supply the makeup air. Referring again to Fig. 70, examination of the curve shows that at 325 CFM delivery, this fan will develop a static pressure of 0.1 inch H_2O, which is sufficient to overcome the less than .01 inch H_2O pressure drop of a 10-inch diameter 90° elbow installed on the outside wall of the building for a weather hood, as well as a filter and bird screen to clean the incoming air.

20 ft of pipe = .012″ H_2O loss
2 elbows at .01 each = .020

.032 total loss,
maximum.

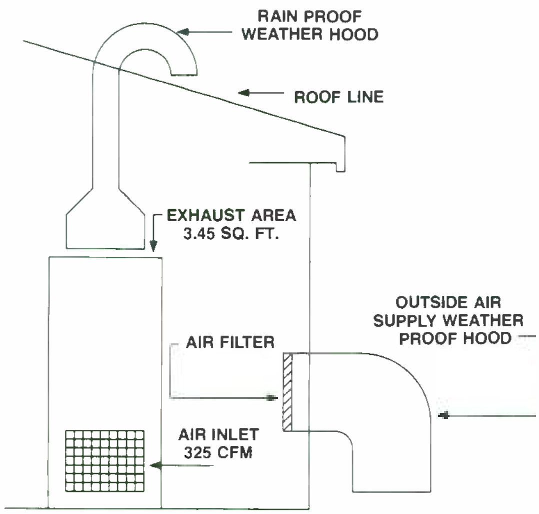


Fig. 71. Suggested intake and exhaust duct installations.

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 (visual and aural)

For combining of sources of the same frequency 90° 3 dB hybrid couplers have found

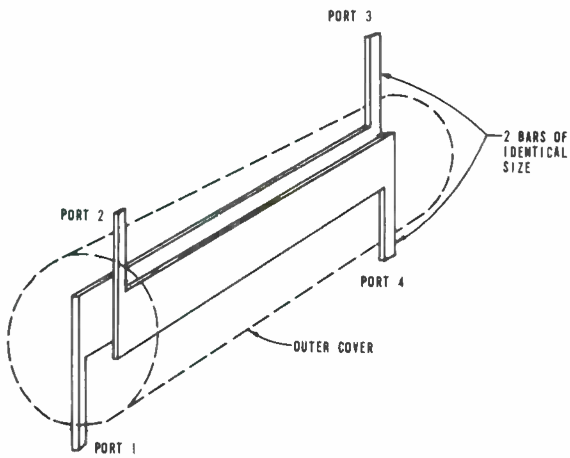


Fig. 72. 3 dB 90° hybrid coupler.

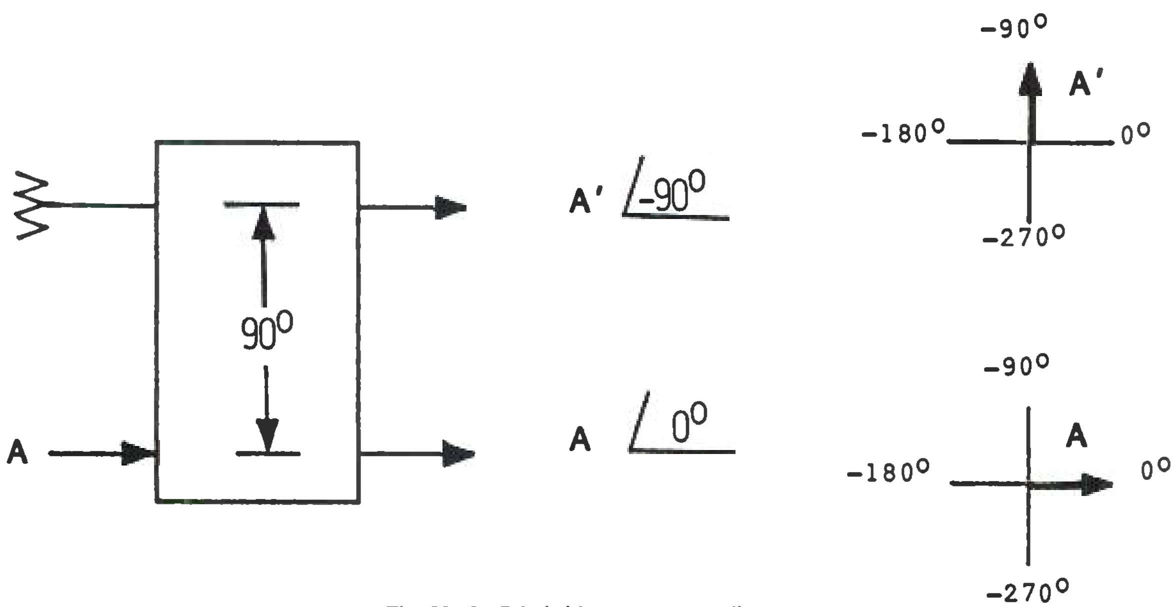


Fig. 73. 3 dB hybrid as a power splitter.

almost universal acceptance. Fig. 72 shows the physical model of a 90° 3 dB hybrid coupler.

A 3dB hybrid coupler consists of two identical parallel transmission lines coupled over a length equal or approximately equal to the quarter wavelength of the frequency to be acted upon 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 hybrid coupler can be used as a power splitter or as a power combiner. Fig. 73 shows the 3 dB hybrid coupler as a power splitter and Fig. 74 shows the 3 dB hybrid coupler as a power combiner.

One of the common uses of the 3 dB hybrid is to combine two frequencies that are close (Visual and Aural) in a back to back hybrid called a notch diplexer. The first step in building a notch diplexer can be shown by combining two 3 dB hybrids back to back as shown in Fig. 75. The input hybrid will act as a splitter while the hybrid at the output will act as a combiner.

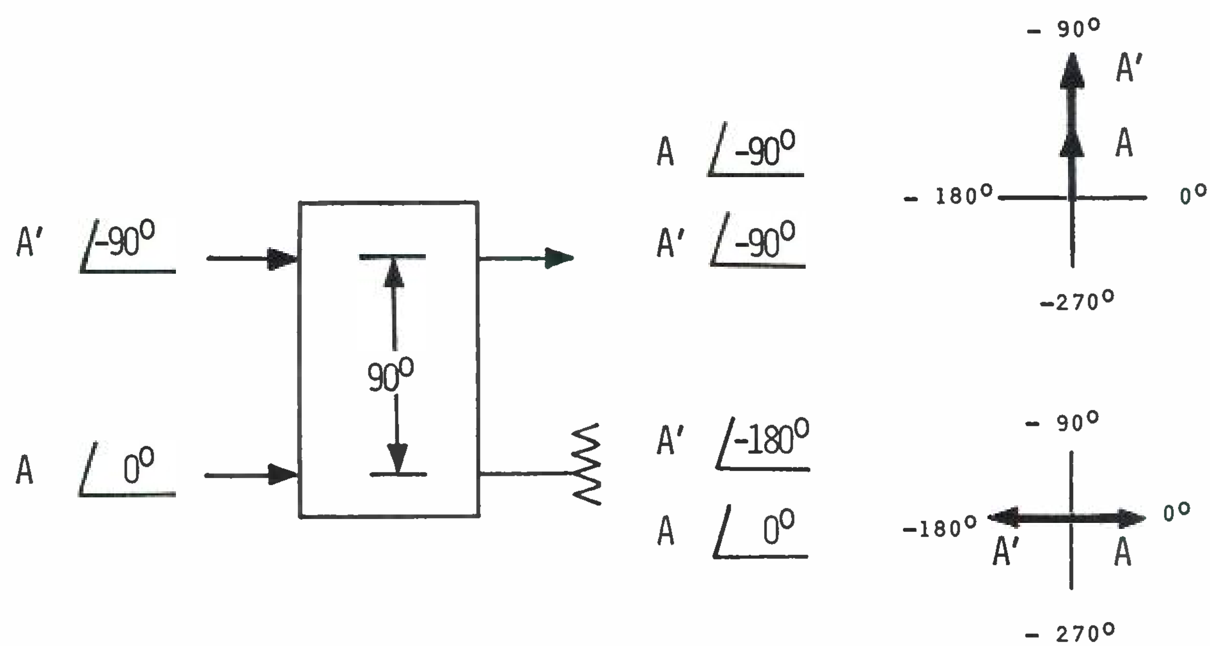


Fig. 74. 3 dB hybrid as a power combiner.

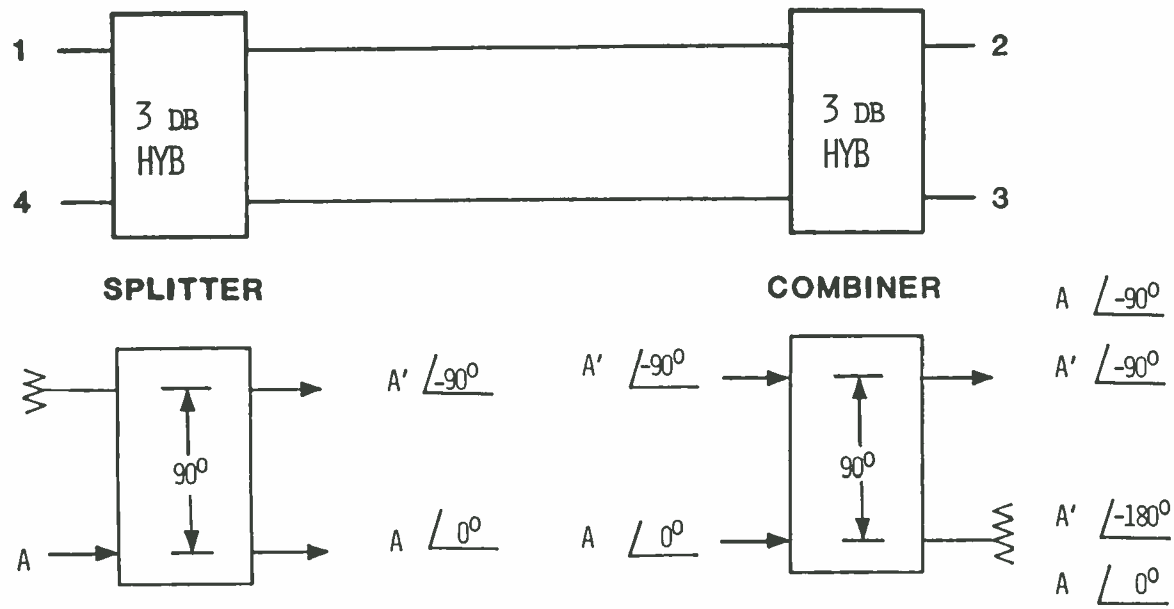


Fig. 75. Back-to-back hybrids.

If we now insert a Visual signal to location *A* in Fig. 75 we have the Visual signal appearing at the Antenna port as seen in Fig. 76.

If we now add the Aural signal as in Fig. 77, we can see that it is passed through the 3 dB hybrids but does not appear at the antenna port as required.

If we now add the Aural signal as in Fig. 77 but also add two high Q notch cavities tuned to the aural frequency, as depicted in Fig. 78, we can introduce a short to the aural frequency at the −3 dB output of its hybrid as seen in Fig. 79.

These shorts cause the aural signal, now split into two −3 dB components 90° out of phase by the 3 dB hybrid acting as a power splitter, to be reflected back into the 3 dB hybrid and where the 3 dB hybrid now effectively acts as a power combiner and passes the combined −3 dB aural signals to the antenna port along with the Visual signal. The Aural notch cavities are effectively not seen by the Visual signal since they are high Q cavities tuned to the Aural frequency.

Another use of the back to back hybrids can be seen in Fig. 80 which shows a single signal source driving a power splitter and dual amplifiers. This same arrangement can be used for Visual or Aural signals. Gain and phase adjustments are used to keep the signals out of the amplifiers (A1 and A2) at equal output power and 90° phase difference so that they can be properly combined by the power combiner with a minimum of reject load power.

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

Differences in amplifier chain gain are compensated for by the attenuators AT1 and AT2 which vary the amount of input signal applied to the amplifier chains. Differences in phase are

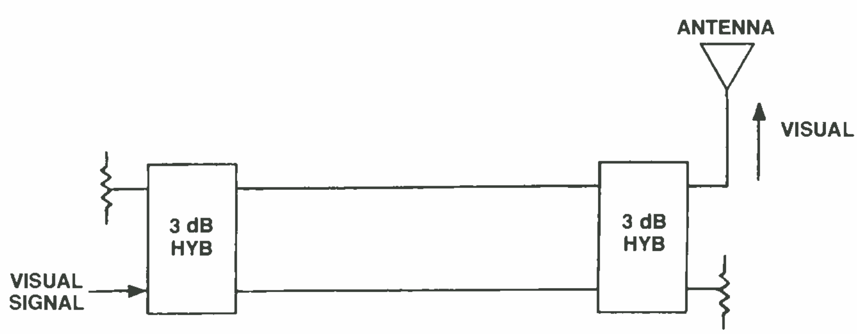


Fig. 76. Visual signal applied to back-to-back hybrids.

Fig. 77. Aural signal applied to back-to-back hybrids.

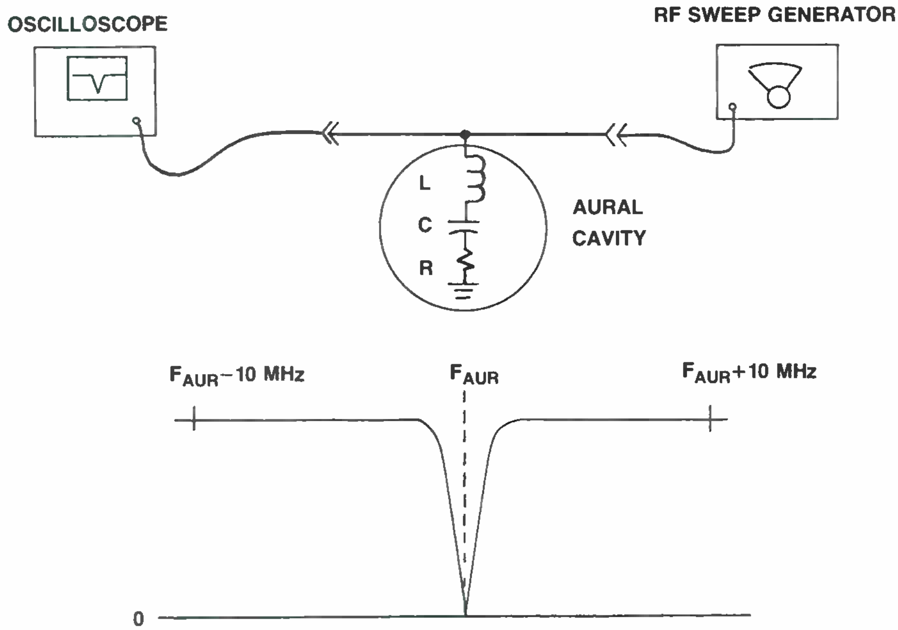


Fig. 78. Aural notch cavities with LC series resonant components and resistor (R) representing non-infinite Q of cavity.

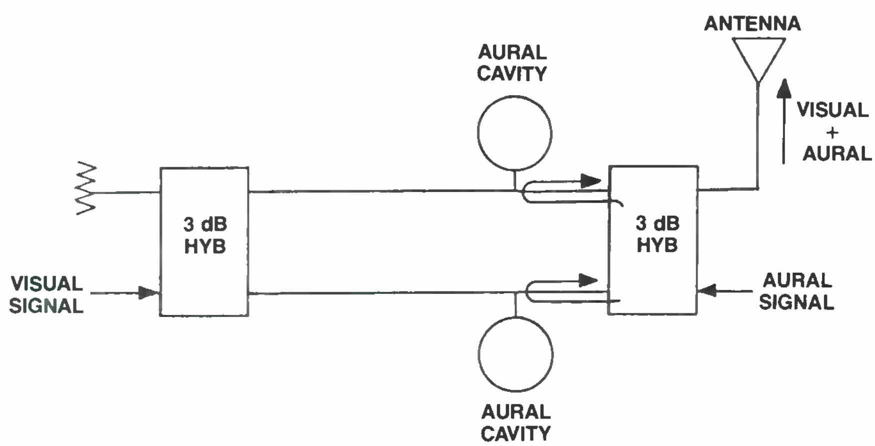


Fig. 79. Aural and visual signals applied to back-to-back hybrids with aural notch cavities to reflect aural signals back to antenna port.

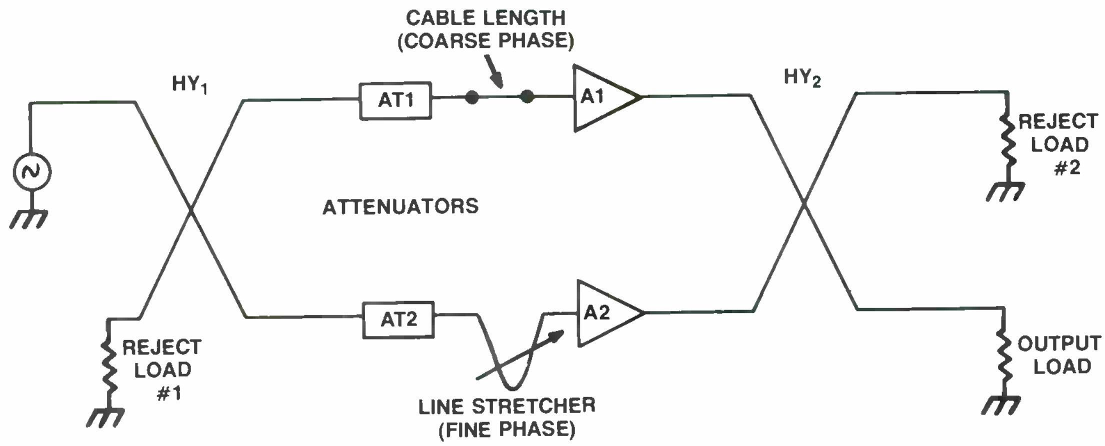


Fig. 80. Typical parallel amplifier with single signal source.

compensated for by the length of cable between the attenuators and the amplifier chain. Optimum phase and gain adjustments are determined by minimum power being dissipated by reject load number 2 and maximum power into output load as shown in Fig. 80.

Effective power transferred to antenna or load with respect to differences in phase or gain of amplifier chains can be seen in Figs. 81 and 82 respectively. As can be seen from these figures, a 60° error in phase will cause only a 25% reduction in output power and if one amplifier has only half the output power of its counterpart, a 3% reduction in power (in reference to the combined input powers) will result. (Transmitter A = 5 kW, Transmitter B = 10 kW, $K^2 = P_a/P_b$ = .5 which gives a K of .707 which equals approximately 3% in Fig. 84 or approximately 14.55 kW of useful power out of hybrid.) Worst case for gain occurs when one transmitter is not producing any output; in this case half of the remaining transmitter's output will be dissipated in the reject load and the other half will be applied to antenna.

A typical dual transmitter with notch diplexer can now be block diagrammed as in Fig. 83.

PRECISE FREQUENCY CONTROL

The limited number of available channels for TV Broadcasting makes it necessary to assign the

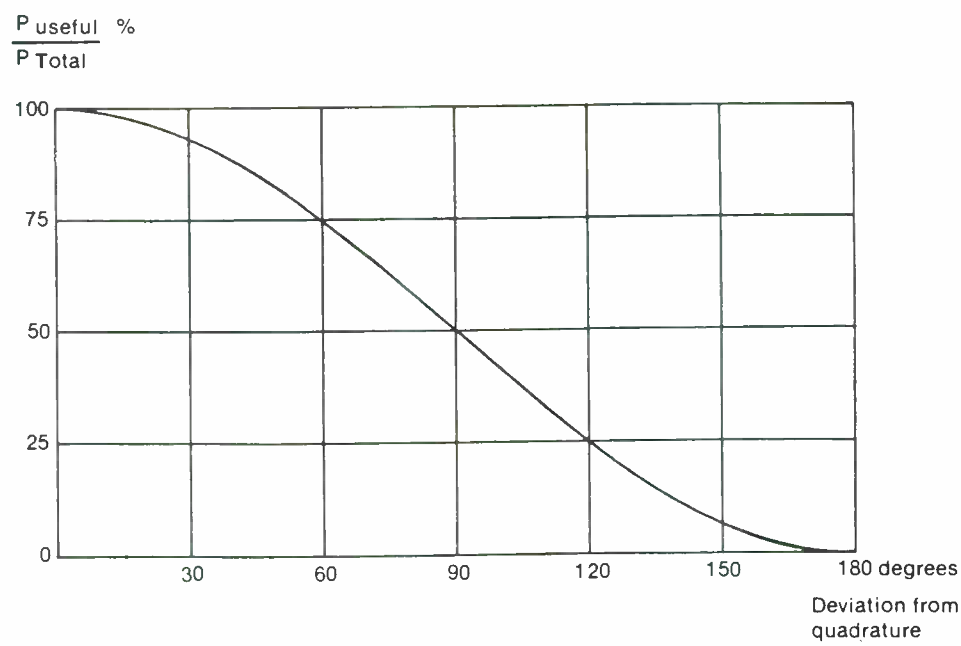


Fig. 81. Hybrid coupler phase sensitivity.

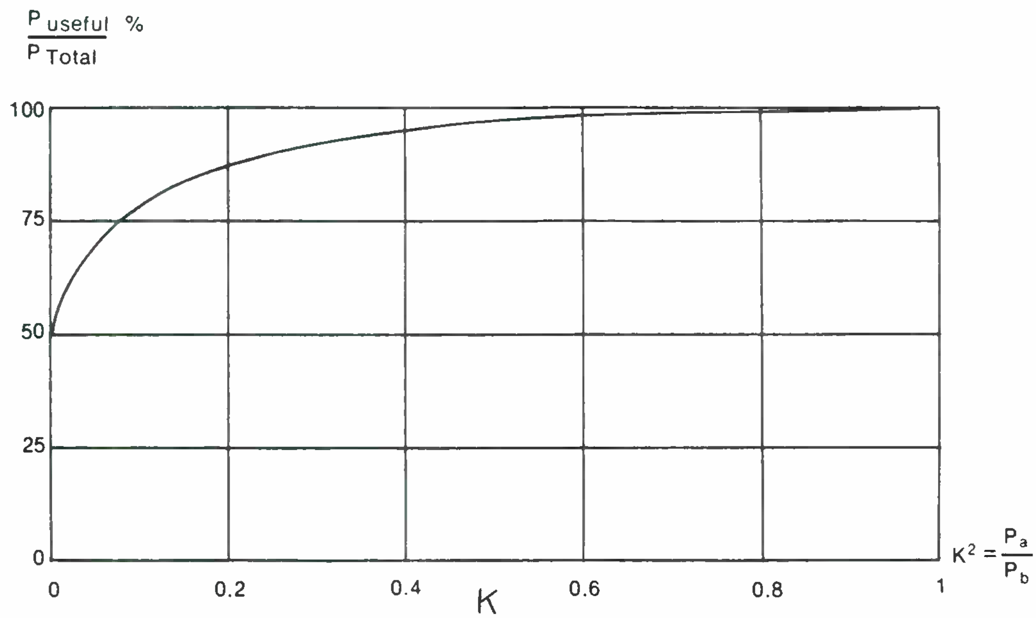


Fig. 82. Power imbalance in hybrid couplers.

same carrier frequencies to many stations. To avoid interference between stations operating on the same frequency (co-channel interference) geographical separation and radiated powers are carefully selected.

Nevertheless, considerable co-channel interference was encountered in many locations and investigation showed that additional means were available to reduce co-channel 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 interfering carrier but to choose operating parameters to reduce the visibility of the interference.

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

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

Co-channel interference between television stations appears to viewers as a horizontal pattern of alternating light and dark bars on the viewing screen—very much like the shadows cast by venetian blinds. It has been demonstrated for many years that the visibility of these bars varies cyclically as a function of the difference in frequency of the interfering carriers (Fig. 84). The interfer-

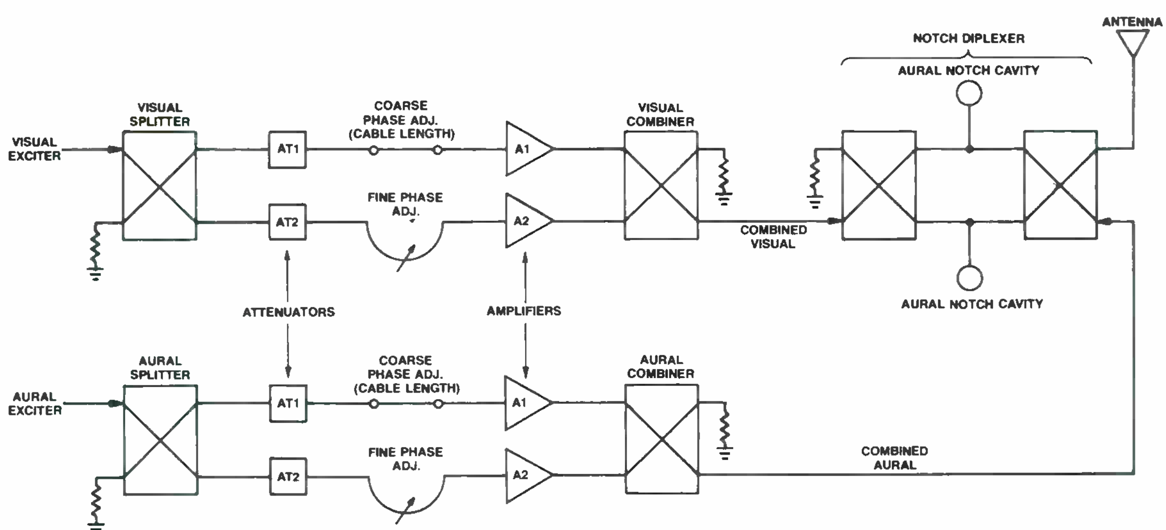


Fig. 83. Typical dual transmitter with notch diplexer.

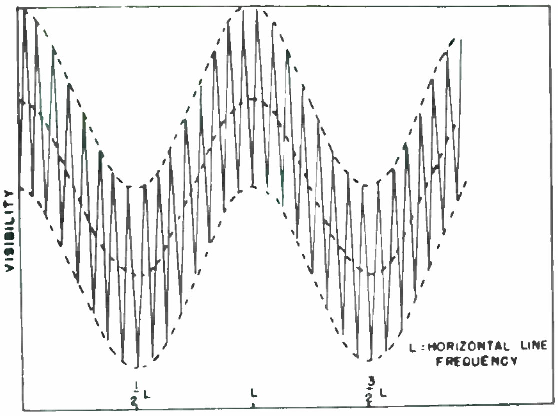


Fig. 84. Co-channel interference.

ence 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 approximately equal reduction of the interference patterns for any number of stations in geographical proximity, refer to Fig. 85.

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 general, it is not practical to build into the visual carrier oscillator the required stability for precise frequency control. Most oscillators used in visual service have sufficiently short time stability for effectively locking to an external frequency standard. Fig. 86 is a block diagram showing

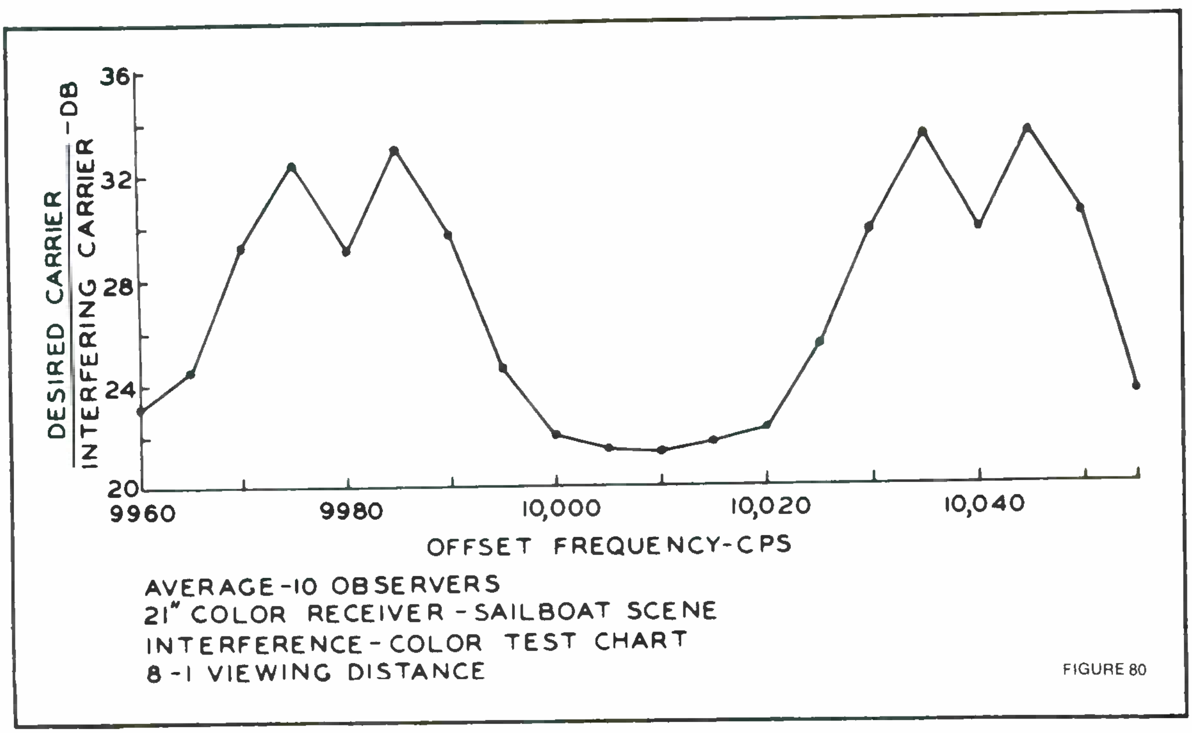


Fig. 85. 10 kHz offset pattern.

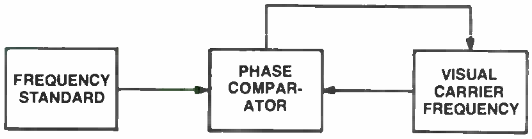


Fig. 86. Precision frequency control block diagram.

the visual carrier frequency in a phase lock loop with a frequency standard.

In exciters using two independent oscillators, the visual carrier feed to the phase detector is derived from mixing the oscillators together. For exciters using synsthesizers, the reference oscillator is fed to the phase detector.

By phase-locking the visual carrier to a stable reference oscillator, the Master Oscillator acquires the stability of the reference source. Sources which use an internal WWVB Receiver/Comparator to self-correct or an Atomic Frequency Standard can easily provide the stable reference source required. These reference sources allow a transmitter to be maintained within a few Hertz of a desired frequency indefinitely. Experimental results have indicated that frequency differences between transmitters can vary as much as ± 5 Hz from the precision offset before co-channel interference becomes noticeably worse. This means that two television stations can minimize their co-channel interference without the requirement to frequently adjust the transmitter frequency.

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

To ascertain the carrier frequency of a station operating under precision frequency control, assuming a measuring accuracy 10 times better than the quantity to be measured, leads to the following accuracy requirements:

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

Frequency counters with the above accuracy and with NBS traceable calibration be used for making measurements.

PERFORMANCE MEASUREMENTS

In the age of deregulation, TV signal quality responsibility lies with the broadcaster. Over the past 10 years, critical signal performance parameters improved two to one. Transmitters led the improvement with the introduction of IF modulation, solid state SAW vestigial filters and sophisticated precorrection circuits. Precision demodulators followed with synchronous detection and SAW filters capable of near ideal detection. New test waveforms and digital signal synthesis provided accurate test signal generation. Home receivers adopted high tech circuits including comb filters for wider band luminance, stable SAW IF filters, high resolution display devices, and sophisticated LSI video processing circuits.

It is difficult to show the correlation between signal quality and audience ratings. The issue of viewer enjoyment is easier to demonstrate. Many scientific studies relating signal degradation to perceived quality have been made over the years. These empirical studies have been the basis for *K* factor ratings and TASO picture ratings.

The Electronic Industries Associates Standard RS508 takes into account the empirical quality factors and reflects a common denominator for new transmitter performance. This standard is a valuable reference document which describes parameters, standards and methods of measurements.

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

Measurements can be broken down into two categories as follows:

1. Frequency response.
2. Linearity.

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

Linearity performance includes distortions which are signal level dependent and are a function of the instantaneous luminance level and on the average picture level over several lines. For example; level dependent chroma gain and phase called differential gain and phase, respectively; luminance gain variation called low-frequency nonlinearity; and frequency response versus brightness which is the change in swept response with static luminance level. The nonlinear responses can also change as a function of average picture level. For this test the video waveform

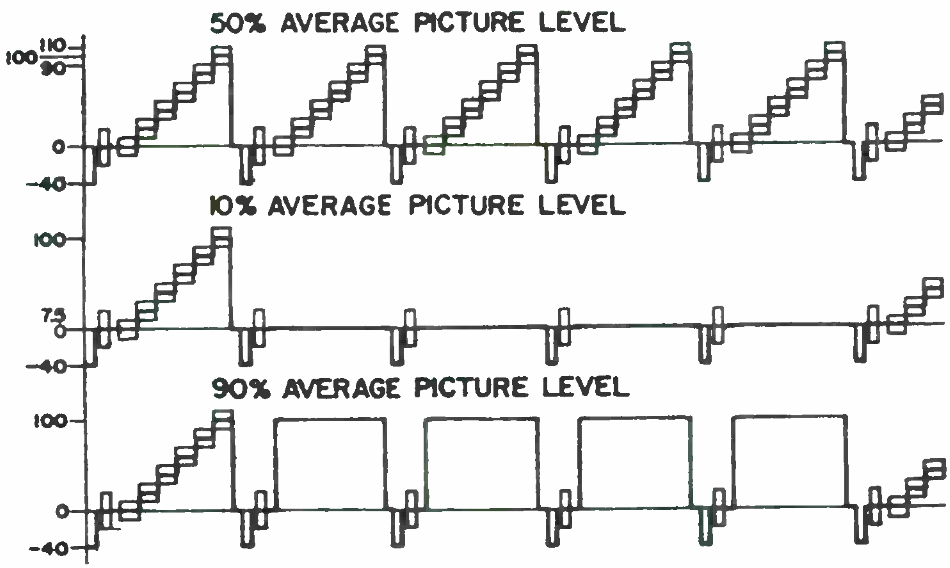


Fig. 87. Modulated staircase waveform.

is alternated with four lines containing a static luminance level. Fig. 87 contains a modulated staircase with three average picture levels.

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

The following is a pretest check and a sequence of tests presented as a general guide for arriving at a properly adjusted transmitter.

Pretest Checks

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

Transmitter Test Sequence

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

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

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

For off-air monitoring, a picture monitor is convenient for accessing gross video degradations quickly. Vertical interval test signals containing the composite waveform or other specialized test signals are useful for keeping track of depth of modulation, and fine grain linear and nonlinear responses. These test signals are displayed on a waveform monitor and vectorscope.

Records of transmitter meter readings and performance data are very useful in maintaining a transmitter at like new operating conditions. Maintenance logs should include date, time duration of outages, corrective action taken, and where possible, identify cause.

Equipment Needed

In order to make satisfactory measurements of video signals, a certain minimum of equipment is required. Measuring the output of a television transmitter at baseband requires a demodulator. For greatest accuracy, this demodulator should provide a zero carrier reference pulse for determining percentage of modulation, and ideally will have both a synchronous and envelope detection mode. For most baseband measurements, a video waveform monitor is all that is needed, but to do a complete analysis of a system requires some additional equipment. The waveform monitor should include a provision for making measurements on vertical interval test signals (VITS) and provide filtering to allow separate examination of the luminance and chrominance components of color television signals. In addition to the waveform monitor, a vectorscope must be used for making certain measurements on color television system, particularly differential gain and phase measurements. The waveform monitor and vectorscope are all that are required to accomplish these measurements, but if a greater level of accuracy is desired in timing measurements, a conventional oscilloscope with an associated digital counter/timer would be a great asset.

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

At one time or another, just about every component part of a video signal must be measured or adjusted. It is good engineering practice that each part of a transmission plant including studio, distribution, transmitter, and monitor equipment provides minimal distortion to the video signal. Furthermore it is not recommended that one part of the system correct for another part. The signal going into the transmitter should be a standard NTSC video signal.

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

REFERENCES

1. Electronic Industry Association. (To Be Published) RS-508 "Electrical Performance Standards For Television Broadcast Transmitters".
2. Electronic Industry Association. RS-462 "Electrical Performance Standards For Television Broadcast Demodulators".
3. Hans Shmid. "The Measurement Of Short-Time Waveform Distortion In NTSC TV Facilities" IEEE Trans. Broadcast, Vol 18, No. 3, Sept. 1971.
4. Charles Rhodes. "The 12.5T Modulated Sine-Squared Pulse For NTSC" IEEE Trans. Broadcast, Vol 18, Mar. 1971.
5. Pieter Fockens, Carl Eilers. "Inter carrier Buzz Phenomena Analysis And Cures" Transmitter Consumer Electric Vol 27, No.3, August, 1981.
6. C. Bailey Neal. "Frequency And Amplitude Dependent Phase Effects In Television Broadcast Systems" IEEE Trans On Consumer Electric, Vol 23, No. 3, August 1977.
7. T. M Gluyas. "Influence Of RF Output Systems On TV Transmitter Picture Quality" NAB Convention Record, April, 1972.
8. L. E. Weaver. "The Multipulse Waveform" Application Note No. 30, Tektronix Corporation.
9. W. A. Resch. "Visual Performance Characteristic Which Affect Multichannel Sound" NAB Convention Record 1982.
10. W. L. Behrend. "Effects Of Incidental Phase Modulation Of TV Transmitters Or Other Circuits, On TV Signals" IEEE Trans On Broadcasting, Vol 19, No. 3, September 1973.
11. C. J. Edgecombe. "The Television Performance Of The Klystron Amplifier" Radio And Electronic Engineer, September 1971.
12. A. Hans Bott. "Television Transmitters Chapter 19" NAB Engineering Handbook 6th Edition.
13. *TV/VIDEO Sync Primer And Product Note* Opt. 005-1, November 1981, Hewlett Packard.
14. Vol. XI Recommendations And Reports Of The CCIR, 1982. "Subjective Assessment Of The Quality Of Television Pictures" Report 405-4.
15. Varian Associates, INC. "Integral Cavity Klystrons for UHF TV Transmitters".
16. Varian Associates, INC. "External Cavity Klystrons for UHF TV Transmitters".
17. C. F. Smiley & G. L. Best. "Improved Efficiency Klystrons in UHF TV Transmitters" NAB Convention Proceedings, 1984.
18. Robert S. Symons. "Klystrons for UHF Television" Proceedings IEEE Vol. 70, #11 Nov. 1982.
19. Gregory L. Best. "Transmitter Conversion for TV Multichannel Sound" NAB Convention Proceedings, 1984.

20. Joseph C. Wu. "Monitoring TV Multichannel Sound" NAB Convention Proceedings, 1984.
21. Carl G. Eilers. "The Zenith Multichannel TV Sound System" NAB Convention Proceedings, 1984.
22. Leslie B. Tyler, Mark F. Davis, and William A. Allen. "The dbx TV Companding System for Multichannel TV Sound".
23. United Sheet Metal Bulletin 100-5-178 "Engineering Design Manual for Air Handling Systems".
24. Rotron Incorporated. "The Caravel Fan" Catalog Sheet E-3050C.
25. McGraw-Hill. Marks' *Mechanical Engineers Handbook.*
26. Andersen Laboratories, *Handbook of Acoustical Processing, Vol. I.*
27. A. G. Uyttendale, "Design Requirements and Operational Features of the Gates Dualtran RF Switching System," Harris Corp., Broadcast Products Division.

3.6

Multichannel Television Sound

Edmund A. Williams
Staff Engineer,
Department of Science and Technology
National Association of Broadcasters
Washington, DC

INTRODUCTION

Multichannel Television Sound (MTS) is a generic term for the process of adding subcarriers to the aural carrier of a television station. Some of these subcarriers are designed to be received by the general public and may be used for a variety of different purposes including stereophonic sound, second language, commercial communications, and data. Other subcarriers may be used for news crew cueing, transmitter telemetry or other digital or analog aural or data services. The subcarriers may or may not be related to the program on the video or aural portion of the television signal itself.

BACKGROUND

The addition of subcarriers to the aural carrier was approved by the Federal Communications Commission on March 29, 1984, in response to a petition for rulemaking from a television station interested in transmitting cueing instructions to remote news crews during live news events. In order to develop a standardized approach to the implementation of multiple sound channels, the broadcast industry, through the Electronics Industry Association (EIA) and its Broadcast Television Systems Committee (BTSC), established laboratory facilities and conducted transmission tests on several potential multichannel sound systems. The system selected by the EIA was a combination using the transmission parameters developed by Zenith Electronics Corporation and the noise reduction system developed by dbx, Inc. The combination is called the BTSC Multichannel Television Sound system after the EIA BTS Committee which conducted the laboratory tests. A chronological table of events for the development of MTS is shown in Fig. 1.

In response to the EIA's recommendation, the FCC, in their Second Report and Order

1. Provided protection to the transmission and reception of the BTSC system,
2. Produced an Office of Science and Technology Bulletin #60 containing the system's specifications, and
3. Permitted stations to implement not only the BTSC MTS system but any other system of subcarriers a station desired so long as no interference resulted to BTSC receivers or regular monophonic sound television receivers.

While not an absolute standard, the Commission's approach provides the next best thing—protection. Transmitter and receiver manufacturers are able to produce equipment operating with the BTSC system with full knowledge that there is a compatible, protected MTS system recognized by both FCC and industry. The industry has also developed a series of recommended operating practices some of which are

MTS CHRONOLOGY

- WCVB PETITION FOR RULEMAKING 1/21/77
- FCC NOTICE OF INQUIRY (DOCKET 21323) 7/1/77
- EIA MTS COMMITTEE FORMED 1/10/79
- FCC NPRM (NOTICE OF PROPOSED RULEMAKING) 11/20/79
- FCC 1ST REPORT & ORDER (CUEING) ... 6/30/81
- FCC FURTHER NPRM 7/28/83
- EIA CONDUCTS VOTE 12/22/83
- FCC 2ND REPORT & ORDER (STEREO) ... 3/29/84
- RECEIVERS SHOWN TO PUBLIC 6/4/84
- FIRST STEREO TEST BROADCAST 7/27/84
- FIRST FULL TIME STEREO BROADCAST . 8/6/84

Fig. 1. MTS Chronology.

summarized later in this chapter. Further, stations may also employ subcarriers for virtually any other purposes which do not interfere with the BTSC system or normal monophonic receivers operation.

It should be mentioned that while the multichannel sound system adopted for use in the United States is unique, multichannel sound transmission schemes were developed and adopted in at least two other countries before the FCC Order. In Japan, an FM subcarrier system for either stereo sound or a second language was adopted in 1977 and stations began transmission of this system in 1978. In 1981, West Germany adopted a two-carrier sound transmission system providing either stereo or second language. In comparison, the U.S. system, adopted in 1984, provides a combination of AM and FM subcarriers to transmit both stereo and second language simultaneously. Further, the US system permits other subcarrier arrangements which serve an individual stations requirements.

THE BTSC SYSTEM OBJECTIVES

1. The system must be compatible with existing monophonic television receivers.
2. The system must provide both stereo and 2nd language simultaneously.
3. There must be provisions for other professional use or station use subcarriers.
4. A noise reduction (companding) system must be employed in both stereo and second audio program channels.
5. Aural carrier deviation must be increased to accommodate the new subcarriers.

TRANSMISSION FORMAT

In order to understand the following description of the transmission format of the BTSC system with respect to the television aural baseband, the reader should refer to Fig. 2.

The design of the basic MTS system took into account several major considerations. Among these were such important factors as compatibility

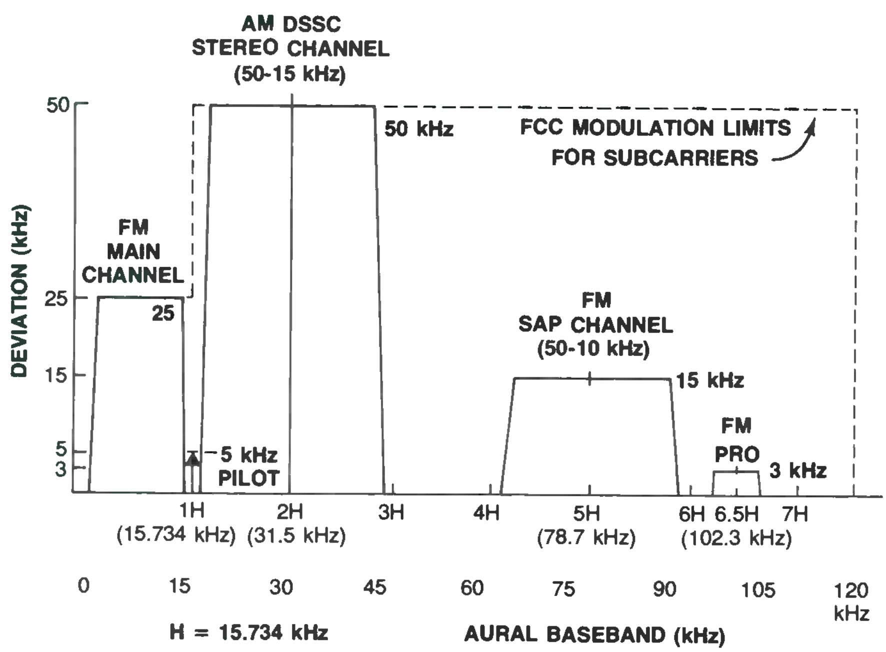


Fig. 2. BTSC MTS baseband.

with existing television receivers, ease of modification of transmitters, full sound fidelity, the ability of the MTS signals to pass through cable TV, master antenna systems, TV translators, the ruggedness to withstand substantial transmission impairments in the path between the transmitter and receiver. These factors influenced the design criteria with the result that the single carrier system with subcarriers was favored over a two-carrier system, and an audio companding or noise reduction system was required. Also, a seperate audio program channel for special programming purposes was provided.

Transmitting the stereophonic component in the form of a difference signal or left minus right (L – R) rather than any other mathematical derivation for the stereo matrix was based mainly on the U.S. FM broadcast model and its familiarity to broadcasters. For example, the West German system transmits the sum (L + R) signal on the main aural carrier and a plus right (+R) signal on the second carrier. Further, additional subcarriers for cueing, data, telemetry or other station related purposes may also be transmitted. Space for these professional use subcarriers is reserved in the aural baseband as well.

MONOPHONIC CHANNEL

The basic monophonic, or sum channel (L + R), is maintained at its present place in the aural baseband, remains at 25 kHz peak deviation, and uses the standard 75 microsecond pre-emphasis. All other subcarriers increase the overall deviation of the main carrier. Increasing the deviation of the aural carrier does not result in any degradation to the normal monophonic sound in existing receivers. Maintaining 25 kHz deviation for the monophonic channel also avoids the need to reduce the modulation level of the monophonic signal during MTS transmission which would mean lowering the signal-to-noise ratio of the aural channel and "loudness" of the sum channel compared to stations not operating with the BTSC system. Increasing deviation provides the "room" for the new stereo, separate audio program and professional use subcarriers to be added to the aural baseband successfully.

The fidelity of the main channel monophonic audio is maintained as before. But, because of the closeness of the sidebands of the stereo subcarrier, BTSC stereo encoder equipment incorporates sophisticated lowpass filters in the audio circuits to limit response of each audio channel to no more than 15 kHz. Nevertheless, the BTSC system is capable of providing audio quality essentially limited only by the quality of the main or monophonic channel.

STEREO SUBCARRIER

In contrast to the monophonic or sum channel which is frequency modulated and deviates the main aural carrier by 25 kHz, the stereophonic subchannel audio is placed on an amplitude modulated, double sideband suppressed carrier, subcarrier at a frequency of 2H or 31,468 Hz (twice the NTSC video horizontal scanning rate of 15,734 Hz.) The subcarrier is modulated with the stereophonic difference signal or left minus right (L—R). The maximum injection of this subcarrier is set to 50 kHz main channel deviation. The suppressed carrier signal is injected into the aural carrier during modulation. The injection level is related to the companding system algorithm. However, because of the nature of stereophonic sound, a form of interleaving between sum and difference channels takes place. . .which prevents the deviation of the main carrier from becoming the sum of the two components which would appear to produce a total of 75 kHz. Instead, the maximum rarely exceeds 50 kHz. This is because a "left only" or "right only" signal originating from the studio would produce only a 50% modulated signal in the mono channel and a similar level in the difference channel.

The difference channel also employs the noise reduction circuit which further controls the maximum level of modulation of the difference signal. The dynamic response characteristic of the compressor is shown in Fig. 6. The result of this combination of events produces no more than 50 kHz deviation of the aural carrier. Finally, most stereo encoder manufacturers incorporate some form of peak limiting device to prevent the stereo subcarrier, the monophonic channel, and perhaps the combined deviation as well, from exceeding the maximum 50 kHz except for occasional peaks.

The selection of 2H for the center frequency of the stereophonic subcarrier was based on its relationship with the horizontal frequency component of the video in both the transmitter and the receiver. Crosstalk from video signal circuits, power supply and RF paths in the transmitter generally occur at multiples of H and therefore may be cancelled or at least dealt with in the receiver without significantly affecting overall stereo audio quality.

The first harmonic of H (15,734 Hz), present to some extent in many transmitter and receiver audio circuits, is above the normal hearing range of most viewers and not normally audible. Receiver audio circuits may also employ notch or low pass filters to eliminate the H signal at the audio output. The second harmonic falls at 2H (31,468 Hz) which is the frequency of the BTSC stereo subcarrier. Because the subcarrier frequency is locked to the video sync, the harmonic and subcarrier remain in phase and no beat occurs.

Modulation of the 2 H harmonic by 60 Hz field rate components of the video signal can also occur in either transmitter or receiver. The resulting buzz can be effectively reduced by the use of high pass filters in the stereo subchannel at the receiver to remove unwanted audio below about 100 Hz. While this has virtually no affect on the stereo audio quality, it substantially reduces low frequency buzz and hum caused by the field rate modulation. Placing the stereo subcarrier at a higher multiple of H is not desirable due to the much higher noise levels encountered in that portion of the aural baseband. A higher subcarrier frequency would also have made the addition of the separate audio program channel more difficult as well.

Fig. 3 shows the noise level build-up in the aural baseband. This illustrates why it was important to place the stereo subcarrier as low as possible in the baseband as well as the need for the audio companding system. The noise increases at a rate of 6 dB for each doubling of the bandwidth. As a result the noise the stereo subcarrier would add to the received dematrixed sound is about 21 dB. This is reduced by 6 dB because the stereo subcarrier deviates the main carrier by 50 kHz compared to the main channel deviation of 25 kHz. The resulting degradation in signal-to-noise is only about 15 dB which is easily overcome by the noise reduction system.

The use of an AM modulated double sideband suppressed carrier, subcarrier also provides several other significant advantages over an FM subcarrier. AM subcarrier technology is well known and developed within the FM broadcast industry and circuit and performance refinements have been raised to near-perfect levels. Second, AM offers lower theoretical distortion levels than FM, given the amount of spectrum available for transmission. The bandwidth for the stereo subcarrier channel is 30 kHz.

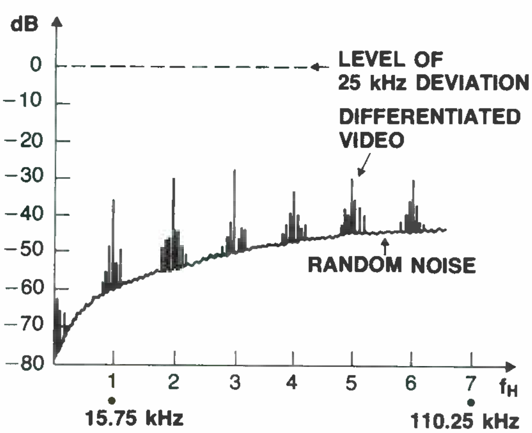


Fig. 3. Spectrum of noise on the received aural baseband. (Courtesty J.J. Gibson, RCA)

AM detection techniques are such that the distortion can be exceedingly low—on the order of 0.1% or lower. An FM subcarrier with enough deviation to produce the desired S/N and limited to only 30 kHz would produce higher distortion (1.5-3.0%) when detected. Calculations which prove this point may be found in the EIA MTS Report Vol 2B. And finally, further enhancements of the aural signal are possible through the use of quadrature modulation of the AM subcarrier for some future aural requirements such as a center channel or surround sound. Frequency response of the stereo subchannel is a good solid 15 kHz which properly compliments the main channel.

PILOT CARRIER

It was mentioned above that the stereo subcarrier is transmitted with its carrier supressed. In order for the AM detector to have a reference, a PILOT signal is transmitted at one-half the subcarrier frequency similar to the technique employed in broadcast FM transmission systems. The pilot is an unmodulated carrier at a frequency of 15, 734 Hz locked to the video horizontal sync rate. It modulates the main aural carrier to a deviation of 5 kHz ±0.25 kHz. The stereo encoder requires a feed from the companion video signal in order to obtain its frequency reference. It might be thought that receivers should not need the pilot and could use horizontal sync information obtained from the receiver video circuits. While this could be the case in some instances, there are receivers which do not process the video when decoding the stereo audio and would not have a reference to synchronize the stereo detector. The pilot also serves as a signal that BTSC is being broadcast and for the receiver to activate the stereo circuits.

A disadvantage with providing the pilot at H rate is that there may be some residual H components in transmitters and receivers which, under normal operating conditions, could cause false indications or even malfunction of stereo decoders in receivers. Problems of this sort must be solved by broadcasters and manufacturers by adequately shielding or isolating the H components of the video and video circuits from the audio circuits in transmitting and receiving equipment designs and modifications to existing equipment. For example, the level of 1H components from the visual transmitter 4.5 MHz above the visual carrier can be controlled by installing lowpass video filters, adjusting transmitter visual RF circuits or improving the 4.5 MHz rejection in the transmitter combiners or diplexers.

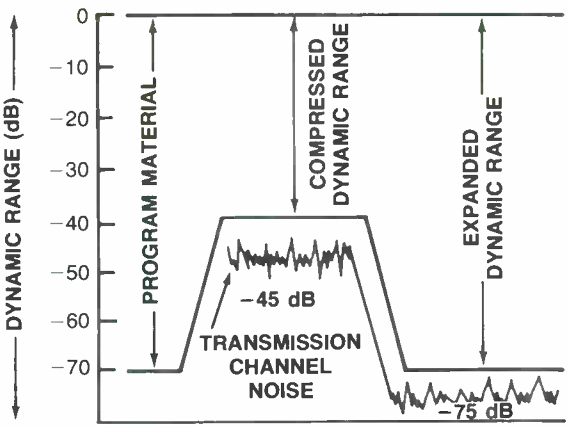


Fig. 4. Typical noise reduction action.

THE COMPANDING SYSTEM

Companding is a term used more frequently in the telecommunication industry to describe what broadcasters call noise reduction, a process which is well known to studio audio technicians. Its use is restricted to systems where there is control over both ends of the circuit. Examples include magnetic analog audio recorders, microwave and satellite circuits and telephone lines used for program transmission. Simply stated, noise reduction is compressing the dynamic range at the sending end and expanding the signal at the receiving end of the circuit. In the process of expansion the noise is reduced (see Fig. 4). Until the development of the BTSC MTS system, noise reduction techniques could not successfully be used for over-the-air transmission because a complementary expander is required in the receiver. MTS required the use of expanders in the receiver from the beginning of the service. As a result, MTS is the first broadcast service to use companding as an integral part of the transmission system.

Some FM broadcast stations employ a simple limited range compression arrangement for an over-the-air noise reduction system. But because few receivers are equipped with the corresponding expander, the compression must be held to about only 10 dB in order to avoid producing objectional side effects and becoming objectional to the listener.

Noise reduction is built into the BTSC MTS system from the beginning of the service and therefore a high performance system may be employed. Figs. 5 and 6 show how the compressor alters both the dynamic range and high frequency response of the program material according to the content. Dynamic range is reduced from 70 dB to about 40 dB. High frequency material is increased by as much as 30 dB at low levels but virtually not at all at high levels.

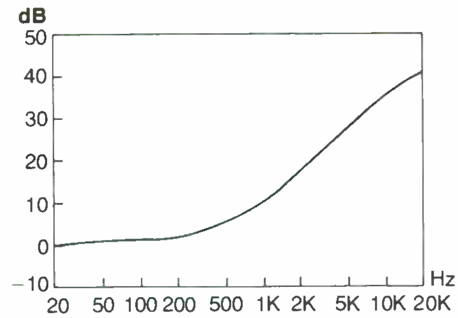


Fig. 5a. Fixed-preemphasis frequency response.

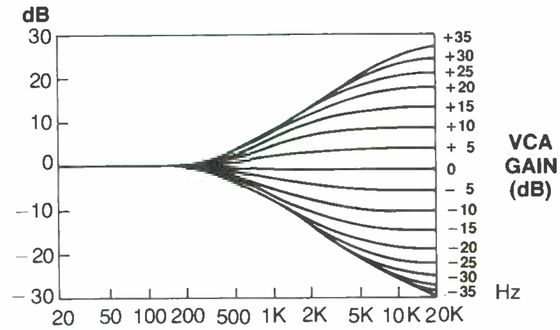


Fig. 5b. Spectral compressor, frequency-response range.

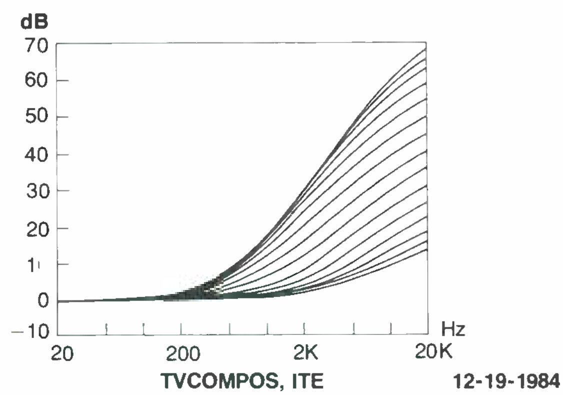


Fig. 5c. Frequency-response range of spectral compressor and fixed preemphasis.

Fig. 5. Spectral compression action. (Courtesy of dbx, Inc.)

The companding system employed by the BTSC MTS system was specifically designed to operate in the comparatively "hostile" environments presented by television transmitters, signal propagation, re-transmission systems (CATV, MATV, translators) and the receiver itself. Thermal and impulse noise, multipath distortion and buzz generated by transmitter and receiver all combine to present formidable obstacles for the companding system to overcome.

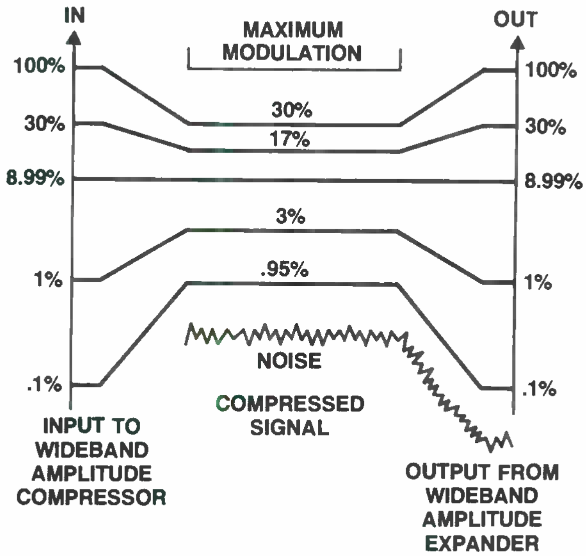


Fig. 6. Companding of (L-R) and SAP. (Courtesy of dbx, Inc.)

The companding system selected for the MTS employs a combination of fixed preemphasis, "spectral" compression, and amplitude compression. The fixed preemphasis combines the familiar 75 microsecond rising frequency response with a 390 microsecond network. The resulting curve is shown in Fig. 5a. By itself this extremely steep preemphasis curve would cause problems with high frequencies. To avoid this, a dynamic "spectral" compressor (variable preemphasis) circuit is employed to increase the gain of low level, high frequency material and reduce the gain of high level, high frequency audio. Only frequencies above about 200 Hz are affected. Fig. 5b illustrates the dynamic characteristics of the spectral compressor. A complicated algorithm involving both the fixed and variable preemphasis produces the sharply rising frequency response shown. Fig. 5c.

The third feature of the companding system is the amplitude compressor which reduces the dynamic range of the input signal by a factor of 2:1 for low frequencies, and 3:1 for high frequencies. In other words, a 40 dB dynamic range audio signal is reduced to 15 to 20 dB or less for transmission (see Fig. 6). In addition to dynamic range compression, the maximum input level applied to the stereo subcarrier is reduced below that required for maximum modulation of the subcarrier. This feature provides some headroom to accommodate instantaneous peaks without resorting to clipping. High level "transients" however are clipped to avoid severe distortion in the reconstruction of the compressed signal in the receiver expander.

The use of companders and noise reduction systems is gaining wide acceptance throughout the broadcast industry. Is is not unusual for a given program to undergo noise reduction processing 5 to 6 times by the time it reaches the viewer. The mis-tracking of a single companding system in the chain may result in substantial errors in the resulting sound which may be more objectionable than the problems the compander was designed to solve. It is essential therefore, that in order to prevent audible artifacts from occurring, only the highest quality systems be employed, properly operated and maintained.

THE SEPARATE AUDIO PROGRAM CHANNEL

Multichannel Television Sound transmission may include a second audio program (SAP) channel in addition to stereo. This SAP channel may be program related or not, as the station chooses. However, if not program related, viewers with the SAP option switch may wonder about the material if it is not related to the picture in some manner. Because it is designed for reception by the public, stations should avoid transmitting non-public material on the SAP channel unless the injection level of the subcarrier is kept below the receiver threshold.

The SAP channel subcarrier is located at 5H in the aural baseband as shown in Fig. 2. Although somewhat noiser at this location in the baseband than if it were placed at 4H, there is less chance for crosstalk to and from the stereo channel. The use of the noise reduction system on this channel provides adequate signal-to-noise ratios and, with the audio response limited to a maximum of 10 kHz, the performance of the SAP is quite good.

The SAP subcarrier is frequency modulated by program material to a maximum deviation of 10 kHz. The subcarrier injection into the main carrier is limited to a maximum of 15 kHz. That is, the subcarrier modulates (deviates) the main carrier by 15 kHz. The subcarrier frequency is locked to 5H when not modulated.

The injection level of 15 kHz by the SAP subcarrier into the main carrier is important. Receivers with SAP decoding equipment will detect the presence of the subcarrier and provide either automatic switching or at least an indicator lamp to the viewer. The receiver decoder will normally be set to activate the SAP circuits when the injection level exceeds about 7.5 kHz. Below this level the receiver will ignore the presence of a subcarrier. To use this portion of the baseband for some purpose other than for program related audio, a subcarrier with an injection of 3 kHz would probably be ignored by most SAP receivers but could of course activate the station's intended receivers. Obviously such a subcarrier could not

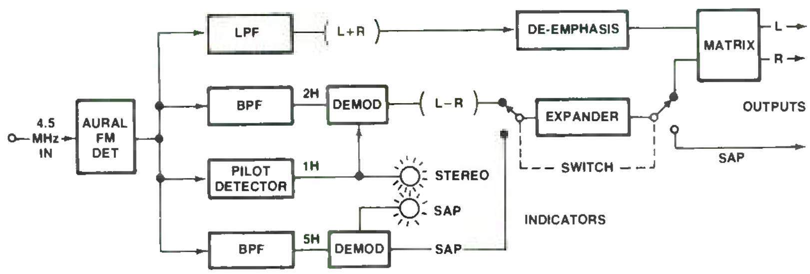


Fig. 7. Typical MTS receiver decoder.

be used at the same time as the regular SAP channel but could be used during stereo or monophonic operation. Fig. 7.

The same noise reduction scheme is used on the SAP channel as for the stereo subcarrier. This permits a single expander to be used in the receiver for either stereo or SAP. Most receivers will provide either stereo or SAP but not both at the same time. To make both available a receiver would need two expanders. Alternatively, the SAP output could be provided without expansion but the output would be highly compressed. The sound would be acceptable for most vocal material but may be objectionable for some music. While the use of a single noise reduction circuit helps the receiver, the broadcaster must deal with maintaining exact levels and performance for two compressors if both stereo and SAP are to be transmitted simultaneously.

OTHER SUBCARRIERS

In addition to providing for stereo and SAP subcarriers, the FCC also allows any other scheme of subcarriers which do not interfere with the operation of BTSC receivers or normal monophonic sound television receivers. When not transmitting stereo or SAP, broadcasters may elect to use the aural baseband for subcarriers for non-program related purposes. Virtually any kind of subcarrier incorporating digital or analog signals of program material or data may be transmitted using AM, FM or other modulation methods. Subcarriers may be placed in the aural baseband between 16 and 120 kHz. The maximum deviation of the aural carrier by other subcarriers is limited to 50 kHz. Any number of subcarriers may be used as long as their total modulation of the main aural carrier does not exceed 50 kHz. The total modulation of the main carrier including main channel sound and non-BTSC subcarriers may not exceed 75 kHz.

Stations should avoid placing other subcarriers within ±15 kHz of the SAP frequency (5H or 78.671 kHz). If the SAP frequency is to be used for non-program related material, the injection level of the subcarrier should be kept below the SAP threshold of the receiver or about 3 kHz deviation. The use of a single subcarrier with 50 kHz deviation should be avoided because of the potential of crosstalk into the main channel in some receivers unless extensive compatibility testing is conducted.

MODULATION SUMMARY

The FCC permits the BTSC specified multichannel sound signal to modulate the main aural carrier a total of 73 kHz. This is produced by the combined monaural and stereo subcarrier deviation of 50 kHz, the pilot subcarrier deviation of 5 kHz, the SAP subcarrier deviation of 15 kHz and the professional use subcarrier of 3 kHz for a grand total of 73 kHz. See Figs. 8 and 9.

Other subcarrier arrangements may modulate the main carrier by as much as 50 kHz. With the main channel at 25 kHz and subcarrier at 50 kHz, the total may not exceed 75 kHz deviation.

PERFORMANCE SPECIFICATIONS

In 1984 the FCC began a series of rulemaking procedures in which many of the technical rules in part 73 of the FCC Rules and Regulations were deleted if they described "quality" characteristics

SIGNAL	MONITORED PARAMETER
MAIN CHANNEL (L+R)	25 kHz PEAK DEVIATION
STEREO CHANNEL (L+R)	50 kHz PEAK DEVIATION
MAIN AND STEREO	50 kHz PEAK DEVIATION
PILOT CARRIER INJECTION	5 kHz PEAK DEVIATION
SAP MODULATION	10 kHz PEAK DEVIATION
SAP INJECTION	15 kHz PEAK DEVIATION
OTHER SUBCARRIER	3 kHz PEAK DEVIATION
TOTAL COMPOSITE	73 kHz PEAK DEVIATION

Fig. 8. Typical modulation monitor functions.

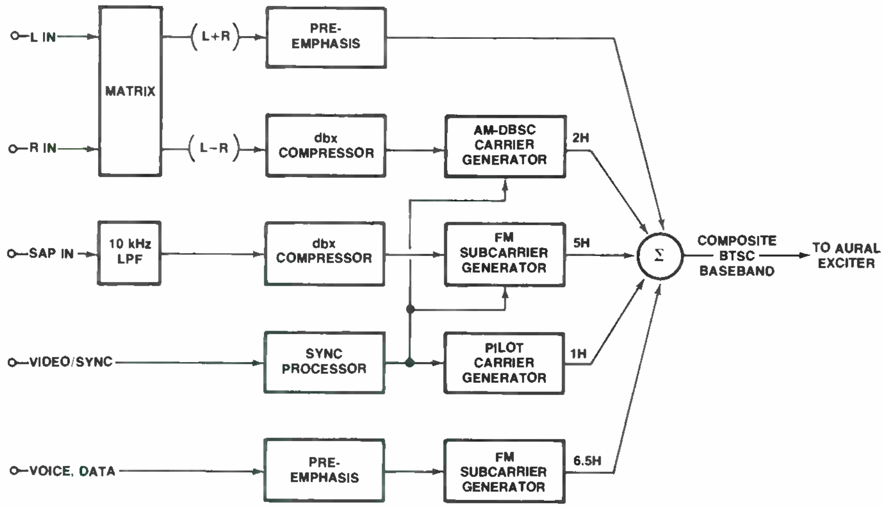


Fig. 9. Typical MTS transmitter encoder.

of the transmitted signal. Hence, the Rules requiring stations to meet certain distortion, frequency response and signal-to-noise performance levels have been eliminated in favor of marketplace pressures to maintain quality levels. Therefore, there are no rules which require minimum audio performance levels. Rules which act to control interference, however, have been maintained where needed.

The Rules which permit multichannel television sound are no different and the user must consult either the manufacturers specifications or the original laboratory test results to determine the sound quality capability of the TV stereo system. The FCC considers new services differently from well established radio broadcast systems that have a track record for quality. Because TV MTS is *new*, several quality specifications are included, not in the Rules themselves, but in a separate document—OST #60, a Technical Bulletin published by the FCC Office of Science and Technology. However, the implementation of MTS is relatively new and the broadcast and receiver industry have produced a series of recommended operating practices, many of which are cited in this chapter.

Because the transmission of stereo sound for television is technically similar to that of FM broadcasting, the reader is encouraged to review the chapter on FM broadcasting and stereophonic transmission elsewhere in this handbook for background information. Portions of the BTSC MTS system description are also contained in the Television Transmitter chapter.

Fig. 10 shows the general performance characteristics of TV MTS which should be achievable under ordinary conditions. While most of the performance objectives indicate the potential for high audio quality, the stereo separation figure may be of some concern to those familiar with the much higher performance of the FM stereo system. The 30 dB figure should be easily acheived in practice. However, it is relatively low because of the use of the companding system and the critical adjustments which must be made in the companding system and the more sophisticated circuits involved compared with the simpler circuits which can produce better than 40 dB separation in FM stereo operation. The tracking gains of the mono and stereo channels determines the relative separation between channels (see Fig. 15a). While it may be possible to maintain a difference of a few tenths of a dB in gain in the sum and difference channels in FM stereo transmission, it is extremely difficult to obtain better than about 0.5 to 0.8 dB in TV MTS. This is because the stereo channel is compressed by the noise reduction system by as much as 20 dB which magnifies the difference when received and decoded in the TV receiver. Improvements in circuit stability in the future may permit an increase in stereo

SIGNAL	RESPONSE	S/N	DIST
MAIN CHANNEL	50-15 kHz	>58 dB	<2.5%
STEREO SUBCHANNEL	50-15 kHz	>55 dB	<1.0%
SAP CHANNEL	50-10 kHz	>50 dB	<3.0%
PROFESSIONAL CHANNEL	100-3 kHz	>40 dB	<3.0%
STEREO SEPARATION	>30 dB		

(Grade B or better TV signal)

Fig. 10. MTS performance objectives.

separation, perhaps as much as 40 dB on a regular basis.

It should be noted that in actual operation there is no "stereo noise penalty" in TV MTS as there is in FM stereo. This is due primarily to the use of the noise reduction system. Further, there is no reduction in deviation of the mono signal during BTSC stereo transmission as there is in FM stereo operation. In FM stereo the main or mono channel must be reduced by 8 to 10% to permit the addition of the pilot subcarrier. This is not the case with TV MTS.

The crosstalk performance values are probably the single most important technical parameter with which the broadcaster must be concerned. Crosstalk less than those values shown in Fig. 11 will be caused by systems external to the stereo transmission equipment itself. Non-linearities in the TV aural exciter and power amplifier stages, combining networks, antenna switching components and the antenna itself constitute the main areas in the station where crosstalk can be generated. In general, the crosstalk into the stereo signals from the SAP will be less than crosstalk from the main channel into the SAP signal. As long as these levels do not become excessive, most crosstalk will be concealed by the companding system.

Other sources of crosstalk are multipath distortion due to poor VSWR of the transmission line and antenna, multipath during the propagation of the transmitted wave and improper matching of the antenna and receiving system. In general, the expander can conceal noise and crosstalk in the stereo and SAP channels by as much as 30 dB.

CROSSTALK	OBJECTIVES
MAIN INTO STEREO	>40 dB
STEREO INTO MAIN	>40 dB
MAIN INTO SAP	>50 dB
SAP INTO MAIN	>60 dB
OTHER SUB INTO MAIN	>60 dB
OTHER SUB INTO STEREO	>60 dB

Fig. 11. Crosstalk objectives.

INCIDENTAL CARRIER PHASE MODULATION

The control of incidental carrier phase modulation (ICPM) in the visual transmitter is the single most important performance characteristic of the transmitter that can affect the quality of multichannel sound. ICPM will cause buzz in intercarrier sound receivers. Intercarrier type receivers will be used for multichannel sound reception because of their high immunity to "common mode" phase modulation found in various systems in the path of the television signal before it reaches the receiver. Such common mode phase modulation can occur in TV translators, MATV downconverters, and CATV headend and set-top converter equipment. In intercarrier type receivers, the aural and visual carriers beat together in the video detector to produce the 4.5 MHz sound signal. Because common mode phone modulation affects both carriers, the phone modulation is factored out of the 4.5 MHz carrier.

On the other hand, phase modulation of the visual carrier in the transmitter which is independent of the aural carrier, will produce undesired phase modualtion of the 4.5 MHz in the receiver and the result will be the familiar buzz. Phase modulation in the visual transmitter can be caused by hum in the carrier frequency oscillator, non-linearities in the IF and power amplifier stages, and especially the visual modulator. UHF transmitters which use klystron anode pulsers to achieve higher operating efficiencies may have phase modulation caused by the pulser modulating the visual amplifier power supply at a 60 Hz rate and a 15 kHz rate. The higher frequency is not heard by most viewers but the lower frequency is heard as a raspy buzz. Pulsing the visual amplifier may cause crosstalk through the common power supply into the aural amplifier. The resulting 15.734 kHz modulation, if high enough, may cause false indications of the stereo indicator on MTS television receivers as will high ICPM.

ICPM in visual transmitters should be adjusted to be less than 3 degrees during luminance and to less than 5 degrees during sync. See Fig. 12.

ICPM manifests itself as differential phase in the demodulated video. Measurement instrumentation requires a TV demodulator with both envelope and quadrature video detectors and a waveform monitor with proper graticules. After measuring ICPM or listening for buzz on a stereo television receiver to determine if ICPM is present, make the necessary phase adjustments and retune RF stages as appropiate. Adjustment of

DEGREES ICPM ⟶	1°	2°	3°	4°	5°
BASEBAND BUZZ LEVEL IN dB BELOW 25 kHz DEV. ("WORST CASE")	−50	−44	−40	−38	−36
STEREO SBR dB	56	50	46	44	42
SAP BUZZBEAT THD dB	−49	−43	−39	−37	−35

$$\text{NYQUIST SLOPE EQV. ICPM} = \begin{cases} 2.4° \text{ AT (L-R) SUBC.} \\ 6° \text{ AT SAP SUBC.} \end{cases}$$

Fig. 12. Intercarrier buzz and buzzbeat levels. (Courtesy J.J. Gibson, RCA)

ICPM will affect video differential phase. Once ICPM is corrected, adjust video correction circuits to compensate for differential phase.

ICPM will be more noticeable on television receivers with multichannel sound capacity. Although some buzz may be heard in the main channel, the ICPM will affect the subcarriers to a greater extent. In addition, the improved audio quality in the new multichannel receivers will make any existing buzz more noticeable even when transmitting monaural sound. Therefore, broadcasters not contemplating multichannel transmission should investigate the ICPM condition of their transmitters with an ear toward improving the quality of the mono sound on both stereo and mono receivers.

COMPATIBILITY WITH EXISTING RECEIVERS

During the testing of the BTSC multichannel sound system considerable effort was spent evaluating the effect of a fully loaded BTSC system on existing monophonic television receivers. Extensive compatibility testing of a wide range of receiver makes and models, under extreme transmission conditions revealed virtually no significant degradation to either the sound or picture of the desired or adjacent channel, as noted by trained observers.

Modern monophonic television receivers should not experience any significant degradation when receiving a TV MTS signal. In fact there may actually be a noticeable improvement in the sound quality on a monophonic receiver due to the increased attention paid to the stereophonic sound signal production and processing and lower visual transmitter ICPM. This desirable side benefit should also offset the slight reduction in monophonic modulation level which will result when transmitting a mostly left or right channel sound. Because the main channel is the sum of left plus right, a right or left-only signal will produce 6 dB less modulation than the sum signal. However, producers seldom employ left or right-only signals and as a result the actual reduction will be something between 0 and 6 dB. It could be argued that these several dB could reduce the apparent "loudness" of the stereo station when compared to a monophonic station. The argument will cease as more television stations install stereo transmission facilities.

COMPATIBILITY WITH CATV SYSTEMS

During the testing of the TV MTS system considerable attention was paid to the ability of the MTS signal to pass through MATV and CATV systems. Cable headend, distribution and subscriber equipment was used to conduct the various tests. A concern that the upper adjacent visual channel might be affected by the wider MTS modulated aural carrier of the lower channel was not confirmed. In general, those CATV systems which process off-air signals in an RF mode (that is, not demodulated) will have little difficulty in successfully processing and passing an acceptable MTS signal through their system.

Cable systems with demodulating type headend and/or subscriber equipment may not be able to pass MTS signals without substantial degradation to the subcarriers due to the use of low pass filters in the audio portions of the demodulator facilities. However, it can be expected that equipment manufacturers will accommodate the BTSC system in future designs.

The BTSC MTS signal should also pass through most amplitude modulated link (AML) microwave systems used by many cable companies for wide area distribution of multiple channels without significant degradation.

COMPATIBILITY WITH TELEVISION TRANSLATORS

Television translators using heterodyne or direct RF conversion techniques should not cause significant degradation to MTS signals if the translator is properly maintained. However, some translators receive little attention unless they fail completely. In general, unless the picture and sound are severely degraded, or visible or audible interference is present due to improper operation of the translator, then the MTS signal will probably pass through the system with little degradation.

Remodulating type translators however, may require modifications to accommodate the MTS signal. A wider audio bandpass, to accept the subcarriers, and improved visual transmitter oscillator stability and low ICPM are some of the improvements required to properly retransmit the MTS signal.

Fig. 13 shows what kinds of transmission impairments may cause various audio impairments.

OPERATING PRACTICE

Fig. 2 shows the composite baseband signal that is used to modulate the aural carrier. In addition to the main, pilot and stereo subcarriers there is the SAP and professional use subcarriers which are discussed in other sections. The main channel contains the sum of the left and right or "sum" audio which may be received without

EFFECT / CAUSE	BUZZ	SAP BUZZ BEAT	NOISE & NOISE PUMPING	REDUCED STEREO SEP	DIST.	X-TALK
NON-LINEAR PIX	✓					
NON-LINEAR SOUND					✓	✓
MISMATCHES						✓
DIPLEX FILTER					✓	✓
ICPM	✓	✓				
MULTIPATH	✓	✓	✓	✓	✓	✓
WEAK SIGNALS			✓	✓		
CO-CHANNEL					✓	✓
MISTRACKING				✓		
REC. FILTERS	✓	✓		✓	✓	✓
FM DETECT	✓				✓	✓

Fig. 13. Effects of transmission impairments. (Courtesy J.J. Gibson, RCA)

degradation on ordinary monophonic receivers. The left minus right or difference channel is contained on a double sideband, suppressed carrier amplitude modulated carrier at two times the TV horizontal rate (2 X 15.734 kHz = 31.568 kHz). The subcarrier is suppressed to avoid wasting modulation capability. The energy in the sidebands however, is permitted to deviate the main aural carrier by as much as 50 kHz.

In order to properly decode the signal it is necessary to reinstate the carrier in the receiver on the exact frequency and nearly the exact phase as the original. To accomplish this a "pilot" signal is transmitted at a frequency of the TV horizontal rate (15,734 Hz). The pilot modulates the main carrier with a deviation of 5 KHz.

The pilot carrier also carries phase information with it for proper demodulation of the stereo subcarrier. The phase of the pilot carrier wave zero crossings in relation to the stereo subcarrier zero crossings should be within 3 degrees. No such tolerance is specified in the FM broadcast rules.

The pilot carrier must be protected from the main audio so as not to contaminate it upon reception which would degrade the ability of the receiver to produce high quality results. A fairly sharp low pass filter must be used in the transmitter to cut off audio frequencies above 15 kHz and provide a notch at 15.734 kHz.

Between 25 and 30 dB protection is considered adequate. Above 15.734 kHz the response should roll off rapidly to 50 to 60 dB in order to protect the stereo difference channel which contains sidebands between 16.5 kHz and 46.5 kHz. Fig. 14 shows the response of a typical low pass filter for use in television stereophonic sound transmission facilities.

The pilot is transmitted at a deviation of 5 kHz ±0.5 kHz. The receiver pilot detector threshold is nominally set to activate the stereo circuits at aural carrier deviation levels of about 2.5 kHz. To provide the receiver with an ample pilot level, it is essential that modulation products at 1H, from whatever source, be reduced to less than 1 kHz aural carrier deviation, or about 30 dB below maximum 25kHz deviation. In *monophonic* transmitters, the band in which the pilot would be located can easily be contaminated by:

1. Audio program material at 15.734 kHz;
2. Horizontal rate crosstalk from the visual transmitter; and
3. RF components from the visual transmitter passed through the aural and visual combining networks at the output of the transmitters.

For example, a simple matter of a studio microphone picking up the "screech" of a horizontal oscillator from a nearby video monitor can produce enough modulation at 1H to produce a false indication in a BTSC equipped receiver. Special

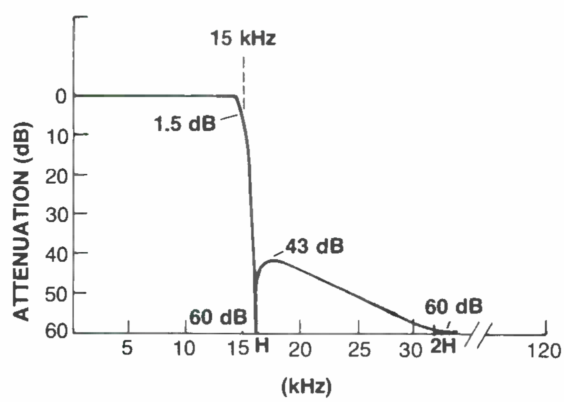


Fig. 14. Audio input low pass filter.

precautions are taken in transmitters designed specifically for multichannel sound operation to protect the pilot. Operators of monophonic transmitters should also check their facilities for residual 1H signals in either the aural transmitter or in the combining network at the output of the transmitters. Monophonic audio inputs to transmitters should incorporate a 15 kHz low pass filter.

For good stereo operation it is necessary that the left and right channels remain well separated before, during and at the end of the transmission chain. The amplitude and phase of the mono and stereo signals must be nearly identical and the phase of the pilot carrier must be maintained. Channel separation as a function of these three factors is given in the following equation:

Separation in dB =

$$20 \log_{10} \left[\frac{(\cos\theta + \frac{S}{M}\cos\phi)^2 + (\sin\theta)^2}{(\cos\theta - \frac{S}{M}\cos\phi)^2 + (\sin\theta)^2} \right]^{1/2}$$

Log10 [" " "] 1/2

Where:

M = the gain of the main (L + R) path and
S = the gain of the stereo (L − R) path
ϕ = the phase error of the reinserted 38kHz subcarrier that is twice the phase error of the 19kHz pilot carrier
θ = the difference in phase between the (L + R) and (L − R) paths.

The implementation of multichannel sound in a television facility which has not dealt with stereo or multiple audio channels in the past will find the experience to be a challenge. Virtually all audio equipment in the station will need to be reviewed, reworked, rewired, or replaced when adding new audio channels. Video tape machines, routing switchers, master control switching and monitoring, studio audio facilities and the STL will require close examination to determine how to implement extra sound channels. Video tape equipment with two sound channels may require new test procedures to determine that the phase relationships between channels is correct.

Studio to transmitter links will need additional channels for stereo and/or SAP audio. Consideration should be given to a composite BTSC signal for transmission over the STL. Special attention must be paid to audio processing. There will be a tendency towards maximizing the loudness of the signal rather than to moderate processing which will result in a more acceptable stereo sound quality.

Multichannel sound equipped television receivers have substantially improved audio system

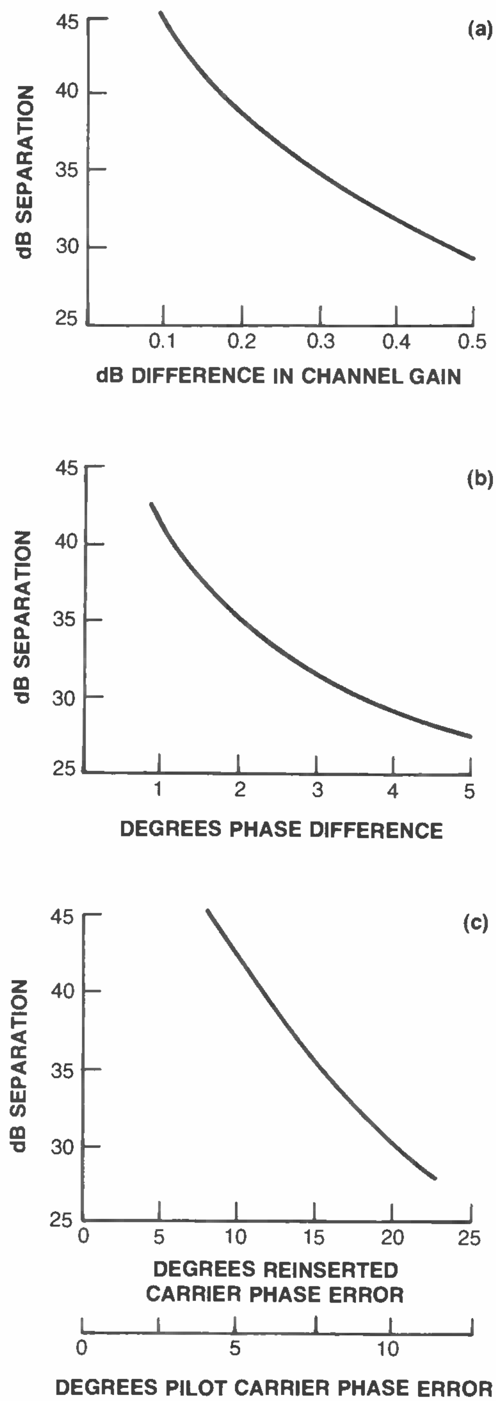


Fig. 15. Stereo separation versus (a) amplitude difference between main and stereo channels, (b) phase difference between main and stereo channels, and (c) error in phase of pilot or reinserted carrier.

which will not require "loudness", but will emphasize audio faults now concealed by most television receivers. As a result, special attention must be devoted to the audio transmission facilities of all television stations, whether transmitting monophonic or multichannel sound.

3.7

Transmitter Control Systems

Fred A. Barbaria and Daniel L. Barnett
Moseley Associates, Inc.
Goleta, California

INTRODUCTION

One of the standard operating techniques used at broadcast stations is the remote control of transmitters. In 1953, the Federal Communications Commission (FCC) first authorized nondirectional stations up to and including 10 kW to operate by remote control. During 1957, the Commission, at the request of the National Association of Broadcasters (NAB), amended its Rules and extended remote control privileges to all types of standard AM and FM broadcast stations regardless of power or mode of operation. In 1963, the Commission, on its own motion, amended Part 73 of the Rules to permit the remote control of UHF television transmitting facilities. In 1971, again at the request of the NAB, the Rules were once again amended to permit the remote control of all television transmitting facilities regardless of frequency or type of operation. This enactment by the FCC permits remote control of all types of broadcast transmitters regardless of the service. In 1977, the FCC amended the Rules to allow control of AM (nondirectional) and FM stations with Automatic Transmission Systems. In 1984, the FCC amended the Rules to combine all classes of service under a uniform code and to relax the requirements of remote control. The history of FCC rule-making essentially has followed the history of transmitter plant reliability.

Several motives exist for using Transmitter Control Systems. The first is to allow unmanned operation at the transmitter site. Transmitter Control Systems allow an operator to activate and deactivate a transmitter from a distant control point. They relay required transmitter operating parameters to the control point. The systems allow the operator to make adjustments to the transmitter from the control point to ensure correct operation of the transmitter. Furthermore, the systems provide remote indications concerning security and environmental conditions at the unmanned transmitting plant.

The second motive for using transmitter control systems is to assist the operator at a control point in performing his duties correctly. Recent Transmitter Control Systems have digital displays, can be calibrated in engineering units, and have built-in power-to-linear and other conversions, allowing the operator to read the transmitter operating parameters directly, without interpolation or chart look-ups. In addition to displaying values, alarm conditions can be programmed into most transmitter control systems. For example, telemetry tolerance alarms can be provided to warn the operator that the transmitter is not operating within its specified limits. Status alarms can be provided by the system to notify the operator that something is amiss at the transmitter plant. To enhance system observation further, CRT displays may be used to provide an overview of the transmitter system. Diagnostics may be provided to the operator to help determine a cause of failure in the transmitter or the control system. Finally, automatic logging may be implemented to record routine information for trend analysis, and may initiate event-triggered logs to record transients and document emergencies of the transmitter plant. The log can also form the basis of a transmitter log to meet company or station requirements and can be provided to the FCC in the case of operating complaints against the station.

The third motive for using transmitter control systems is to automate certain functions of the operator. Automatic Transmitter Systems (ATS) provide the first of two methods for automation. These systems allow completely unmanned operation, providing a means to turn the transmitter on and off still exists. The ATS will monitor and adjust antenna input power for AM (or transmitter output power for FM) and modulation levels according to a rigid formula spelled out in the rules. The ATS will also take the transmitter off the air in case of failure or deviation from the acceptable operating limits.

Automatic Control Systems provide the second method of automation. These systems are intended to be adjuncts to standard remote control systems. Control ladders may be set up to perform sequences of action, and the systems may be used to perform trims, i.e. take action to keep particular telemetry values within specified tolerances. Operation of these sequences may be triggered on day, date, time, trouble conditions, operator activation, or other trigger conditions accessible via the remote control system.

CURRENT FCC REGULATIONS CONCERNING TRANSMITTER SYSTEMS

The December 1984 FCC Rules changes significantly altered the character of the sections pertaining to the Remote Control of transmitters. The intent of these changes was to "place responsibility on each licensee to assure compliance with the station authorization." As a result, the specific details of remote control operation have been left for the most part to the station licensee. This should allow maximum flexibility in meeting the requirements of remote control operation.

The key requirement of the new Rules is "Broadcast stations operated by remote control must provide at remote control points sufficient control and operating parameter monitoring capability to allow technical operation in compliance with the Rules applicable to that station and the terms of the station authorization." What is "sufficient" is left up to the licensee, and leaves room for individuality of station operation.

The failsafe requirements of the new Rules state that if a malfunction occurs which causes a loss of accurate meter readings, remote control operation may not continue beyond three hours after the malfunction is detected. Also, if the station is operating out of its authorization, and the problem cannot be solved by remote control, then operation must be terminated immediately. The telemetry failsafe function of remote control systems meet the first requirement above. If metering of particular channels vanishes or if the Control Terminal does not receive communications from the Remote Terminal then an alarm will occur to alert the operator that corrective action must be taken within three hours.

Requirements for television operation include off-air visual demodulators, and modulation and waveform monitors. Further information may be found in this chapter under "Elements of a Transmitter Remote Control System," and elsewhere in this handbook.

ATS failsafe requirements are somewhat different than those of remote control systems. An ATS failsafe is required to terminate transmission when any of the following conditions occur for a 3-minute period:

1. The automatic power-adjustment circuit fails.
2. The automatic modulation-adjustment circuit fails.
3. The mode-switching time clock fails (AM only).
4. The monitoring and alarm point cannot communicate with the ATS.
5. Any of the alarm systems functions fail, or
6. Sampling function are lost.

Be sure to consult the latest FCC Rules before making any changes or additions to your station operation. The sections applicable to transmitter control systems are:

AM ATS

73.140 Use of Automatic Transmission Systems (ATS)
73.142 Automatic Transmission Facilities
73.144 Fail-Safe Transmitter Control for Automatic Transmission Systems
73.146 Automatic Transmission System Monitoring and Alarm Points

FM ATS

73.340 Use of Automatic Transmission Systems (ATS)
73.342 Automatic Transmission Facilities
73.344 Fail-Safe Transmitter Control for Automatic Transmission Systems

73.346 Automatic Transmission System Monitoring and Alarm Points

Non-commercial Educational FM ATS

73.540 Use of Automatic Transmission Systems (ATS)

73.542 Automatic Transmission Facilities

73.544 Fail-Safe Transmitter Control for Automatic Transmission Systems

73.546 Automatic Transmission System Monitoring and Alarm Points

Remote Control

73.1400 Remote Control Authorizations

73.1410 Remote Control Operation

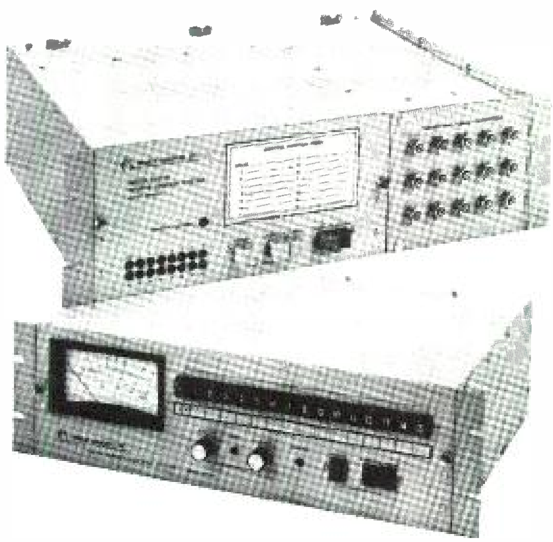


Fig. 1. Analog remote control system.

BASIC THEORY OF OPERATION

There are essentially two kinds of remote control systems, analog and digital. Early remote control systems were analog designs, i.e. their circuitry used a continuous range of voltage levels to operate. A typical analog remote control system is shown in Fig. 1. Its block diagram is shown in Fig. 2. Recent remote control systems are digital designs, i.e. their circuitry uses discrete voltage levels to operate. A binary (two-level) scheme is usually used. A typical digital remote control system is shown in Fig. 3; a typical digital block diagram is shown in Fig. 4. In analog systems, meters are used to take readings. Meters have inherent limitations in accuracy, generally limited to about 2%. Thus analog systems are also limited in their accuracy to about 2% which is the requirement under FCC rules. In digital systems, multi-digit displays are used to take readings. The accuracy of digital systems is theoretically limited only by the number of digits displayed, and the number of bits (binary pieces of information) used internally to convert the analog inputs to digital data. Accuracy of 0.5% or better is available with today's remote control systems. Calibration of analog systems typically requires two people, one at the transmitter tweaking the calibration potentiometer on the remote control system, and the other watching the meter from the studio, telling the first when to stop. In addition, daily calibration from the studio is often required to maintain specified accuracies. Calibration of digital systems, however, only requires one person. This is accomplished at the transmitter site by entering a series of keystrokes

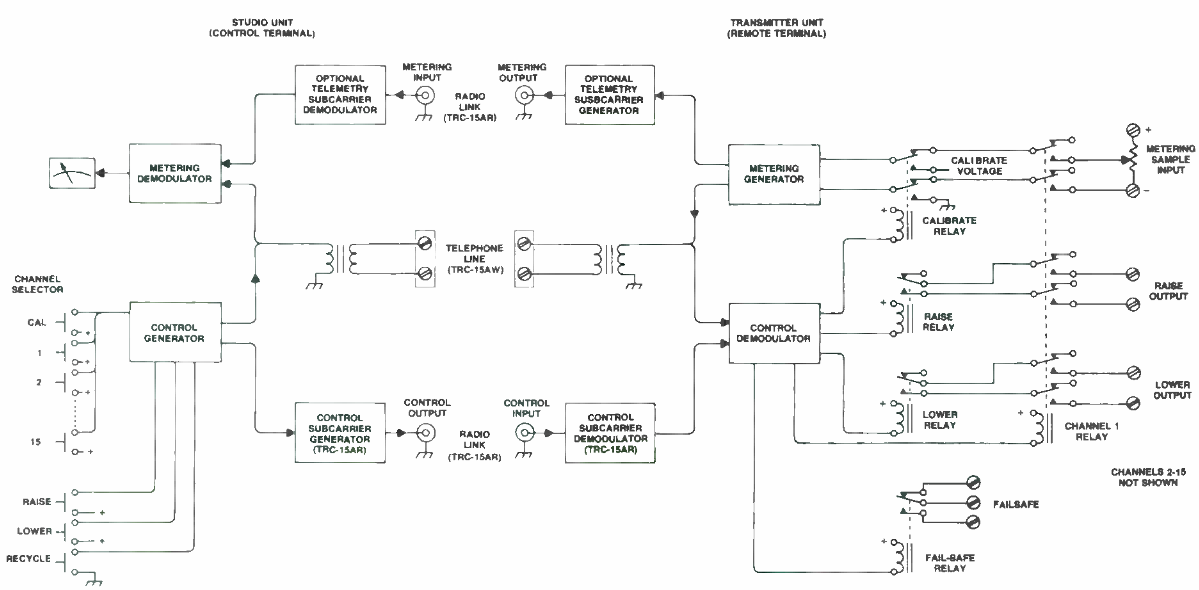


Fig. 2. Analog remote control block diagram.

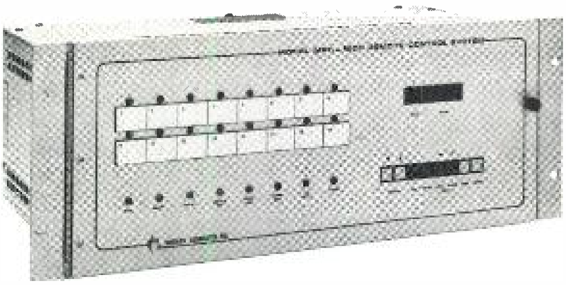


Fig. 3. Digital remote control system remote terminal.

on the front panel of the remote control system. Daily calibration is not required to maintain system accuracy. In addition, power-to-linear and other conversions are usually built in, eliminating the need for some external interface boxes. Communication between the studio and transmitter units of an analog system is usually accomplished with a series of tones; the duration or frequency of the tones determines the action to be taken. In digital systems, the tones take the form of digital words so that, in essence, the studio and transmitter units talk to each other in sentences. Much of the added flexibility of digital remote control systems has resulted from the use of microprocessors in their design.

Microprocessors can be found in almost any recent digital design of any complexity, and transmitter control systems are no exception. A microprocessor is a machine which executes instructions, usually stored in some form of ROM (Read Only Memory), which modify the contents of RAM (Random Access Memory), or which perform some I/O (Input/Output) function (see Fig. 4). By using microprocessors, designers can specify an arbitrary set of logic instructions (software) that will most readily accomplish the task at hand. In our case, that task is the (remote) control of transmitters. A microprocessor is usually the heart of each main component in a transmitter control system.

The part of the transmitter control system that interfaces directly to the transmitter can be called data acquisition and command. This is the I/O between the system and the transmitter. Data acquisition includes telemetry (analog) inputs, status (on/off) inputs, and binary or BCD (Binary Coded Decimal) inputs. Commands are the raise and lower outputs to the transmitter. Other outputs to the transmitter plant can include analog or digital (binary or BCD) outputs.

User setup (calibration) is mostly performed from the Remote Terminal front panel located at the transmitter plant. The setup parameters for each channel can include analog input mode (millivolt, linear, power, indirect, etc.) analog limits (one or more sets, upper and lower), status attributes (invert, latch), status alarms (rising edge, falling edge), and command output type (momentary, pulsed, latched). Setup parameters must be protected from power outages and several methods can be used. Battery backup is usually fairly inexpensive, but has a limited backup time (a few hours). Disk drives hold large amounts of data, but have moving parts that must occasionally be serviced. EEPROMS (Electrically Erasable Programmable Read Only Memories) have long retention times (about 10 years) and take up very little room, since they are integrated circuits.

Communication between the studio and transmitter can either be full-duplex (both ends talking at the same time) or half-duplex (one end talking at a time). Because of bandwidth limitations of phone lines, the rate of communication is usually limited to 300 baud for full-duplex and 1200 baud for half-duplex. The medium between the studio and the transmitter may be dedicated phone lines or subcarriers on an STL. The medium between the transmitter and the studio may be a dedicated phone line, a TSL, or a subcarrier or subaudible channel on the main carrier. Subaudible is not recommended for digital systems because only very limited bandwidth is available for the data.

All day-to-day operation is done at the studio from the Control Terminal. The operator may take telemetry and status readings, and make adjustments to the transmitter using the appropriate raise or lower command as required. Optionally, commands may be initiated via a direct command panel. Alarms generated by the system may be acknowledged by the operator and/or recorded by the automatic logger. In addition, the operator may overview the transmitter control system with a CRT display.

ELEMENTS OF A TRANSMITTER REMOTE CONTROL SYSTEM

A transmitter control system, whether it be for radio or television transmitter control, consists of several key elements. These key elements are the operator, who is normally located at a studio or master control room location; the Control Terminal sometimes called a ("studio unit"), located adjacent to the operator, the Remote Terminal (sometimes called a "transmitter unit"), located adjacent to the equipment being controlled; and last, the equipment actually being controlled, which includes a broadcast transmitter and other equipment located at the transmitter plant, which is often called the "remote site".

The transmitter control system provides the necessary interfacing between these four key elements to allow the operator to properly monitor and control the transmitter plant,

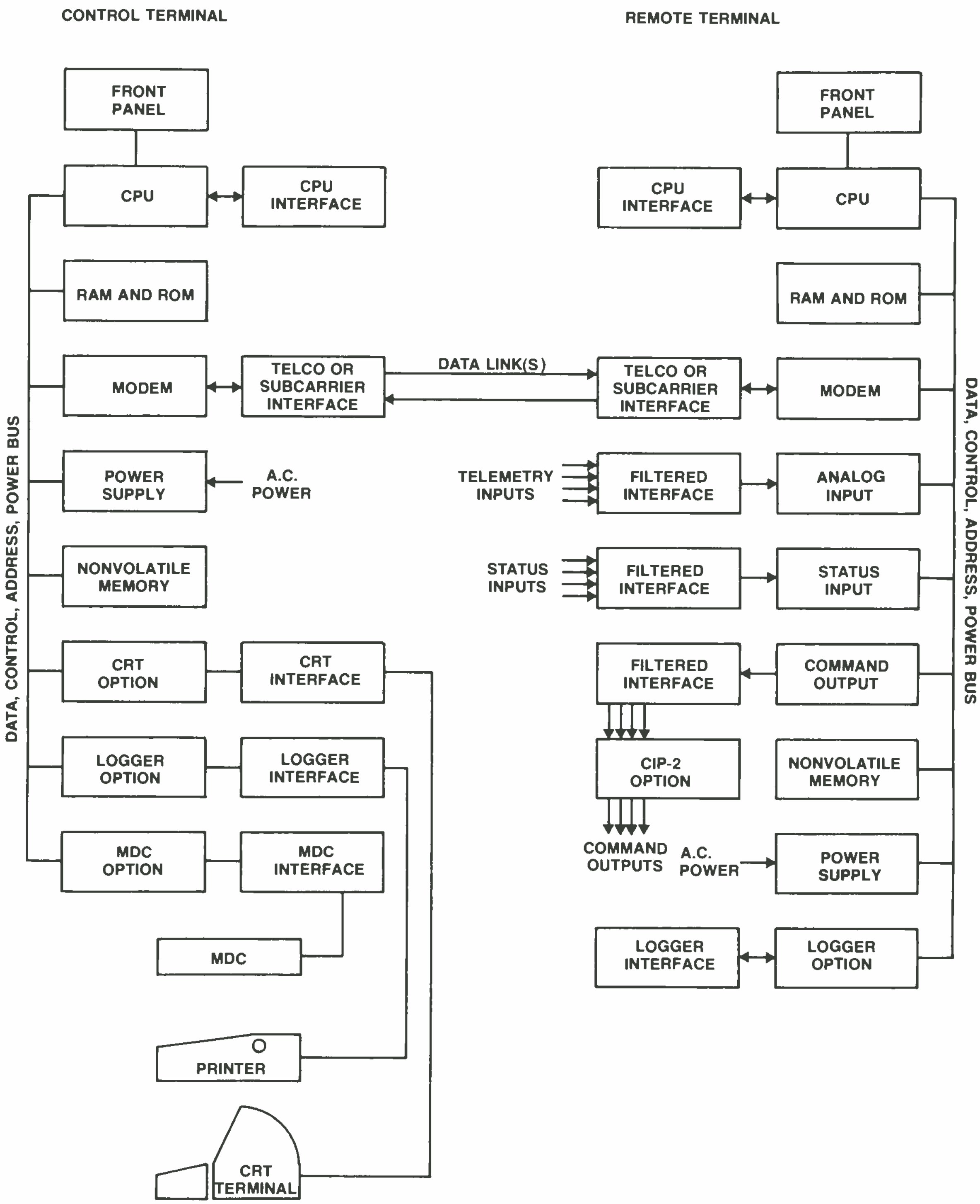


Fig. 4. Digital remote control block diagram.

whether it is on the other side of the building or miles away at a mountaintop transmitting site.

Control Terminal

The Control Terminal, located at the main control point, normally consists of a rack-mounted unit. Since most present-day remote control systems are microprocessor-based, the Control Terminals have become quite small, on the order of two to four rack units high. This minimizes the transmitter control system space requirements in the control room, where space is often at a premium. A typical Control Terminal, such as the one shown in Fig. 5, normally includes digital displays for telemetry and perhaps status monitoring as well as keyboards for calling up the individual telemetry and/or status channels and issuing commands to the transmitter plant. Modern transmitter control systems usually include the ability to program tolerance limits on telemetry channels so that the operator does not have to continually monitor those channels for compliance with the FCC Rules and Regulations.

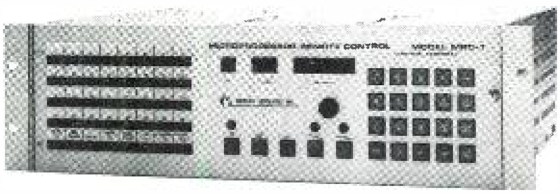


Fig. 5. Remote control system control terminal.

Aural transmitter power output, for example, is often monitored automatically for tolerance limits. Similarily, status channels (on/off) may be monitored for changes in states that require alerting the operator. Smoke detectors, intrusion alarms, and transmitter overload indications are typically set up within the transmitter control system for operator alarms. The benefit of having the transmitter control system monitor both telemetry and status alarms is that the control room operator or disc jockey can concentrate on their other duties while being assured that the transmitter system is operating within the predetermined limits programmed into the remote control system, and that the transmitter plant buildings and equipment are inaccessible by unauthorized persons.

The typical interface between the operator and the Control Terminal is via the Control Terminal front panel, with its digital displays, LEDs, and keyboard. Accessory CRT terminals are often available for transmitter control systems to simplify the interface between the operator and the control terminal. Some CRT displays, such

Fig. 6. CRT display.

as the one shown in Fig. 6, include alphanumeric text capability. This text is normally programmed on site by the user, so that it can be changed if desired. On some systems, multiple CRT terminals may be installed at alternate control points within a studio complex to allow transmitter control to be shifted between control rooms during the broadcast day. The current FCC rules regarding control points for broadcast transmitters should be studied carefully before planning such a system, however.

An automatic logging option is available for many transmitter control systems. This option prints selected (or all) parameters on a printer at predetermined intervals. Most systems also will print the parameters if an out of tolerance telemetry condition or status alarm occurs. Some systems also allow the user to program heading text that will print at the top of each log page. A typical remote-control-system-generated log page is shown in Fig. 7.

Multiple direct command panels are available for some remote control systems; these panels allow the operator to access a given control function by pressing a single button. Similarly, remote status panels are also available for some advanced remote control systems. If access to multiple direct command panel buttons as well as the status panel indicator lamps is provided by the manufacturer, a relatively inexpensive "mimic" panel can be built by station personnel, customizing the installation to the station's transmitter plant configuration. Automatic control units, described in a previous section, are also available for some Control Terminals.

Some advanced transmitter control systems support more than one Control Terminal for those installations that have widely separated control points. Control of the transmitter may be

MRC-1600 Automatic Logging Option

MOSELEY ASSOCIATES, INC.
111 Castilian Drive
Goleta, CA 93117

The header, logged channels, and logging intervals are all user programmable.

TIME	Filament Voltage	Plate Voltage	Plate Current	Output Power	Refl. Power	Intake Temp.	Exhaust Temp.	Alarm System	Tower Lights
08:00	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
08:30	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
09:00	6.2	7.43	2.12	98.8	97.	27.5	56.9	On	On
09:30	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
10:00	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
10:30	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
11:00	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
11:23	ALARM - Status Channel 14: *Off								
11:24	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	*Off
11:30	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
12:00	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
12:30	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
13:00	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
13:30	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
14:00	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
14:30	6.2	7.43	2.12	98.7	97.	27.5	56.9	On	On
14:46	ALARM - Analog Channel 04: 105.1>								
14:47	6.6	7.93	2.12	105.3>	97.	27.5	56.9	On	On
15:00	6.6	7.93	2.12	105.3>	97.	27.5	56.9	On	On
15:30	6.6	7.93	2.12	105.3>	97.	27.5	56.9	On	On
16:00	6.6	7.93	2.12	105.3>	97.	27.5	56.9	On	On
16:30	6.6	7.92	2.12	105.2>	97.	27.5	56.9	On	On
17:00	6.6	7.92	2.12	105.2>	97.	27.5	56.9	On	On
17:30	6.6	7.92	2.12	105.2>	97.	27.5	56.9	On	On

Fig. 7. Automatic logger output.

passed back and forth between the Control Terminals. Again, current FCC Rules regarding control points for broadcast transmitters should be studied carefully before planning such a system.

Remote Terminal

The transmitter control system Remote Terminal, although tucked away in a rack at the transmitter site, forms the heart of the remote control system. It must be able to withstand high intensity RF fields, and "dirty" electrical power, yet still properly relay commands from the control point as well as convey telemetry and status signals from the transmitter plant to the studio. It is extremely important that the Remote Terminal be installed with adequate ventilation, which is a requirement for any rack-mounted equipment. In addition, the rack must be properly grounded to minimize RF interference (RFI). Lightning is sometimes a problem with transmitter control systems, especially as telephone lines are often used to interface to the Control Terminal. It is recommended that both telephone lines and the ac mains be installed underground for the last several hundred feet at the transmitter plant. In addition, lightning arrestor equipment should be used to protect both the ac mains and incoming telephone lines. Most Remote Terminals include some surge protection, but this protection is not intended to dissipate the enormous amounts of power generated by a direct lightning hit or even a near miss.

Newer radio and television transmitters include facilities for supplying telemetry samples, status closures, and command inputs to the transmitter. Accessory sampling and control kits, described later in this chapter, are available for assisting in interfacing older transmitters to remote control systems. The transmitter and/or remote control system manufacturer should be able to recommend interfacing techniques for transmitters that were not initially designed for remote control operation. Even some of the newer transmitters may require interface accessories to properly connect to the transmitter control system selected.

Telemetry (or analog or metering) samples into the Remote Terminal should normally be in the range of 1 to 4 Vdc full scale. Larger samples will normally have to be brought down by the use of voltage divider networks; smaller samples may be prone to inaccuracy due to hum and RFI on the samples, but if relatively clean, may be amplified by using a commercially available dc amplifier.

Metering inputs to a Remote Terminal may be either single ended (one side of input grounded) or floating (two sample lines, neither grounded, sometimes called differential). Some Remote Terminal analog telemetry inputs will tolerate a large amount of "float", (common-mode voltage) while others will only allow a few volts of potential above ground (or ground one side of the input). Samples with a large amount of common-mode voltage, such as cathode current samples, can be brought down near ground potential by the use of a dc isolation amplifier. In any event, it is not a good idea to apply samples floating more than about 50 volts above or below the ground to a Remote Terminal due to potential shock hazard while operating or servicing the equipment.

The transmitter control system manufacturer's input specifications should be studied carefully as to telemetry sample level, impedance, and configuration requirements. The Remote Terminal should be located directly adjacent to the transmitter(s) it is metering and controlling, and shielded cable may be required to eliminate RFI on the sample lines.

Most transmitter control systems allow telemetry inputs to be scaled for the proper readings given a representative sample of an actual parameter such as plate voltage. To illustrate, the actual plate voltage may be 7500 volts, the sample input to the Remote Terminal may be 2.5 volts, and the remote control system calibrated for a reading of 7500 by either the use of a potentiometer or a keyboard. Keyboard calibration simplifies metering calibration and verification at the transmitter site. In addition to linear calibration, many remote control systems offer power-to-linear conversion and product calibration (for indirect power). These types of calibration are accomplished in software and stored in the transmitter control system.

Some advanced transmitter control systems will accept direct digital telemetry inputs, usually in the form of binary or binary-coded decimal (BCD). This feature allows direct interface between, for example, a frequency meter and the transmitter control system. Once inputted, this telemetry may be processed like its analog counterparts as to tolerance limits, etc.

Status inputs to the Remote Terminal are normally required to be contact closures or voltages. Some remote control systems offer the ability to use either, and interfacing is possible on site between the two types of inputs using readily available components such as optical isolators.

Command outputs for transmitter control have traditionally been supplied as dry relay contacts, with the relays either contained in the Remote Terminal or on an accessory relay panel. Open collector transistor switch outputs are standard on some remote control systems, and the user may need to obtain either the manufacturer's accessory relay panel or build up relay interfacing on site for some control functions, especially for older transmitters. Some remote control systems offer latching command outputs, but external latching relays may be required so that the latches are held during all conditions. Optically isolated command outputs are also available for some remote control systems to simplify transmitter interface for dc control.

Control failsafe contacts are normally provided on Remote Terminals. This usually consists of a closed (during normal operation) set of relay contacts that will open after a short predetermined interval if control has been lost from the Control Terminal. A second set of relay contacts is added (or integrated into the control failsafe function) to allow a telemetry failsafe shutdown.

Maintenance override ("go home") contacts are also usually provided at the Remote Terminal in addition to a front-panel indication to notify the maintenance personnel that they should leave the remote control system in "remote" mode before they leave the transmitter site. Maintenance override is essential to provide Remote/Local capability to protect personnel from inherent equipment hazards.

Control Terminal/Remote Terminal Communication

Several choices are available to the broadcast to provide communication between the Control Terminal and the Remote Terminal. The oldest method employs a telephone line or lines to carry control signals from the Control Terminal to the Remote Terminal as well as telemetry and status information from the Remote Terminal to the Control Terminal. Most transmitter control systimes called a voice-grade circuit). Some transmitter control systems require additional telephone line conditioning, so the user should check the transmitter control system requirements before purchasing the system to be sure that the required type of line is available between the studio and transmitter sites. Maximum allowable telephone line attenuation is also normally specified by the transmitter control manufacturer and should be checked with the telephone company. Both "two-wire" and "four-wire" telephone data circuits are available; both can be full-duplex. Almost all lines are a combination of types, as amplification via hybrids and unidirectional amplifiers is normally required at the telephone company central office to offset line losses.

Aural studio-to-transmitter links (STLs) are popular for relaying signals from the Control Terminal to the Remote Terminal. In this system, an aural subcarrier is generated either in or adjacent to the Control Terminal on a frequency capable of being carried by the aural STL without interfering with the program signal. A 26 kHz aural subcarrier is often used on a monophonic aural STL system, while 110 kHz, 152 kHz, or 185 kHz is often used on a composite aural STL system. The user should be sure that the frequency chosen will be properly carried by the STL system. Most aural STL systems accept a subcarrier of approximately 1.5 volts p-p (for 10% injection). It is then transmitted through the system and delivered at the same level by the STL receiver. The subcarrier signals are normally separated from the monophonic or composite baseband signal so that the control subcarrier is not sent on to the broadcast transmitter for general broadcast.

Television broadcasters have two options for carrying control information over their video STL systems. The first option is to employ the video STL manufacturer's standard audio subcarrier system, using a separate 15 kHz audio channel for the control information. The second (and more difficult) option is to inject a 26 kHz aural subcarrier into an existing audio subcarrier channel on the video STL. Many video STL systems do not offer subcarrier injection and recovery as part of their audio subcarrier scheme, so some ingenuity may be required to effect the required interface.

Low power UHF radio links, designed for continuous operation on the Group P channels in the 450 MHz range, may also be used to carry control signals. These low power radio links are intended for point-to-point communications, and may be thought of as one-way phone lines under the control of the broadcaster. Fig. 8 shows a typical UHF radio link for this application. These radio links are sometimes called Telemetry Return Links (TRLs) or Transmitter-to-Studio Links (TSLs).

Most of the choices described above are also applicable to relaying telemetry and status communications from the transmitter site back to the control point. All of them are available if one includes the possibility of program material originating at the transmitter site being relayed to the studio, such as in the case of television electronic news gathering (ENG) signals being received at the transmitter site and relayed to the studio.

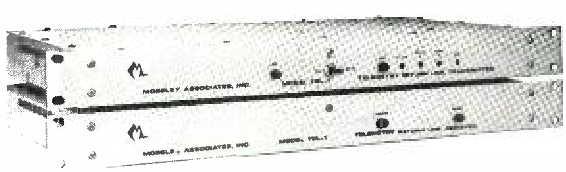


Fig. 8. Telemetry return link system.

In addition to those methods, the FM or television broadcaster may also insert a subcarrier on the broadcast transmitter aural carrier. FM broadcasters may use either 41 kHz, 67 kHz, or 92 kHz subcarriers for monophonic use; FM broadcasters transmitting in stereo may use either 67 kHz or 92 kHz. Television broadcasters normally use a 39 kHz subcarrier, although the transmission of stereo television requires that a secondary audio program channel (SAP) on 102.271 kHz be used for the telemetry signals.

Over-the-air subcarrier transmission of telemetry information is relatively inexpensive, but does have the drawback of not being available when the transmitter is off and does tie up subcarrier bandwidth that might be used for other services.

Some of the advanced transmitter control systems allow the use of two data transmission channels in either a simultaneous or alternate/main configuration to provide redundancy in case of failure of one data link. Dial-up telephone lines are not currently an alternative for broadcast transmitter remote control, as the control and telemetry capability must be continuous. Some systems do, however, allow the use of dial-up data links as an alternative during primary data link failure. (The dial-up link must be an "open" line to maintain continuous control and telemetry.)

A limited number of transmitter control systems allow the use of subaudible telemetry communications from the Remote Terminal to the Control Terminal. This subaudible information is either injected on an AM carrier or an FM subcarrier to avoid interference with the main signal. This is an inexpensive method of providing telemetry information. However, it does have the drawback of not being available when a transmitter is off the air, as well as employing a relatively low data transmission rate which translates into slow response time to changes at the transmitter. Figs. 9, 10, and 11 show some typical Control Terminal/Remote Terminal communication schemes.

Transmitter Facility Interfacing

As mentioned before, some interfacing is normally required between even the newest broadcast transmitters and the typical transmitter control systems. These interface requirements may include command outputs, telemetry inputs, and status inputs. Command Outputs provide for the actual functions to be accomplished at the

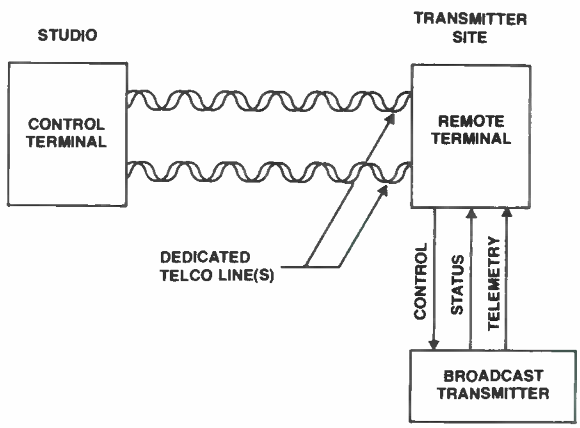


Fig. 9. Telco communications.

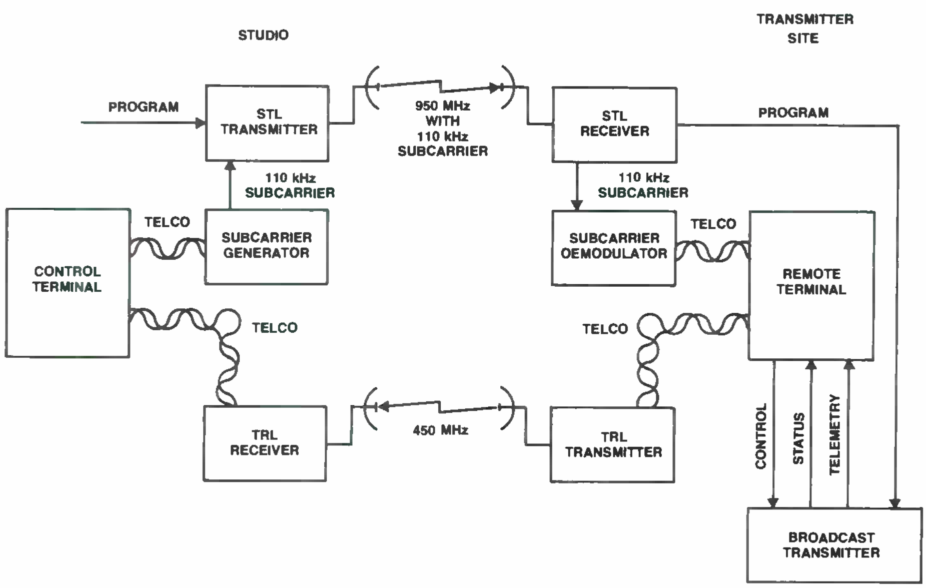


Fig. 10. Wireless communications.

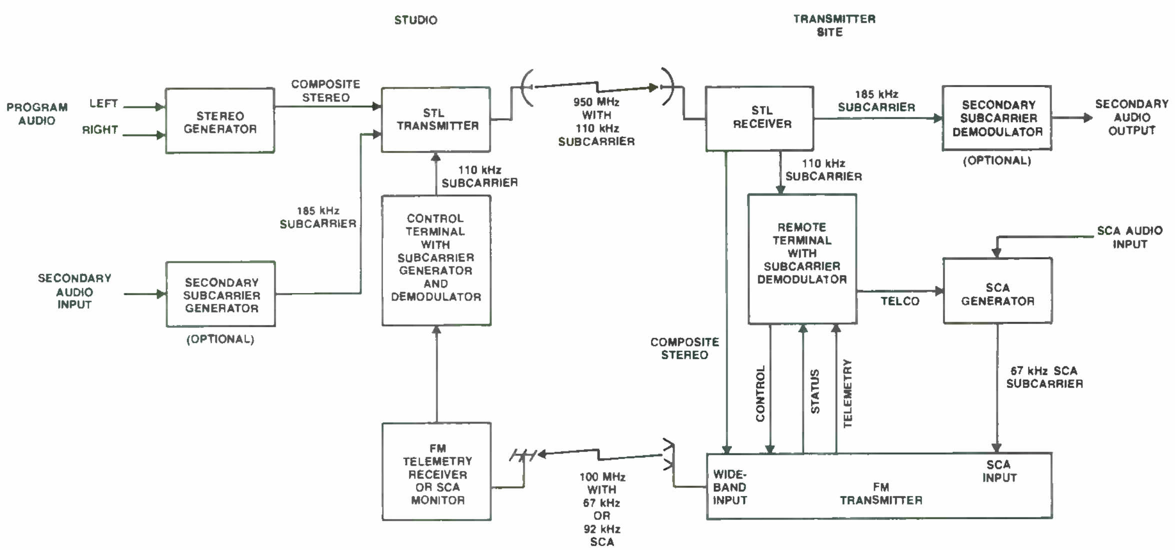


Fig. 11. Wireless communications via subcarrier.

transmitter plant such as filament on/off, plate voltage on/off, etc. Telemetry, or metering, is the sampling of the desired parameters, the result being a dc voltage of low value that is proportional to the sampled parameter. Status is the "go"/"no-go" input showing change of state.

Control/Command

Control from a transmitter control system is normally supplied as a momentary contact closure, either through a relay, through an optical isolator, or directly from a transistor switch. Individual transmitter control networks are normally the determining factors for the type of control interface, newer models requiring only a closure between a common terminal (usually ground) and various other terminals to remotely control the transmitter. In some cases, a common terminal and/or control terminations may be at such a voltage that relays are required to avoid shock hazard and/or excessive voltage within the transmitter control system. These relays also provide protection to the transmitter control system should an abnormal condition occur in the control network of the transmitter due to excessive voltage.

Some controls will be used to operate relays within the transmitter (filaments on/off, plate voltage on/off, etc.) while other closures in the transmitter system will be used to rotate devices such as potentiometers and variable capacitors. Most transmitter manufacturers provide, either as standard or as accessories, motor assemblies for controlling these variable devices. External reversible motors may also be used to translate transmitter control system commands to rotation.

Metering/Telemetry

Certain parameters must be monitored by the operator at the remote control point. These parameters are defined in the FCC Rules and Regulations relating to the particular type of service (AM, FM, or TV). These parameters typically include plate voltage, plate current, and output power. Most newer broadcast transmitters provide representative samples of these and other parameters for the remote control system. External interface units, described below, may be used for plate voltage, plate current and other parameters, such as tower light current, not provided as samples from the transmitter itself.

Plate Voltage

Plate voltage is normally sampled using a resistive voltage-divider network that reduces the plate voltage to a potential on the order of 1 to 5 volts dc. Fig. 12 shows a schematic of a commercially available plate voltage sampling unit.

This type of unit is available for use up to 20 kV,and can be extended to 30 kV by the use of an external resistor between the sample point and sampling unit input; this resistor would be in the same order of value and rating as those used within the plate voltage sampling unit.

Plate Current

Plate current can be sampled as a voltage across a low value resistor between the transmitter ground and the cathode. This current sample will, of course, include any grid and screen currents present. Universal plate circuit sampling kits are commercially available, providing the ability to monitor either plate current or plate voltage with

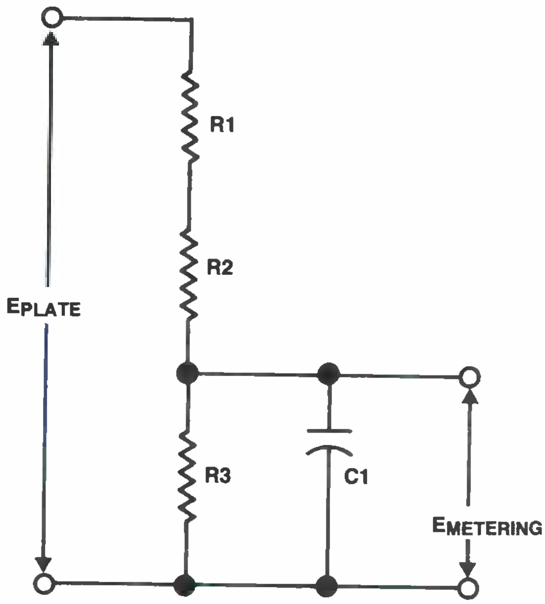


Fig. 12. Plate voltage sampling schematic.

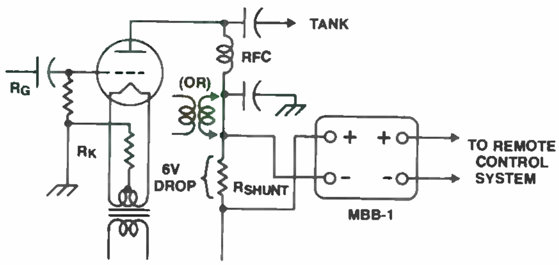


Fig. 13. Monitoring plate current.

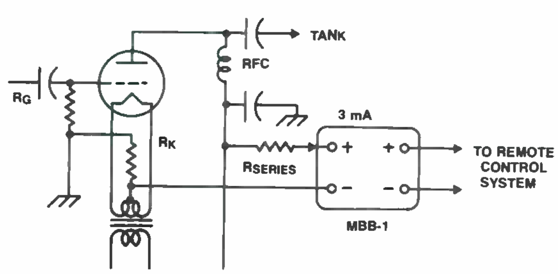


Fig. 14. Monitoring plate voltage.

up to a 10 kV potential. The voltage drop across these units is low, on the order of 4 to 6 volts. For current, external shunting is used across the input of these devices to keep the voltage drop quite low. Figs. 13 and 14 show typical installations to monitor plate current and plate voltage using this type of device.

Tower Lights

Tower light operation can be observed at the control point through the use of the transmitter control system. Both a flashing beacon as well as steady side lights may be monitored, with the preferred method being to monitor them separately. Tower light current sampling units are commercially available, consisting of a transformer winding and a rectifier circuit attached to a laminated core.

A current carrying conductor for the tower lighting being monitored is wrapped around the core up to several times to provide the proper sample output voltage.

AM Base Currents

Standard AM broadcast stations operating via remote control or ATS must provide a remote indication of antenna base current phase and amplitude. For those stations using directional arrays, all active towers plus the common point would have to be included. Commercially available AM tower monitoring units are available, and some include the ability to relay telemetry information to a remote control system via analog outputs. These telemetry outputs tend to be sequential on some units, however, so additional interfacing may be required in order to provide proper sample treatment for easy monitoring and/or automatic logging. The appropriate FCC Rules and Regulations should be consulted to determine the type of antenna sampling device to use for a given application.

RF Power Output

The output of broadcast transmitters may be sampled in a variety of ways. Most transmitters include a dc output for RF output sampling. External RF sampling devices are available from a number of manufacturers, and may be used to measure forward and reflected power. Inexpensive nondirectional output sampling units are also available. In some cases, it may be simpler to amplify the sample already present in the transmitter system if that will not adversely affect the reflectometer calibration. Power-to-linear conversion is normally done within the transmitter control system, so external devices for that application are no longer required. Care should be exercised so that the transmitter's forward and reflected power metering is not adversely modified by AM modulation.

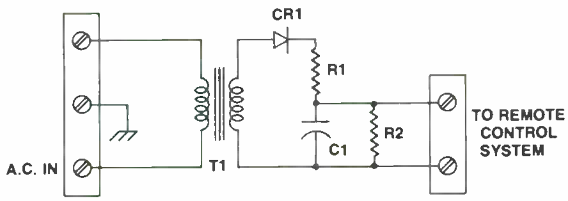


Fig. 15. Line voltage metering schematic.

Line Voltage

It is often desirable to monitor ac power mains voltage at the transmitter control point. Fig. 15 shows a typical line voltage sampling unit schematic, which may be used from 110 volts to 440 volts ac. Three-phase power may be monitored by using one of these units on each phase. Line current may be monitored by using a tower light sampling unit as described before.

Filament Voltage

A simple voltage divider and rectifier circuit may be used to adapt an incoming filament voltage for remote control monitoring. Fig. 16 shows a typical filament voltage sample schematic. Alternatively, an ac line voltage sampling unit may be used to sample the line voltage applied to the filament transformer.

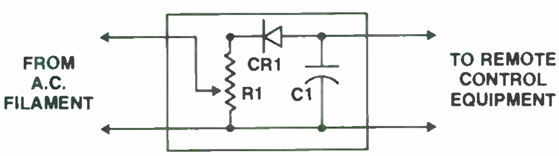


Fig. 16. Filament metering schematic.

Fig. 17. Temperature sensing kit.

Temperature

It is often desirable to be able to monitor critical temperature inside and outside the broadcast transmitter from the control point. This can be accomplished by the use of on/off thermostats, commonly available from heating supply stores, or by the use of temperature sensing units that convert temperatures above and below zero degrees to proportionate voltages positive and negative relative to zero volts. Some of these units allow the user to extend the temperature sensor away from the temperature sensing unit chassis. A typical air temperature sampling unit is shown in Fig. 17.

Status Inputs

Status inputs to a transmitter control system are typically either contact closures or dc voltages. In some cases, it is necessary to be able to convert the sample from one type to another for proper interface. Low voltage samples can be converted to contact closures (or their equivalent) by the use of readily available optical isolators with the appropriate resistance in series with the input. Higher voltages can often be converted to contact closures by the use of relays. In the case of conversion from contact closures to voltage input, the same type of interface units may be used, with external wetting voltage applied from a separate power supply. Various types of sensors are available commercially to provide on/off indications. These include smoke detectors, pressure switches, temperature switches, magnetic intrusion switches, photoelectric switches, sail switches for air movement, key switches, and motion detectors. In short, many types of parameters may be monitored for status input to the transmitter control system.

Uninterruptable Power Supply

Some installations have a requirement to know what is happening at the transmitter plant even during power failures. In this case, an Uninterruptable Power Supply (UPS) is probably required. The Remote Terminal and any communications equipment at the transmitter plant (STLs and TSLs) would be plugged into the UPS. The UPS would then supply ac power inverted from its own batteries to the transmitter control system when main ac power fails. UPSs are available from many manufacturers.

Television Monitoring Equipment

An off-air video demodulator is needed at the control point to permit continuous monitoring of the waveform and other characteristics of the transmitter visual signal. As an additional requirement, a separate video demodulator will be needed at the transmitter site for use in making measurements and setup adjustments.

A video waveform monitor is required for continuous monitoring of the waveform and other characteristics of the video signal including the percentage of modulation. A vectorscope or other instrument designed to depict the instantaneous phase and amplitude relationships of color components must be provided if any portions of the transmissions are in color. This apparatus must be capable of providing full field displays, and displays of test signals inserted on selected lines in the vertical blanking interval. The waveform monitor at the remote control point must be calibrated against the waveform monitor maintained at the transmitter during each required inspection.

Although not required by the Rules, color monitors should be provided if color program material is transmitted. It is suggested that both a monochrome and a color video monitor be provided.

Vertical interval test signal (VITS) generation is no longer required by the FCC Rules and Regulations. The VITS generated signals, do however, allow monitoring of selected parameters any time during the broadcast day as well as providing a basis for automatic correction during picture transmission.

Modulation Monitors

Audio modulation monitors are no longer required by the FCC to be type accepted and located in each broadcast facility. Each licensee is, however, responsible for maintaining modulation within the limitations imposed by the FCC. For this reason, some sort of modulation monitoring must be employed, either on a continuous or intermittent basis, to be sure that overmodulation does not take place. A number of manufacturers produce modulation monitors for AM, monophonic FM, stereophonic FM, monophonic television, and television subcarrier modulation measurement. Stereophonic television modulation monitors are required for stereo television transmission as well as the secondary audio program channels (subcarriers).

3.8

Frequency and Aural Modulation Monitoring

Joseph C. Wu
President
TFT, Inc.
Santa Clara, California

GENERAL

Broadcast station monitors fall into two categories—**frequency monitors** and **modulation monitors**. These two functions are sometimes combined into one unit and are sometimes packaged as separate units.

Broadcast aural modulation monitors are also classified according to the type of modulation to be monitored:

- AM Monaural
- AM stereo
- FM monaural (and SCA as needed)
- FM stereo (and SCA as needed)
- TV monaural
- TV stereo (and multichannel sound)

Two of these monitoring functions are sometimes combined in one instrument.

The various types of monitors are discussed in the following paragraphs, together with the FCC rules that the monitored transmitter's emissions must comply with.

WHY MONITOR YOUR TRANSMITTER

There are three major reasons for using frequency and aural modulation monitors at the broadcast station:

Coverage.

Since coverage of the broadcast service area is a function of transmitter carrier power and modulation level, it is desirable to maintain the modulation level at the maximum legal limit.

Equipment Performance.

By means of the monitors, the station engineer can perform equipment performance measurements to ensure that the transmitter is operating within the technical specifications of the FCC Rules.

Compliance With Legal Requirements.

The monitors enable station employees to operate the transmitter in accordance with FCC Rules regarding aural modulation level and carrier frequency.

FCC RULES AFFECTING MONITORING

The FCC does not specify the type of aural modulation monitor or measuring equipment a broadcast station must use. It is up to the station engineer to decide what monitoring equipment is needed to ensure that the station transmitter emissions comply with FCC frequency and modulation requirements.

Frequency Monitoring

Parts 73.1540 and 73.1545 of the FCC Rules require that the carrier of center frequencies of AM, FM, and TV stations must be measured or determined as often as necessary to ensure that they are maintained within the following tolerances:

AM Stations - The carrier frequency for monophonic transmissions or the center frequency for stereophonic transmissions must be within ±20 Hz of the assigned frequency.

FM Stations - The center frequency must be within ±2000 Hz of the assigned frequency (±3000 Hz for transmitters having a power output of 10 watts or less).

TV Stations - The visual carrier frequency must be within ±1000 Hz of the assigned frequency. The aural carrier frequency must be 4.5 MHz ±1000 Hz above the actual visual carrier frequency.

Aural Modulation Monitoring

Although type-approved monitors are no longer required, Part 73.1570 of the FCC Rules states that modulation percentage is to be maintained at the highest level consistent with good transmission quality and broadcast service, not to exceed the following limits:

AM Stations - Modulation of the carrier must not exceed 100% on negative peaks of frequent recurrence, or 125% on positive peaks at any time. There are additional limitations for AM stereo and telemetry transmissions.

FM Stations - Total modulation must not exceed 100% on peaks of frequent recurrence referenced to 75 kHz deviation. However, stereo stations simultaneously providing subsidiary communication services on subcarriers may increase the total peak modulation 0.5% for each 1.0% of subcarrier injection modulation; but the total carrier modulation must not exceed 110% (82.5 KHz peak deviation).

TV Stations - In general, the total modulation of the aural carrier must not exceed 25 kHz deviation (monaural) on peaks of frequent recurrence.

Stations transmitting multi-plexed subcarrier signals on the aural carrier must limit the modulation of the aural carrier by the arithmetic sum of the subcarrier(s) allowable deviation and the total modulation must not exceed 75 kHz deviation. Modulation requirements for stations transmitting aural subcarriers as part of encoded subscription programs are stated in the application for advance FCC approval.

ESSENTIAL FEATURES OF FREQUENCY MONITORS OR METERS

AM frequency monitors or meters must be capable of measuring and displaying carrier frequency error with a resolution sufficient to ensure compliance with the FCC carrier frequency requirements. Recommended resolution for AM monitoring is 1 Hz while FM and TV monitors should provide a resolution of 10 Hz or better. These tolerances call for an extremely accurate and stable internal frequency standard with an aging rate of 1 ppm per year or better which is traceable to the National Bureau of Standards (NBS).

If the transmitter is to be monitored at some distance from the transmitter site, a built-in or external preselector is required for receiving the signal off the air.

Fig. 1. A general purpose frequency counter may be used to measure carrier frequency as well as other frequencies within its range. Specified accuracy and resolution for the model shown here make it suitable for measuring carrier frequencies of AM, FM and VHF TV. (Courtesy of Leader Instruments Corp.)

The following features are also highly desirable in the frequency monitor:

- A means of presetting the frequency of broadcast transmitters off the air.
- An output (BCD or analog) for automatic logging of frequency error.
- An output to operate an alarm when present frequency limits are exceeded.
- Provision for calibrating the internal frequency standard against an NBS station or other highly accurate standard.

Fig. 2. This combined frequency and modulation monitor includes capabilities for monitoring various aspects of the FM stereo signal. (Courtesy of Belar Electronics Labs, Inc.)

ESSENTIAL FEATURES OF AN AURAL MODULATION MONITOR

General

As a minimum, all aural modulation monitors must have a peak-reading modulation meter to give a direct indication of modulation percentage. Meter accuracy should be ±4% or better. Other desirable features common to all modulation monitor are:

- Flashers, accurate to ±2% or better, to indicate when maximum positive and negative modulation peaks are occurring.
- Adjustable flashers to indicate when modulation peaks exceed preset levels.
- An internal calibrator to check the accuracy of the modulation meter and peak flashers, and a means for recalibrating the meter and flasher circuits.
- An output to operate an overmodulation alarm.
- An output to operate an external alarm when the modulation drops below a certain level (e.g. 10%) for a specified period of time.
- Outputs to operate a remote meter and remote peak flashers.

AM Monitors

AM monaural modulation monitors should possess the features described in the preceding paragraph. If a preselector is used with the monitor, its sensitivity should be approximately 100 μV for a 35 dB SNR, and 1 mV for a 50 dB SNR. Selectivity should be at least −40 dB at ±40 kHz, and image rejection should be at least 50 dB.

AM stereo monitors should have the following additional features:

- L + R and L − R channel decoding.
- L and R channel separation measurement capability.
- L and R channel signal-to-noise ratio measurement capability.
- L and R channel frequency response measurement capability.
- Pilot carrier injection level measurement capability.
- Channel crosstalk measurement capability.
- Signal output for distortion measurements.

FM Monitors

FM monitors must have a high signal-to-noise ratio, at least 70 dB. The discriminator must have a distortion figure of 0.1% or better and a frequency response of at least 25 Hz to 100 kHz.

If a preselector is used for off-the-air monitoring, its IF amplifier must have a bandwidth of at least ±275 kHz, and yet be narrow enough to reject alternate channels. A built-in multipath detector is highly desirable to minimize multipath interference. The FM monitor should be able to measure AM noise on the carrier as well as the FM signal.

FM stereo monitors must be able to measure the (L + R) channel level (30 Hz to 15 kHz), the (L − R) channel level (23 to 53 kHz), the 19 kHz injection level, and the 38 kHz subcarrier level. It must also be able to measure crosstalk (separation of left and right channels) up to 60 dB.

If an SCA signal is being transmitted, an SCA monitor must be used in addition to the FM modulation monitor. The SCA monitor must be able to measure modulation percentage, the SCA injection level on the composite signal, signal-to-noise ratio, and crosstalk. The SCA modulation measurement circuit must be adjustable for whatever deviation is defined as 100% modulation.

TV Monitors

Desirable features for TV monaural and stereo monitors are the same as for FM monitors. When

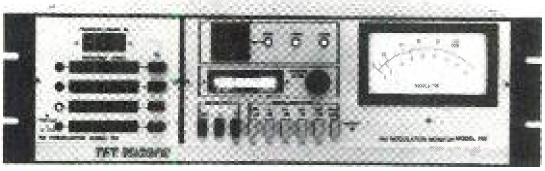


Fig. 3. For remote monitoring, a preselector is used to provide sufficient gain so that the signal will drive the modulation monitor. A multichannel preselector allows monitoring modulation characteristics of other stations in the market. (Courtesy of TFT, Inc.)

multichannel sound (MTS) is being transmitted, the operator should be able to monitor the main channel independently of the other channels, and vice-versa. The MTS monitor must receive and demodulate the composite baseband, and measure composite deviation as well as injection levels.

ON-SITE MONITORING TECHNIQUES

When monitoring is to be done at the transmitter site, the monitor is normally connected directly, or through an attenuator, to the RF sampling point of the transmission line feeding the antenna.

AM Monitoring

Most AM frequency monitors display carrier frequency error rather than frequency. Typically, a front-panel digital display indicates the carrier deviation from its assigned frequency in Hz, and a lighted "+" or "−" indicates whether the carrier is above or below its assigned frequency.

The percentage modulation of a monaural AM carrier is read directly on the monitor's front-panel meter or other indicating device. Monitors usually permit selection of either positive or negative modulation peaks by means of a front-panel switch. Peak flashers, used on some monitors, are intended to catch fast transients and peaks that the meter cannot respond to. There may be one flasher to indicate maximum allowable negative peaks (100%) and another to indicate maximum allowable positive peaks (125%). The monitor may also have an adjustable peak flasher. This can be present by means of digital switches in that it flashes when the modulation percentage exceeds the preset value.

For modulation monitoring of an AM stereo transmitter, an AM stereo monitor or a stereo monitor plus a compatible AM modulation monitor is required. The system used should enable the operator to read simultaneously the percentage modulation on both left and right channels, and to measure separation between channels and crosstalk between the main channel and subchannels.

FM Monitoring

A typical FM frequency monitor digitally displays carier frequency error in Hz, and indicates whether the carrier is above or below its assigned frequency. The RF input level must be adjusted so that sufficient signal is available for measurement but not so high that overloading occurs. When an FM stereo transmitter carrier frequency is being monitored, the stereo pilot carrier frequency can also be measured.

Monaural modulation percentage is normally displayed on a front-panel meter. Typically, either positive, negative, or combined modulation peaks can be selected for monitoring by a switch. Some monitors also provide peak flashers to indicate modulation peaks that exceed a preset percentage. The RF input must be adjusted to the correct level, as described in the instruction manual, before accurate readings can be taken.

For monitoring a stereo FM transmitter, an FM stereo modulation monitor or an FM modulation monitor with a compatible stereo monitor must be used. Left-channel, right-channel, and total modulation are read on front-panel meters. With the typical monitor, these meters can also be used to measure separation between the left and right channels as well as crosstalk between the main channel and subchannels, as described in the monitor's instruction manual.

TV Monitoring

A typical TV frequency monitor provides separate displays of the visual carrier, aural carrier, and intercarrier frequency errors. The RF input to the monitor should be adjusted to the correct level as described in the instruction manual before measurements are made.

Aural modulation percentage can be read on the monitor's front-panel meter. If the TV transmitter is also transmitting stereo and/or MTS, a monitor with a stereo decoder for left and right channels and a decoder for the SAP channel must be used.

OFF-SITE MONITORING TECHNIQUES

For remote, off-the-air transmitter monitoring, the monitor must contain a preselector (RF amplifier) or an external preselector must be used.

Input to the preselector is from a suitable antenna. Some preselectors are capable of measuring and indicating carrier frequency error. Such preselectors combined with a modulation monitor provide a complete transmitter monitoring capability.

Off-site monitoring techniques are generally the same as on-site techniques. RF input level to the preselector or monitor must be carefully adjusted, as described in the instruction manual, for accurate frequency and modulation measurements.

CALIBRATION AND MAINTENANCE OF MONITORS

Frequency Monitor Calibration

Frequency monitors should be calibrated periodically to ensure accurate measurement of transmitter carrier frequency error. The time between calibrations depends on the frequency of the monitor's internal standard, which in turn depends on the operating frequency of the transmitter. For a monitor having an internal crystal standard that has been correctly calibrated once and that has an aging rate of 1 ppm per year, the following calibration schedule is recommended:

Transmitter Frequency	Calibration Interval
AM band	Every 12 months
FM band	Every 12 months
Low VHF TV band (Chs 2 - 6)	Every 12 months
High VHF TV band (Chs 7 - 13)	Every 6 months
UHF TV band	Every 3 months

After the first few calibrations, the interval may be lengthened if the drift observed in the first few checks warrants it.

There are a number of ways the internal frequency standard can be calibrated:

a. A secondary standard such as an HP5245 Counter or an HP105A Quartz Oscillator can be used with an oscilloscope. The output of the secondary standard is used to synchronize the oscilloscope and the internal standard's output is applied to the vertical input of the oscilloscope. The frequency of the internal standard is then adjusted for the least movement of the oscilloscope display.

b. If the monitor's frequency display can be operated in a general-purpose counter mode, a frequency standard of higher accuracy than the monitor's internal standard (e.g., 3.58 MHz chromatic subcarrier transmitted on TV network originated programs) can be used with the counter to calibrate the internal standard. The external standard is connected to the counter input, and the internal standard (which furnishes the time base for the counter) is adjusted so that the counter displays the exact frequency of the external standard.

c. If a receiver capable of receiving one of the NBS stations (WWV, WWVB, etc.) is available, and if the monitor has an output from its internal standard that is nominally an exact subharmonic of the frequency of the NBS station being received, the monitor's internal standard can be calibrated by adjusting its frequency for zero beat with the NBS station frequency.

Aural Modulation Monitor Calibration

The monitor's aural modulation meter and peak flashers should be calibrated regularly. Most monitors have built-in calibrators, so that meter and flasher accuracy can be checked by simply pressing front-panel switches and observing the peak flashers and the meter reading. If the reading is in error, a simple adjustment corrects the error.

If an FM modulation monitor has no internal calibrator, the monitor must be calibrated against a laboratory standard or by means of a Bessel-null measurement using a spectrum analyzer and a precision audio frequency generator. If an AM modulation monitor has no internal calibrator, the monitor can be calibrated using an RF generator. The generator must be capable of very low distortion amplitude modulation, and the level of this modulation should be accurately observed by using a high quality oscilloscope that has high linearity.

Maintenance

A broadcast station monitor should be maintained in the same way as other precision laboratory instruments. It should be calibrated regularly as described in the preceding paragraphs.