Modulation, Overmodulation, and Occupied Bandwidth:

Recommendations for the AM Broadcast Industry

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HARRISON J. KLEIN, P.E.[1]

I. Executive Summary

This technical report was prepared on behalf of the AM Improvement Committee of the National Association of Broadcasters. Its purpose is to provide recommendations for AM engineers on how to assure (1) the transmission of a clean and full-fidelity AM signal, (2) the prevention of AM overmodulation, and (3) the prevention of "splatter" interference.

A computer analysis of AM modulation and overmodulation was performed to assess the extent that out-of-band emissions result from overmodulation, improperly processed audio, and RF networks in transmitters and antennas. In addition, tests were performed to obtain data at an operating AM station. The effects of antenna bandwidth are discussed, and suggestions are given on the optimal locations for monitoring AM modulation and the best ways to measure occupied bandwidth.

The following six conclusions and recommendations are provided:

1. The primary cause of splatter interference is not the disappearance of the carrier during overmodulation, but instead is the presence of excessive high-frequency content in the audio signal that modulates the transmitter.
2. Meeting FCC bandwidth limits is no guarantee of a "clean" transmitted signal.
3. Splatter interference is minimized by (1) low-pass filters on audio prior to modulation, (2) final protective clippers in processors or at transmitter inputs, and (3) elimination of DC level shift in AM transmitters.
4. Modulation percentage as observed in the field is often inaccurate, differing from the modulation percentage observed at the transmitter, because of the effect of RF networks in the transmitter and the antenna system; to prevent overmodulation a station must have an accurate modulation monitoring point.
5. AM stations should evaluate modulation performance using appropriate analysis techniques, and tailor transmitter modulation adjustments accordingly.
6. The AM industry should develop a high-quality synchronous detector AM demodulator for accurate analysis of modulation characteristics in the field.

Each of these recommendations is discussed herein. Industry-wide attention to these recommendations would raise the quality of AM listening and reduce objectionable interference.

II. Introduction

Interference between stations is a serious problem for AM radio. Interference has been exacerbated by the increasing number of stations and the rising ambient radio-frequency noise level, but a major factor has been the increase in the occupied bandwidth of typical AM stations. Greater occupied bandwidth results in greater adjacent-channel interference, especially as perceived on wider bandwidth receivers. Excessive occupied bandwidth does *not* improve AM reception quality.

A station's occupied bandwidth depends upon its programming, its audio processing, its modulation level, and its transmitter and antenna bandwidths. Occupied bandwidth must be optimized to most efficiently transmit the desired signal to the receiver while minimizing undesired signal components that can cause interference. To optimize its bandwidth, a station must be able to control and monitor its modulation effectively, and must understand how modulation practices affect occupied bandwidth.

For example, overmodulation is one of the basic mechanisms by which an amplitude modulated carrier can be degraded. On the surface, overmodulation appears a straightforward phenomenon which should be easily understood and prevented. Yet many stations regularly overmodulate, some intentionally, implying that the overmodulation mechanisms are not fully understood or that the stations do not believe there are benefits to preventing overmodulation. On the other hand, receiver manufacturers consider overmodulation to be a significant AM problem. In numerous industry meetings, they have indicated their belief that *splatter caused by overmodulation is a major factor preventing the manufacture of improved, high-fidelity AM radios.*

A review of AM modulation, overmodulation, and occupied bandwidth is timely. This report provides an overview of the subject. Modulation and overmodulation are defined and their effects on occupied bandwidth are described. The spectral energies created by "clean" and by overmodulated AM signals are compared. The distribution

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of this energy is discussed. The effects on occupied bandwidth due to audio processors, transmitters, and antenna systems are reviewed, with emphasis on the roles of audio and carrier clipping.

Most importantly, techniques are discussed that engineers can use to prevent excessive occupied bandwidth. Processing techniques for minimizing overmodulation and splatter are described. Guidelines for modulation monitoring are provided. The benefits of clean signals to both broadcasters and receiver manufacturers are demonstrated.

III. Definitions of Modulation and Overmodulation

An amplitude-modulated carrier with a single modulating frequency can be described by the equation

$$e(t) = A(1 + m \cos 2\pi f_m t) \cos 2\pi f_c t, \qquad (1)$$

where A = amplitude of unmodulated carrier,
m = AM modulation index,
f_m = modulating frequency in Hertz,
and f_c = carrier frequency in Hertz.

The AM modulation index is merely the relative level of modulation; the term has a different and simpler meaning for amplitude modulation than it does for frequency modulation. Figure 1 shows four different waveforms corresponding to the above equation with varying values of m: 0.25, 0.5, 1.0, and 1.25.

Although modulation index is related to percentage of modulation (as discussed below), Section 73.14 of the FCC Rules officially defines modulation percentage in terms of the modulated RF envelope:

Positive percentage modulation = 100 (MAX - C) ÷ C,
Negative percentage modulation = 100 (C - MIN) ÷ C,

where MAX = instantaneous maximum envelope level,
MIN = instantaneous minimum envelope level,
and C = unmodulated carrier level (identical to A in equation for $e(t)$ above).

It can be seen from Figure 1 that, up to 100%, modulation index is equivalent to modulation percentage.[2] However, the two terms are not identical. There can be no negative modulation percentage over 100%, according to the above definition, because the envelope can never be smaller than 0. Yet there is no upper limit on the modulation index. As shown in Figure 1d, the carrier does not disappear when $m > 1$ in the equation, but merely "folds over" during negative modulation peaks.

There are two types of overmodulation discussed in this report. AM broadcast engineers are familiar with the type of overmodulation found in a typical AM transmitter and shown in Figure 2: as the audio input level to the transmitter is increased above that necessary to achieve 100% negative modulation, the carrier clips, or disappears, during negative peaks. It is essential to recognize the fundamental difference between this conventional overmodulation, and the overmodulation shown in Figure 1d in which $m > 1$ but no carrier clipping occurs. The second type of overmodulation is produced when an integrated-circuit multiplier is used to generate amplitude modulation. This is an alternative modulation technique used by some commercially available low-power AM transmitters. Amplitude and phase errors in RF networks, examined in Section VIII of this report, can also produce this effect. For clarity, this report will sometimes differentiate between "transmitter-type overmodulation" (Figure 2) and "multiplier-type overmodulation" (Figure 1d). Where "overmodulation" is used without qualification, transmitter-type overmodulation is implied.

Related to overmodulation is the term "splatter." Splatter is loosely defined to be any radio-frequency spectral components that increase the normally occupied bandwidth of the AM signal. Some splatter may be unavoidable or even intentional, such as the increased sideband energy produced by audio preemphasis. Other splatter is unintentional and can be prevented.

IV. The Amplitude Modulation Spectrum

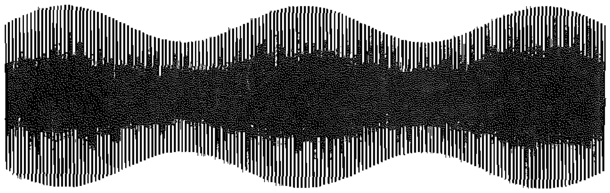
Using trigonometric identities, the amplitude modulation equation (1) can be written to distinguish the frequency components of the signal. For single-tone modulation (using angular frequency ω instead of f in Hertz, where $\omega = 2\pi f$, to simplify the notation):

$$\begin{aligned} e(t) &= A(1 + m \cos \omega_m t) \cos \omega_c t \\ &= A \cos \omega_c t + A\, m \cos \omega_c t \cos \omega_m t \\ &= A \cos \omega_c t \\ &\quad + 1/2\, A\, m\, [\cos(\omega_c - \omega_m)t + \cos(\omega_c + \omega_m)t]. \end{aligned}$$

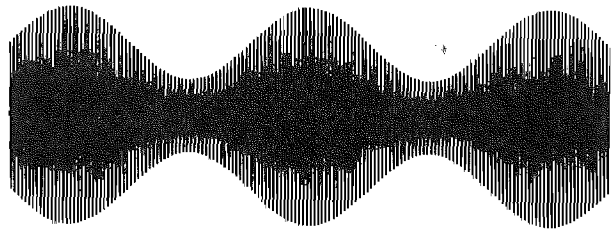
This equation shows the well-known three-component spectrum of an AM wave. For $m = 1$ (100% modulation) the spectrum contains the carrier component at frequency ω_c and two sidebands, each 6 dB below the carrier, separated from the carrier by the modulating frequency ω_m.

From this equation, it can be seen that an AM spectrum has several notable characteristics. The carrier component does not vary with modulation and the two sidebands are symmetrical about the carrier. Most importantly, the spectrum of the sidebands is the frequency-shifted spectrum of the modulating audio (inverted for the lower sideband). Audio-frequency components present at the input of the transmitter will appear on both sides of the carrier at their same relative levels. This means that under normal modulating conditions (i.e., no overmodulation or other non-linear effects) the sideband energy is determined directly by the modulating audio spectrum. An audio component of 20 kHz will appear 20 kHz on each side of the carrier. If its audio level is 10 dB below 100% modulation, the sideband levels will be 10 dB below 100% modulation (16 dB below carrier). In a modern transmitter with good high frequency response, any high-frequency distortion

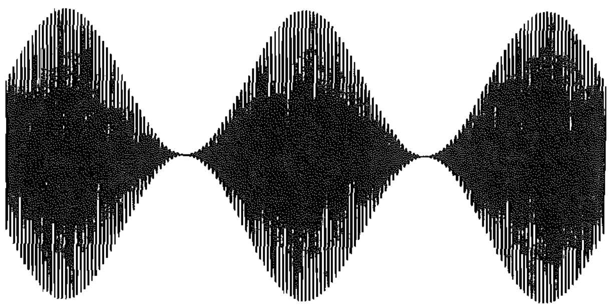
[2] This statement is valid for any symmetrical modulating audio waveform.



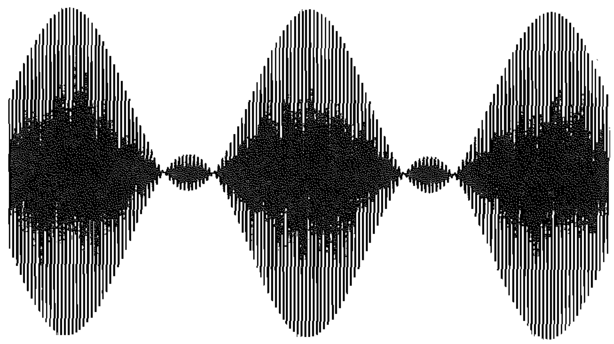
a. m=0.25



b. m=0.5



c. m=1.0



d. m=1.25

Figure 1. Waveforms of AM Carriers for Various Modulation Indices.

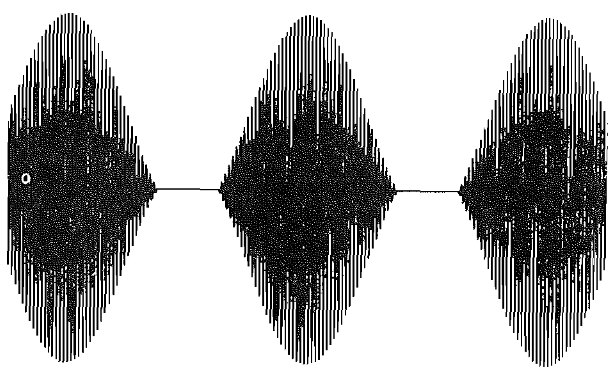


Figure 2. Overmodulation in a Typical AM Transmitter.

products or other processing artifacts reaching the transmitter input will be converted to splatter energy at the transmitter output.

V. AM DEMODULATION

The two conventional methods used to demodulate AM signals are envelope and synchronous detection. Envelope detection relies on the property that the envelope of an AM signal is identical, under certain conditions, to the waveform of the modulating audio.[3] An envelope detector merely rectifies and filters the AM signal to re-create the original modulating waveform, a technique that works properly only if the envelope remains undistorted during transmission. Unfortunately, imperfections between the transmitter and the receiver, some of which are discussed in Section VIII of this report, can cause envelope distortion. The resulting behavior of the AM signal during detection is important to the discussion of overmodulation.

Synchronous detection is not yet common in commercial AM receivers, but has several desirable properties that should make it much more common in the future. A synchronous detector (sometimes known as a product detector) merely repeats the multiplication process that was used to modulate the carrier. The AM signal is multiplied by a sine wave of the same frequency and phase as the transmitted carrier. It can be shown mathematically that this process, together with some simple filtering, results in the re-creation of the original modulating audio signal.[4]

A synchronous detector is not sensitive to envelope distortion. Amplitude and phase errors in the transmission path affect the received frequency response but do not cause an increase in distortion. This property, and its value in monitoring modulation, is discussed later in more detail.

It should be noted that synchronous detection is not without its practical limitations. In order to multiply the AM signal by a sine wave of the same frequency and phase as the transmitted carrier, the carrier must somehow be recovered in the receiver. If, due to such phenomena as skywave fading or co-channel interference, the carrier cannot be properly reconstructed, the quality of reception will be degraded.

VI. THEORETICAL AMPLITUDE MODULATION ANALYSIS

A. Fast Fourier Transform Techniques

Modulation analysis has been greatly enhanced by computer technology. Fast Fourier Transform (FFT) techniques now make it possible to quickly and easily perform conversions between the time domain and the frequency domain, without the necessity for extensive measurements.[5] For example, a digital representation of a modulated carrier can be generated, based on a desired modulation condition. This waveform can then be converted with a forward FFT program to the frequency domain, where the spectrum can be analyzed. The spectrum can be manipulated to simulate the effect of an electrical network such as a directional antenna, and then can be reconverted with an inverse FFT program to the time domain where the envelope can be inspected for distortion. This distortion can be quantified by digitally simulating an envelope detector to produce the time domain representation of the detected audio waveform, transforming again to the frequency domain, and calculating the percentage of distortion from the levels of harmonic components.

This technique has been used herein to theoretically analyze the modulation effects of several components in the broadcast system: clippers, transmitters, directional antennas, and receivers. These effects will be discussed throughout this report.[6]

B. FFT Modulation Analysis

Figure 3 shows the results of FFT analysis of single-tone modulation under various conditions of level and clipping. The following conditions were analyzed:[7]

a. 50% modulation
b. 100% modulation
c. 100% + 3 dB (141%), no clipping or carrier pinchoff (multiplier-type overmodulation)
d. 100% + 1 dB (112%), carrier pinchoff only
e. 100% + 2 dB (126%), carrier pinchoff only

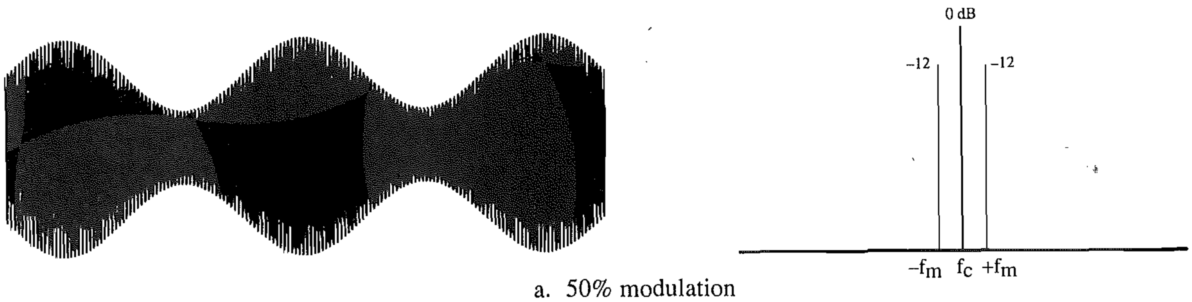
[3] Figure 1 demonstrates that, up to 100% modulation, the envelope of the signal is the modulating sine wave.

[4] See Appendix A for a mathematical analysis of synchronous detection.

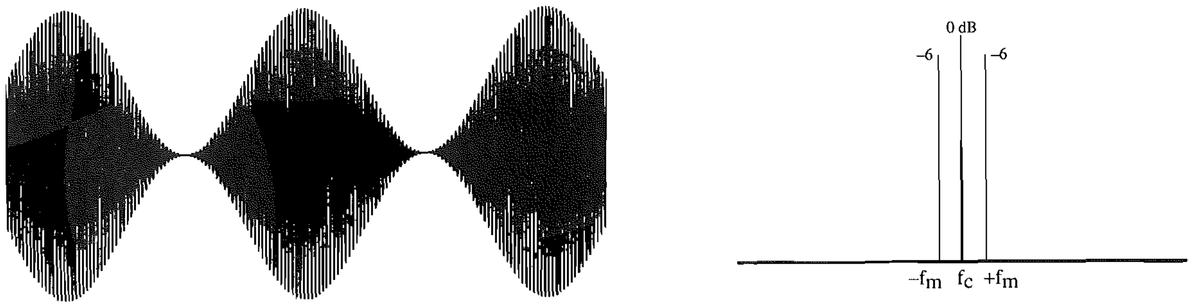
[5] The Fourier Transform is a mathematical technique to convert between the time-domain representation, or waveform, of a signal, and its frequency-domain representation, or spectrum. Before the advent of computers, the use of Fourier Transforms was time consuming and limited to certain mathematical functions such as sine waves. A computer algorithm known as the Fast Fourier Transform speeds the computation of the Fourier Transform and operates on essentially any waveform or spectrum. The forward FFT converts a waveform to its spectrum, while the inverse FFT converts a spectrum to its associated waveform.

[6] See Appendix B for a more complete description of FFT parameters.

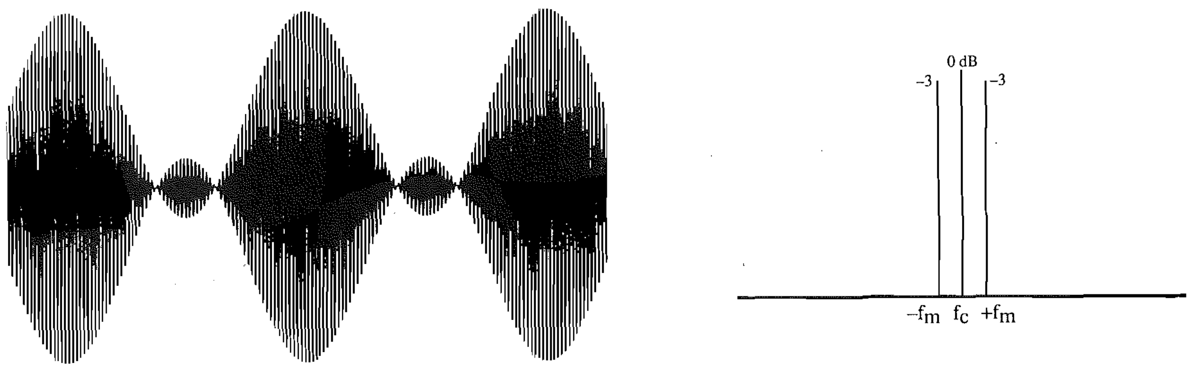
[7] Most AM transmitters, whose audio stages are AC-coupled, experience a DC level shift under conditions of asymmetrical clipping that affects the peak modulation level. For example, increasing the audio level by 3 dB above 100% modulation would theoretically result in a positive peak modulation of 141%. If the negative peak were clipped to avoid negative overmodulation, an AC-coupled transmitter would shift the DC level of the audio, reducing the positive peak modulation to just over 100%. The waveforms analyzed here were analytically generated and had no DC level shift under any modulation conditions. Therefore, the FFT spectra shown here will differ slightly from those of most physical AM transmitters.



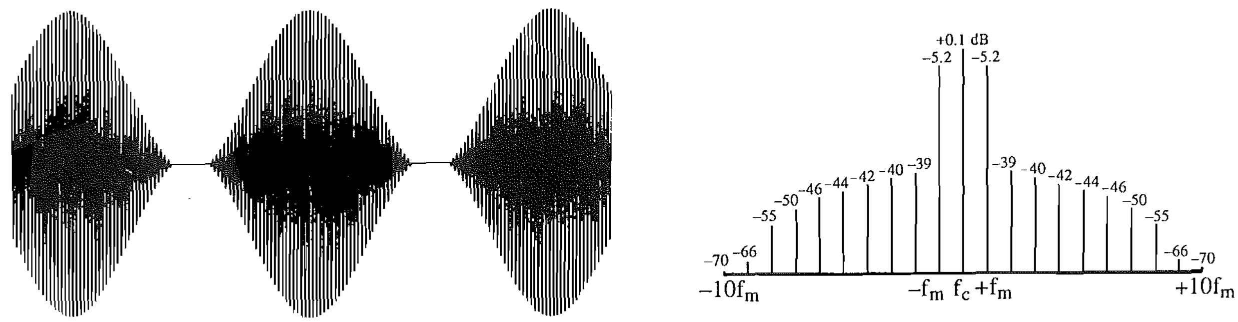
a. 50% modulation



b. 100% modulation

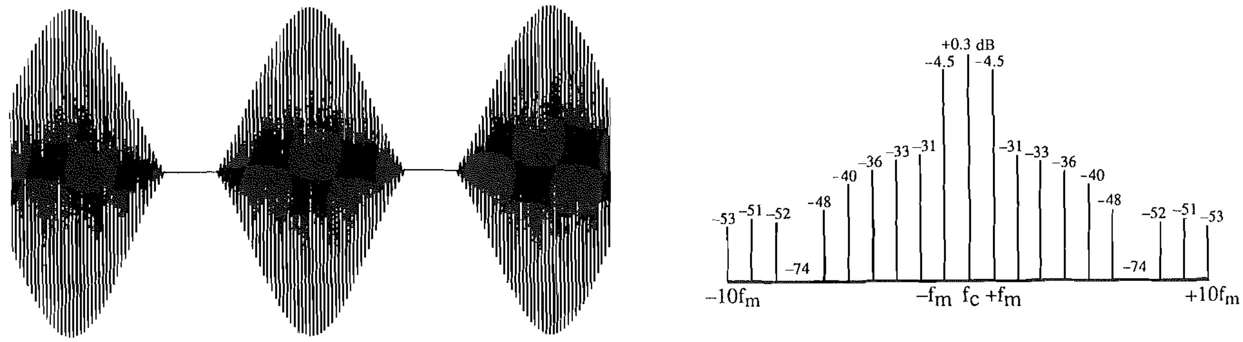


c. 100% + 3 dB (141%), multiplier-type overmodulation

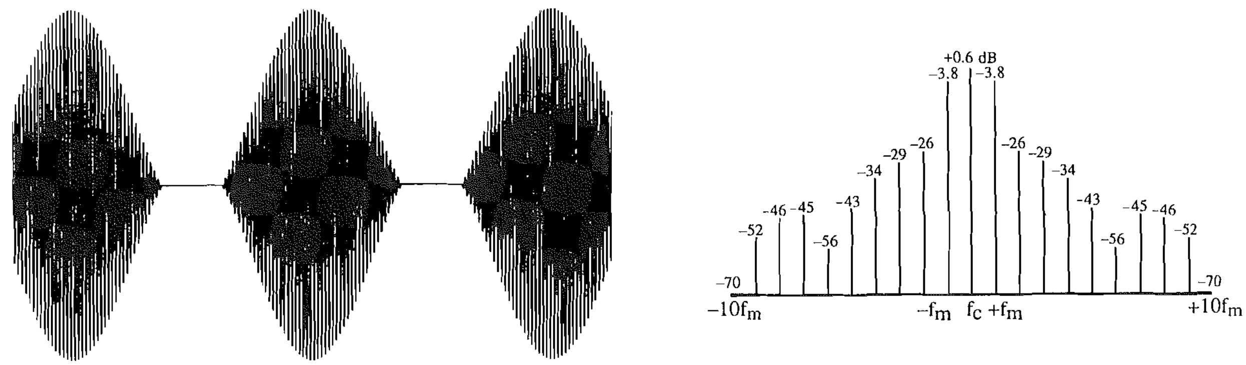


d. 100% + 1 dB (112%), carrier pinchoff only

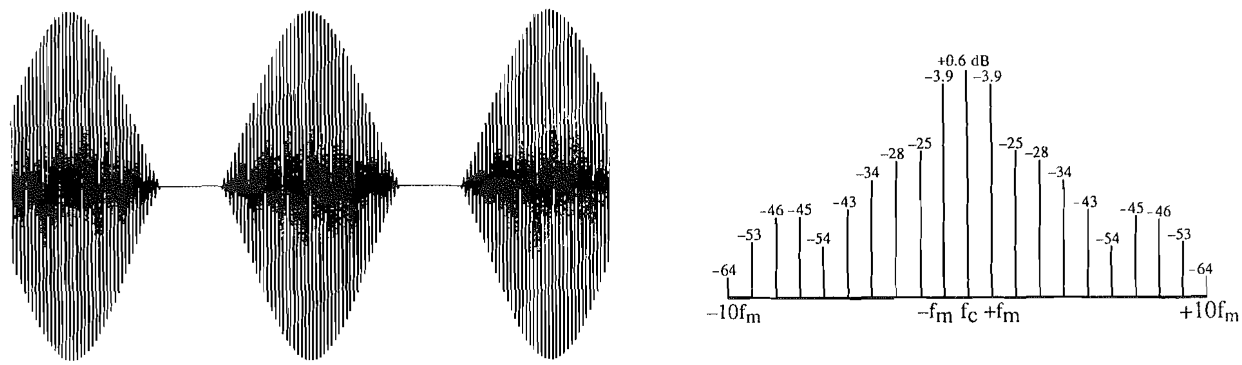
Figure 3. FFT Modulation Analysis - Waveforms and Spectra.



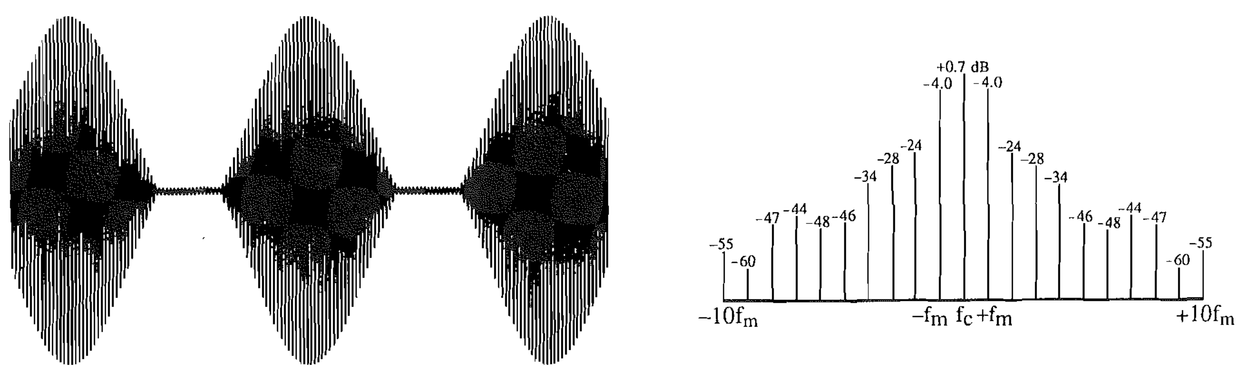
e. 100% + 2 dB (126%), carrier pinchoff only



f. 100% + 3 dB (141%), carrier pinchoff only

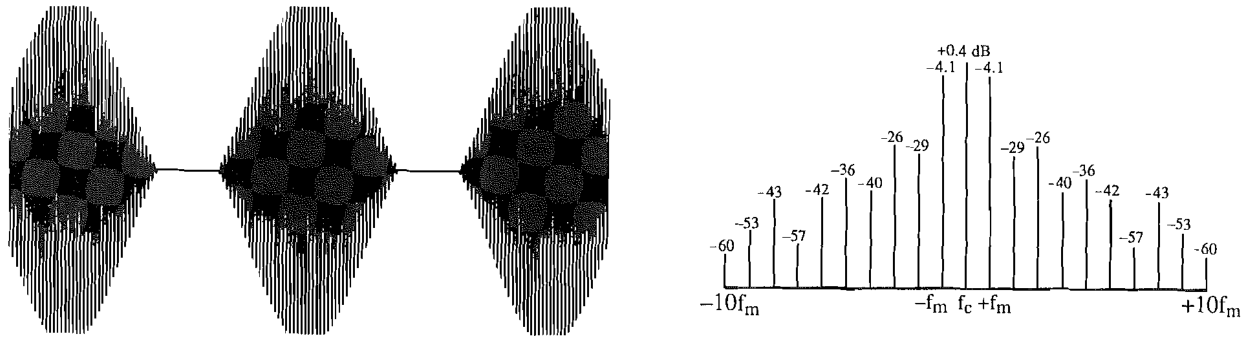


g. 100% + 3 dB, –99% audio clip

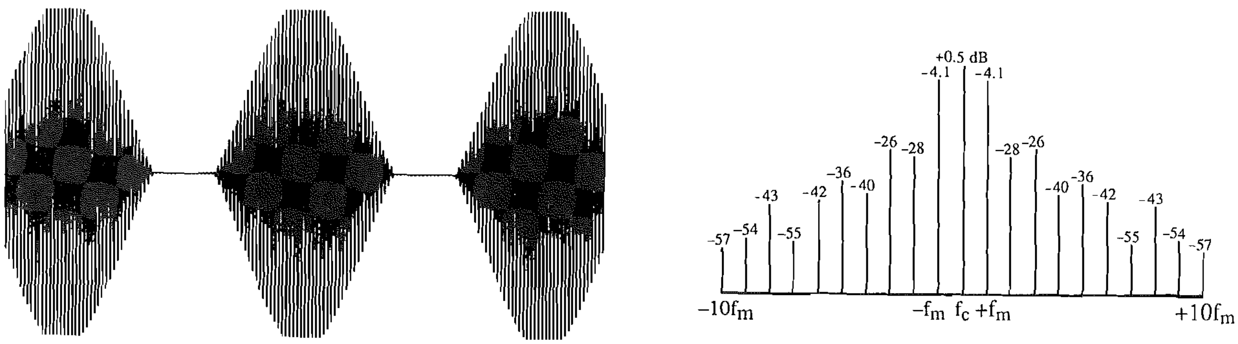


h. 100% + 3 dB, –95% audio clip

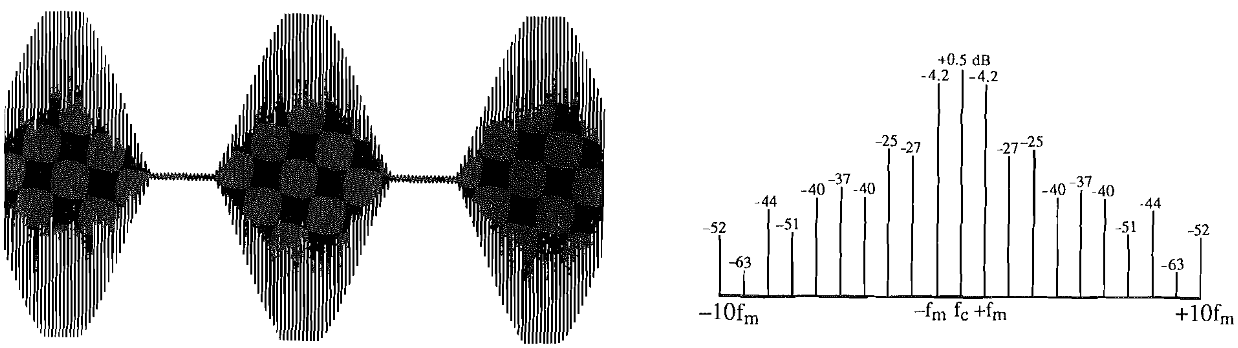
Figure 3 (cont.). FFT Modulation Analysis - Waveforms and Spectra.



j. 100% + 3 dB, carrier pinchoff, +125% audio clip



k. 100% + 3 dB, –99% audio clip, +125% audio clip



l. 100% + 3 dB, –95% audio clip, +125% audio clip

Figure 3 (cont.). FFT Modulation Analysis - Waveforms and Spectra.

f. 100% + 3 dB (141%), carrier pinchoff only
g. 100% + 3 dB, -99% audio clip
h. 100% + 3 dB, -95% audio clip
j. 100% + 3 dB, carrier pinchoff, +125% audio clip
k. 100% + 3 dB, -99% audio clip, +125% audio clip
l. 100% + 3 dB, -95% audio clip, +125% audio clip.

These figures demonstrate some important basic properties of modulation and splatter. First, the intermittent disappearance of the carrier, which is believed by many broadcast engineers to be the cause of splatter, actually has very little to do with it. Multiplier-type overmodulation, shown in Figure 3c, results in the carrier often going to zero but does not result in the generation of any new frequency components; conversely, Figures 3g, 3h, 3k, and 3l show that splatter can occur even though the carrier does not disappear. Second, it is not "sharp edges" of the modulation, per se, that produce splatter; again, the zero crossings of multiplier-type overmodulation are sharp but cause no splatter.

Most important, it is apparent from Figure 3 that negative peak clipping is the major cause of splatter and that, theoretically, it matters little whether the clipping occurs as carrier pinchoff in the transmitter or as clipping in the audio prior to modulation. Comparing the spectra of 3 dB overmodulation in Figures 3f through 3l, there is almost no difference between carrier pinchoff and audio clipping. The conditions of carrier pinchoff, –99% audio clipping, and –95% audio clipping, produce spectral components which are stronger than –35 dB out to the fourth harmonic of the modulating frequency, and produce additional significant components even farther out.

When +125% audio clipping is added to an overmodulated signal, as in Figure 3j, or to a negatively clipped signal, as in Figures 3k and 3l, the effect is less significant. While some spectral components are increased if positive clipping is added, others are actually reduced due to the phase relationships between the components produced by positive and negative clipping. Also, the larger components produced when negative peaks are clipped from 141% down to 100% masks the smaller components produced when positive peaks are clipped from 141% down to only 125%. The net additional splatter energy produced by +125% clipping is small.

It can be concluded from this data that the existing +125% FCC modulation limit is counterproductive. In practice, a station adjusts its modulation level to the –100% limit, then clips positive peaks to comply with the +125% limit. Contrary to the original FCC intention, such clipping has only a negative effect on occupied bandwidth. Although the effect is a small one, occupied bandwidth would benefit by removing the artificial restriction on positive peaks.

VII. Transmitter Modulation Measurements

To confirm the FFT results, and to analyze the spectral characteristics of broadcast program material, spectrum analyzer measurements were made using the facilities of nearby AM Station KOFY, San Mateo, California. The station is a one-kilowatt non-directional daytimer with two alternate transmitters, a new all-solid-state model and a much older plate-modulated model. The test configuration is shown in Figure 4. Figure 5 shows how to interpret the spectrum analyzer photographs.

Initial tests were made using both the dummy load and the antenna. Due to the wide bandwidth of the single tower, its test results were virtually identical to those of the dummy load so, as a convenience, the dummy load was used for subsequent tests. More complex antenna systems would produce different amounts of sideband energy than these test results show; most will have a narrower bandwidth than does a single tower so will produce less sideband energy. However, at certain azimuths from a directional antenna the sideband energy may be greater, relative to the carrier, than it would be in a wideband system.[8]

A. Sine Wave Measurements

Figure 6 shows the spectrum analyzer displays for the solid-state transmitter under the following modulation conditions:

a. Unmodulated carrier
b. 50% modulation, 1 kHz tone
c. 100% modulation, 1 kHz tone
d. 100% + 3 dB, 1 kHz tone, carrier pinchoff only
e. 100% + 3 dB, 1 kHz tone, -95% audio clip[9]
f. 100% modulation, 10 kHz tone
g. 100% + 3 dB, 10 kHz tone, carrier pinchoff only
h. 100% + 3 dB, 10 kHz tone, -95% audio clip.

By comparing these measured results with the corresponding figures showing the calculated FFT spectra, it can be seen that the methods correlate well. The sideband amplitudes are similar, and show the equivalence of carrier pinchoff and audio clipping for lower modulating frequencies.

For the 10 kHz modulating frequency, shown in Figures 6f through 6h, audio clipping generates fewer sidebands far from the carrier. This occurs because, although FFT analysis showed that audio and carrier clipping have essentially identical effects, carrier clipping will produce more damaging interference in practice because of limited transmitter high-frequency response. A carrier that is overmodulated with a 10 kHz audio signal is clipped at a 10 kHz rate, generating spurious components at 20 kHz, 30 kHz, 40 kHz, etc. Their amplitudes are limited only by the bandwidth of the antenna. If the same audio signal is clipped prior to reaching the transmitter,

[8] For example, a directional antenna system is usually tuned for minimum radiated field at carrier frequency in a pattern minimum. The null will probably not be as deep at sideband frequencies in a narrowband system, so sideband energy will be enhanced relative to the carrier.

[9] Because clipped sine waves were DC-level-shifted by the AC-coupled transmitter, +125% sine wave modulation was never achieved.

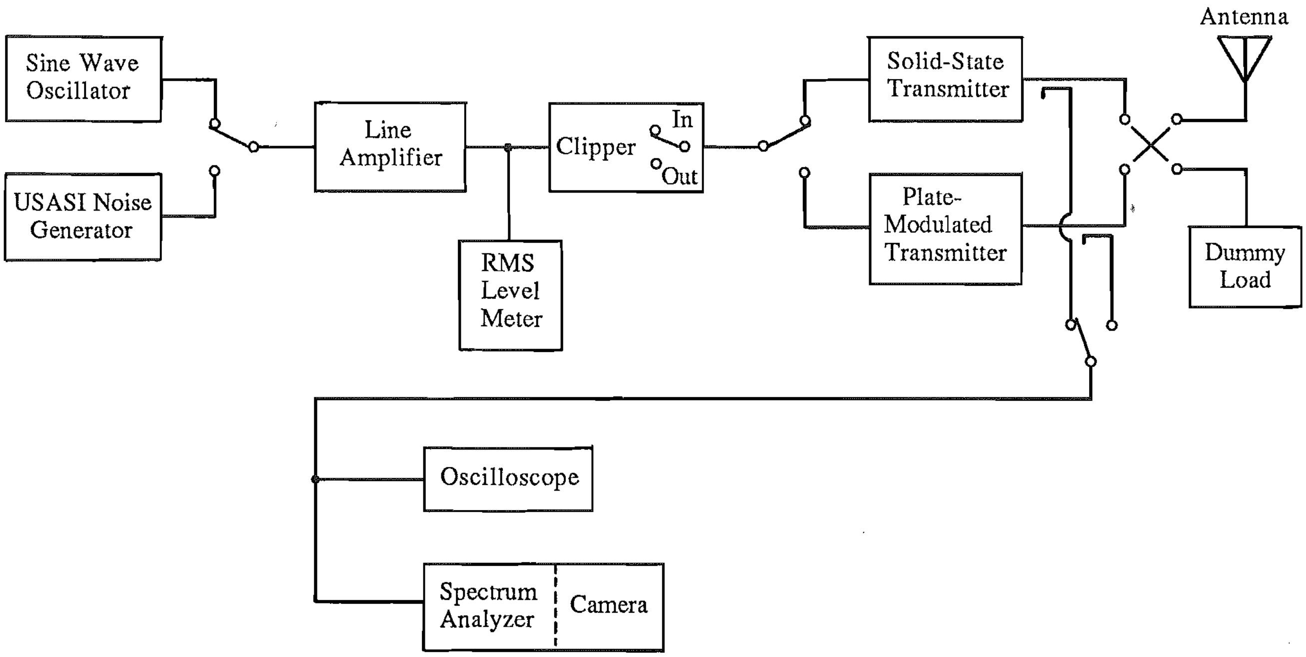


Figure 4. Transmitter Modulation Measurements - Equipment Configuration.

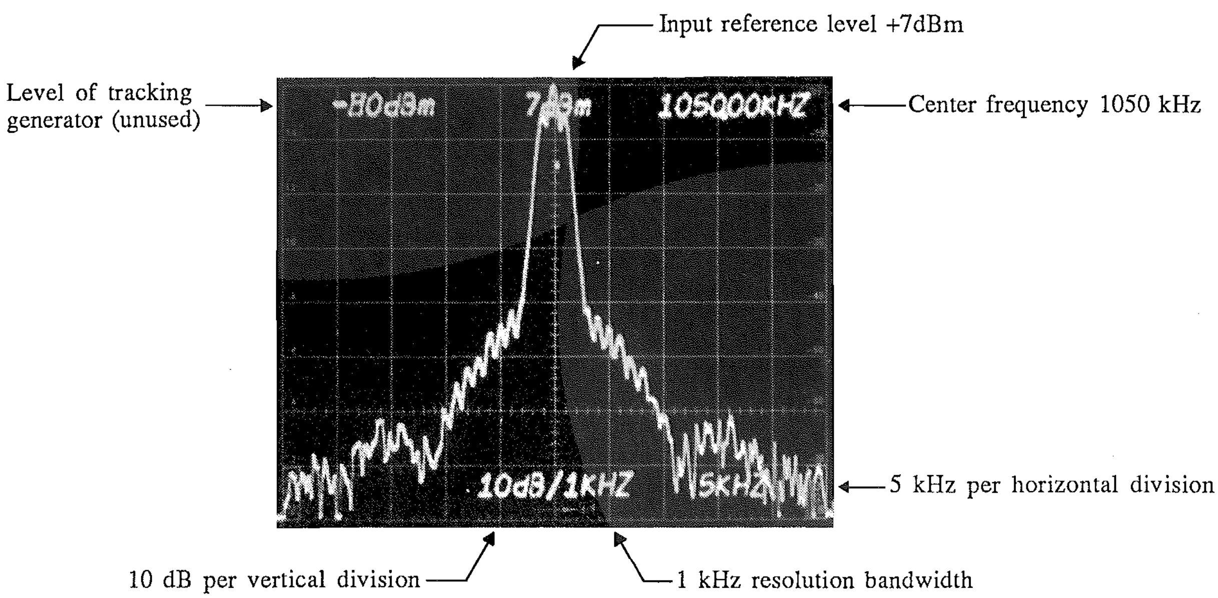
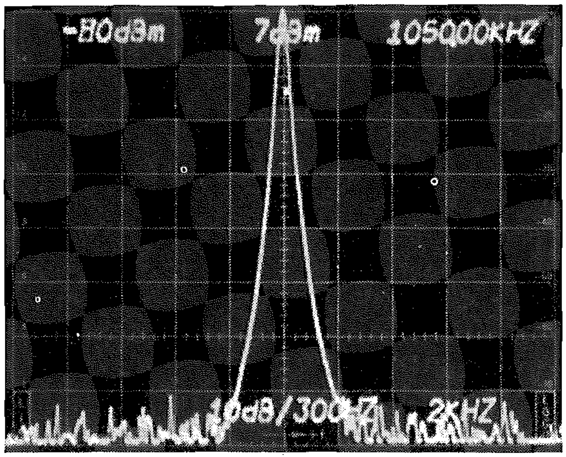
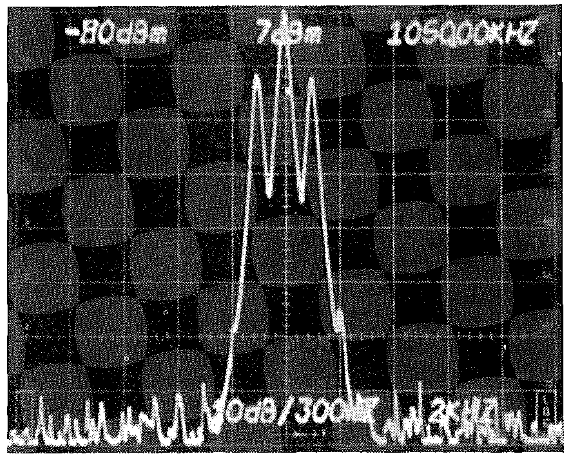


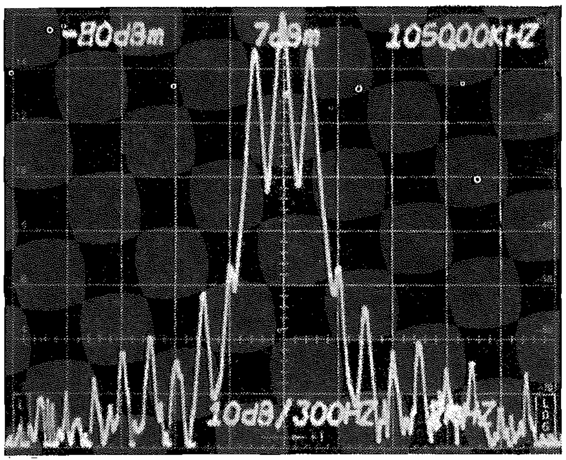
Figure 5. Interpretation of Spectrum Analyzer Photographs.

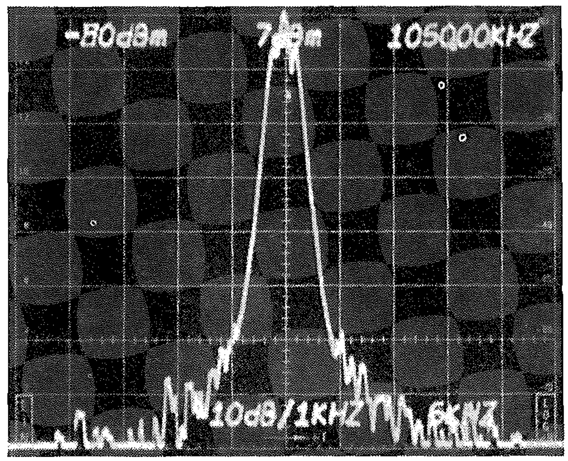


a. Unmodulated carrier



b. 50% modulation, 1 kHz tone

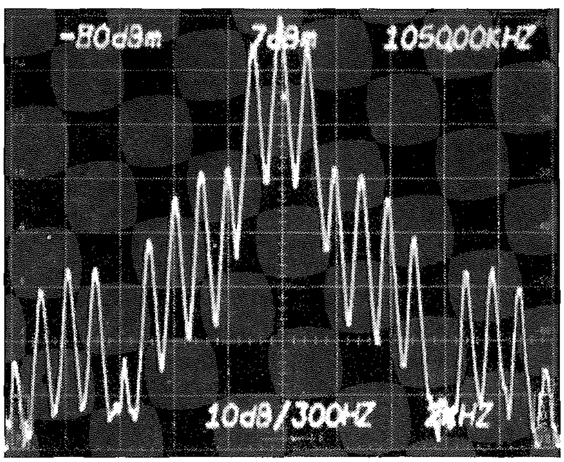


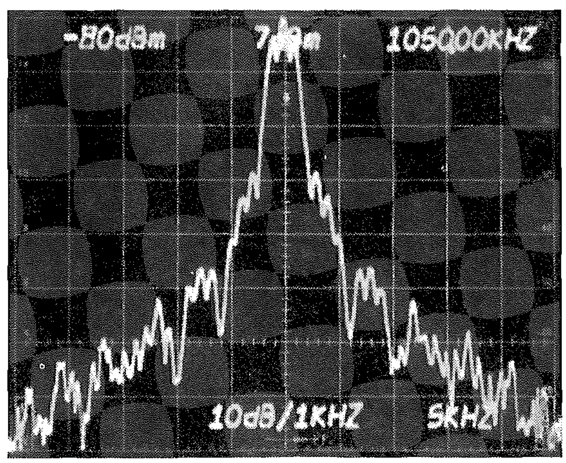


2 kHz/division

5 kHz/division

c. 100% modulation, 1 kHz tone



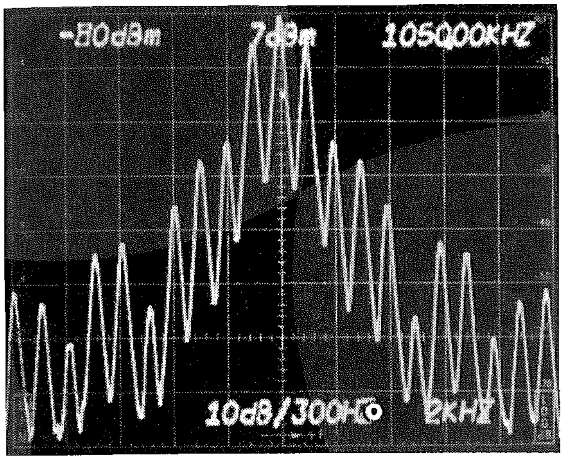


2 kHz/division

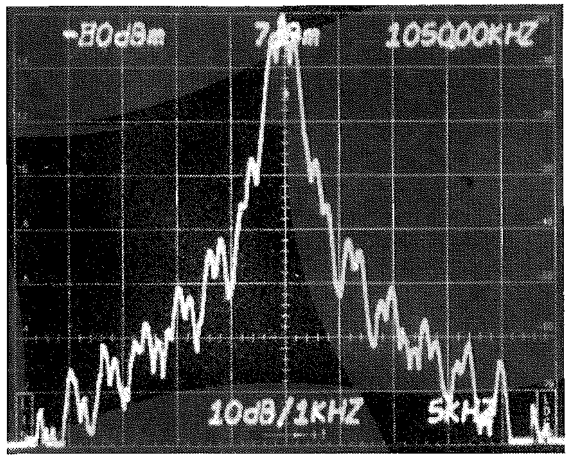
5 kHz/division

d. 100% + 3 dB, 1 kHz tone, carrier pinchoff only

Figure 6. Spectrum Analyzer Displays, Solid-State Transmitter.

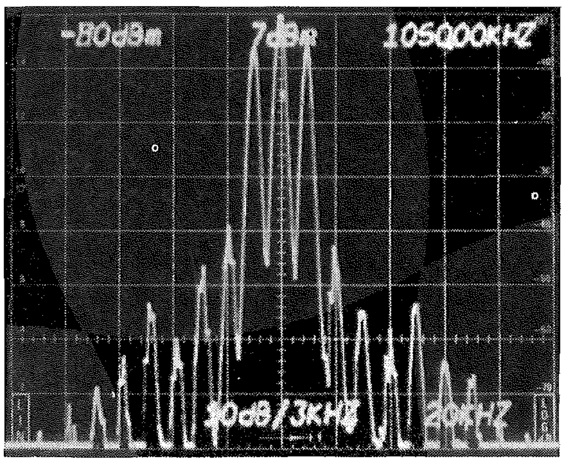


2 kHz/division

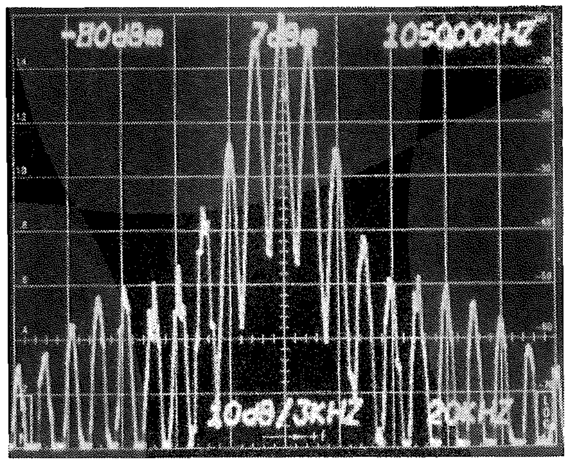


5 kHz/division

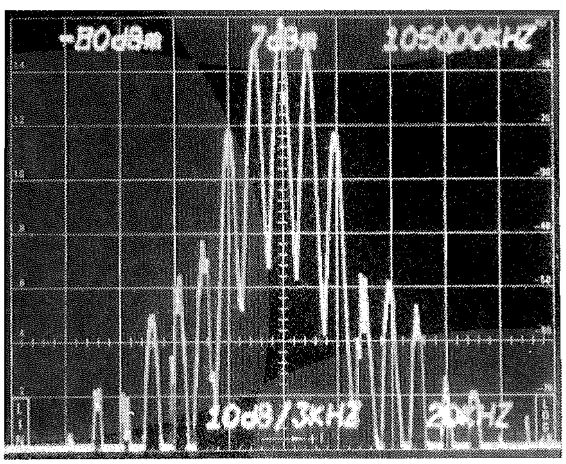
e. 100% + 3 dB, 1 kHz tone, –95% audio clip



f. 100% modulation, 10 kHz tone



g. 100% + 3 dB, 10 kHz tone, carrier pinchoff only



h. 100% + 3 dB, 10 kHz tone, –95% audio clip

Figure 6 (cont.). Spectrum Analyzer Displays, Solid-State Transmitter.

much of its harmonic energy will be rolled off in the transmitter and will not reach the antenna.

Modern transmitters have excellent high-frequency response so, despite the roll-off, audio clipping can produce significant spurious energy at frequencies as much as 40 kHz from the carrier; significant energy from carrier clipping extends even farther. The figures show that, with only a 1 kHz modulating frequency, splatter components up to 20 dB above the noise level occur 17 kHz from the carrier. 10 kHz overmodulation produces such splatter components 80 kHz away.

These figures also strikingly demonstrate that *meeting the FCC occupied bandwidth limitations is no guarantee of a clean signal.* Section 73.44 of the FCC Rules requires emissions from an AM station to be 25 dB below carrier amplitude between 15 kHz and 30 kHz from the carrier, 35 dB below carrier between 30 kHz and 75 kHz from the carrier, and 67-80 dB below carrier (depending on station power) more than 75 kHz from the carrier. Even the worst-case scenario of 3 dB overmodulation at 10 kHz, a condition that would never occur in practice, meets these limits up to 75 kHz.

Figure 7 shows similar spectrum analyzer displays for the plate-modulated transmitter:

a. 100% modulation, 1 kHz tone
b. 100% + 3 dB, 1 kHz tone, carrier pinchoff only
c. 100% + 3 dB, 1 kHz tone, -95% audio clip
d. 100% modulation, 10 kHz tone
e. 100% + 3 dB, 10 kHz tone, carrier pinchoff only
f. 100% + 3 dB, 10 kHz tone, -95% audio clip.

This transmitter had significantly more distortion than the newer solid-state version, and showed some differences in levels of specific sidebands, but its splatter characteristics were not significantly different. Nothing in the test results indicated that old transmitters generated more splatter than did newer ones.[10]

B. Noise Measurements

Sine waves are useful analytical tools with which to gain a basic understanding of overmodulation, but any accurate description of the effects of overmodulation on actual broadcast stations must focus on program material. For these tests, program material was simulated by "USASI" (United States of America Standards Institute) noise. This is a type of weighted noise, developed for sound level meters, with a spectral characteristic that was empirically designed to be similar to average programming. It consists of white noise that has been filtered by a 100 Hz, 6 dB per octave high-pass network and a 320 Hz, 6 dB per octave low-pass network. It has recently been concluded that USASI noise continues to be a close approximation to current music program material.[11] The source for USASI noise in these tests was the General Radio Model 1382 Random Noise Generator. Figure 8, taken from the unit's instruction manual, shows the spectrum of USASI noise compared to white and pink noise.

It was desired to simulate the modulation density of a typical AM radio station. By analyzing the peak and average level characteristics of a number of stations, it has been found that the lowest peak-to-RMS ratio, corresponding to the most heavily processed station, is approximately 6 dB to 9 dB, depending on the integration time of the RMS detector.[12] 9 dB was assumed as an appropriate value for these tests. Broadcast stations have a peak modulation level of approximately 100%; a sine wave has a peak-to-RMS ratio of 3 dB. Therefore, a reference 0 dB noise level for these tests was defined to be 6 dB below the RMS sine wave level needed to modulate the transmitter to 100%.

When applied to the solid-state transmitter, the 0 dB USASI noise produced frequent negative modulation peaks of 100%, with positive peaks exceeding 110%. As the level of the noise was increased by up to 6 dB, the amount of negative overmodulation increased and the positive modulation exceeded 150%. To evaluate the different effects that carrier pinchoff and audio clipping had on occupied bandwidth, the audio clipper was adjusted for –95% and +125% modulation, and the spectrum was compared at several noise levels with and without audio clipping.

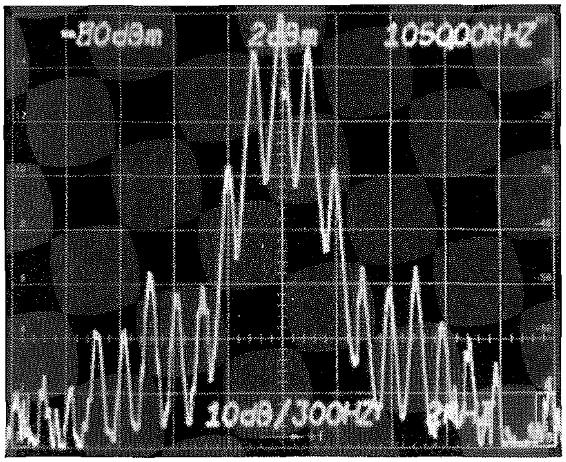
The noise was not preemphasized for the tests. The amount of high-frequency boost used by AM stations varies widely, so any particular choice for a preemphasis curve would have been arbitrary and applicable only to some stations. USASI noise itself is believed to be representative of the spectral characteristics of actual program material, so the actual spectrum expected from a particular radio station can be extrapolated by adding the station's preemphasis curve to the curve shown on the spectrum analyzer.

The noise was also not band-limited for the tests. Although the noise amplitude decreases by 6 dB per octave above 320 Hz, there is still some energy present above 20 kHz, differing from normal programming in which little spectral content exists above 15 kHz. The effect of this difference on the test results is discussed in Section VII–C, below.

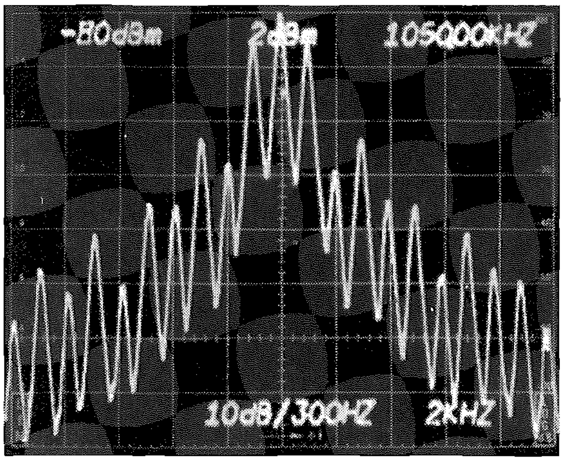
[10] Transmitter Transient Distortion (TTD) has been suggested as a possible cause, in some transmitters, of spurious emissions that are not noticed in steady state measurements. This phenomenon was not investigated during this project. For a discussion of TTD, *see AM Technical Improvement Report*, NAB AM Improvement Subcommittee, October 1984 at 44.

[11] *See* Payne, Christopher P., *The Characterization of Amplitude Response in Audio Systems Employing Program Dependent Variable Equalization,* March 1986, submitted to the National Radio Systems Committee and available from the author at (202) 862-1549.

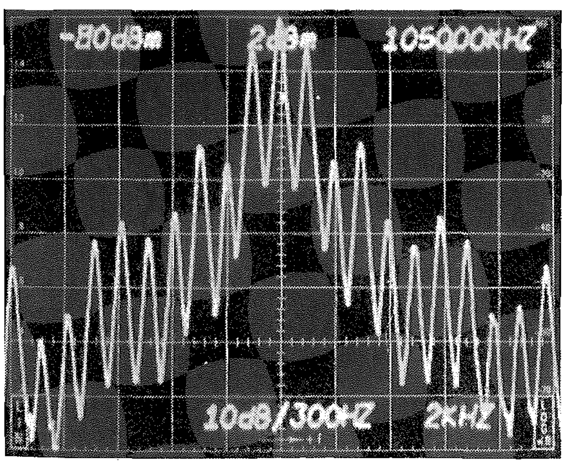
[12] Personal communications with Christopher Payne (id.) and Robert Orban, Orban Associates, Inc., (415) 957–1063.



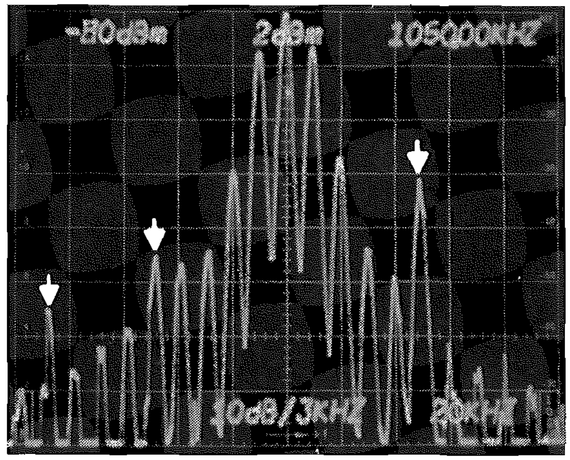
a. 100% modulation, 1 kHz tone



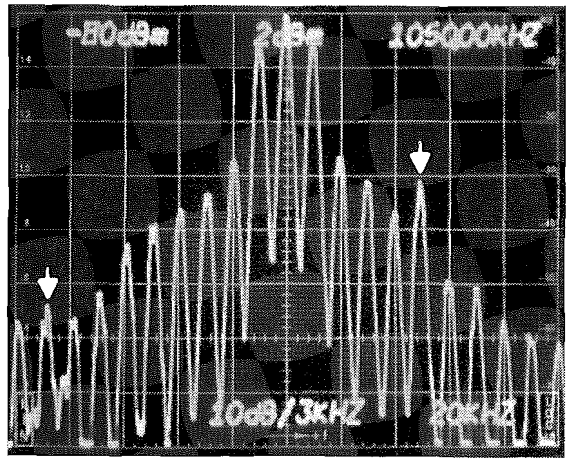
b. 100% + 3 dB, 1 kHz tone, carrier pinchoff only



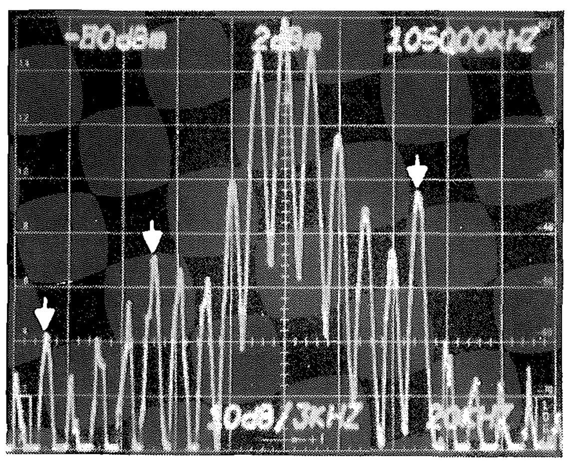
c. 100% + 3 dB, 1 kHz tone, -95% audio clip



d. 100% modulation, 10 kHz tone (note: 960 kHz, 1000 kHz, and 1100 kHz components are other stations)



e. 100% + 3 dB, 10 kHz tone, carrier pinchoff only (960 kHz and 1100 kHz components are other stations)



f. 100% + 3 dB, 10 kHz tone, -95% audio clip (960 kHz, 1000 kHz, and 1100 kHz components are other stations)

Figure 7. Spectrum Analyzer Displays, Plate -Modulated Transmitter.

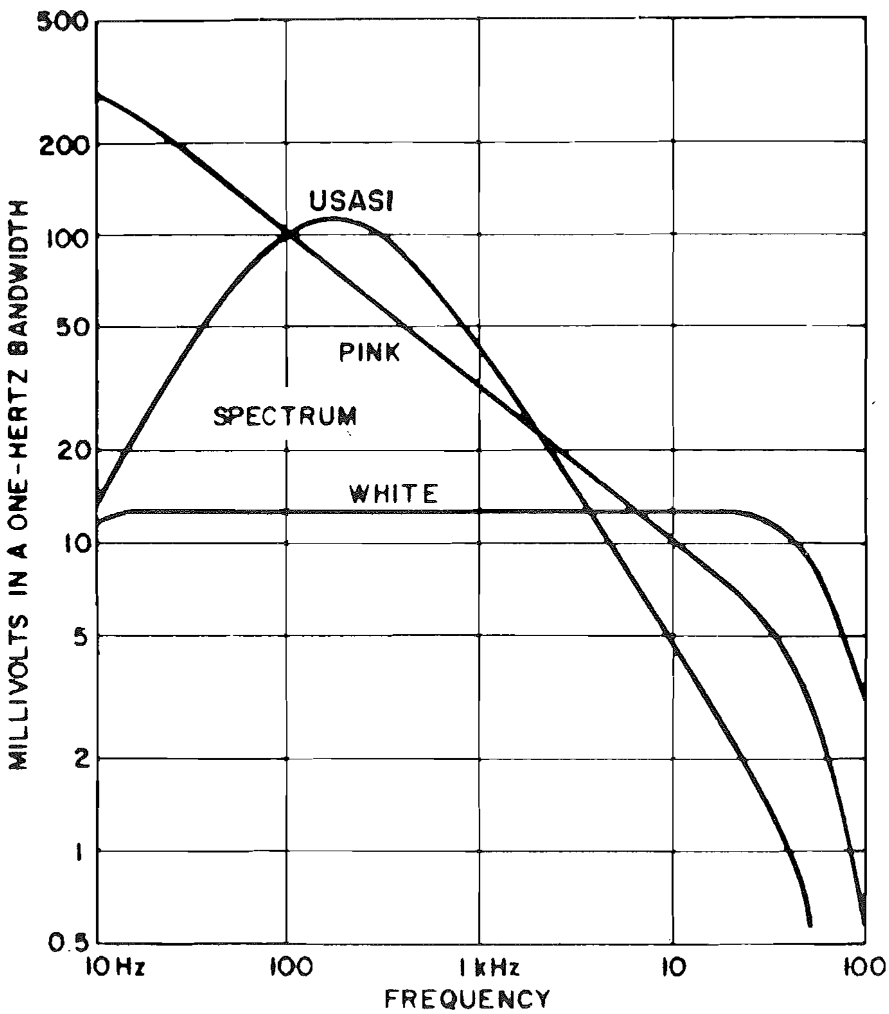


Figure 8. Spectra of White, Pink, and USASI Noise.

Figure 9 shows the spectra of the measured noise-modulation conditions. The spectrum analyzer was adjusted for a display of ±50 kHz, with a resolution bandwidth of 3 kHz. The analyzer was left in the MAX HOLD mode for approximately two minutes, after which virtually no further change in the curve was noted. The reference amplitude at the top of the display was adjusted to be the unmodulated carrier amplitude.[13] Spectrum measurements were made for noise levels of –3 dB, 0 dB, +3 dB, and +6 dB. Since the 0 dB noise reference level was chosen as equivalent to a highly processed station, +6 dB is a true "worst case." It is a far greater average program level than could possibly exist in practice, equal to the level of a constant 100% sine wave.

The spectra shown in Figure 9 are quite surprising. A USASI input noise level of –3 dB (Figure 9a) results in peak modulation levels of less than 100%. A noise level of +6 dB (Figure 9d) results in almost continuous negative overmodulation and positive modulation levels exceeding 150%, and requires heavy audio clipping if overmodulation is to be avoided. Yet the RF spectrum shape is identical at these two extremes, for both carrier clipping and audio clipping conditions. For each 3 dB increase in noise level, the spectrum increases by approximately 3 dB at all frequencies. One might have expected the sidebands of the heavily clipped +6 dB signal to be significantly greater than 9 dB above the level of the lightly modulated –3 dB signal, due to the generation of distortion components at the higher levels. This shows that the spurious components produced by either overmodulation or clipping are masked by the higher frequencies already present in the modulating signal. There is little penalty in occupied bandwidth due to clipping or overmodulation under these modulation conditions.[14]

C. Band-Limited Noise

The use of a low-pass filter on USASI noise, to more closely simulate the spectral characteristics of broadcast program material, would affect the results of these tests. Overmodulation and clipping has such little effect on the measured USASI noise spectra because the spurious components are masked by the high-frequency components already present. Sharp low-pass filtering of the input USASI noise would remove these masking components, making the spurious components more significant.

In particular, Orban has measured the output power spectrum of an Orban 9100A processor, which includes a sharp 12 kHz low-pass filter.[15] Harmonic components caused by the processor's clipper generate a long spectrum "tail" beyond 14 kHz. These components are approximately 45 dB down at 15 kHz, dropping to 80 dB down by 45 kHz. Accordingly, the spurious products of an actual clipped or overmodulated broadcast signal, although still within FCC limits, would be likely to cause greater adjacent-channel interference than would a clean signal without such products. The better the processor is at controlling spurious high-frequencies, the more noticeable will be any splatter caused by post-processor clipping or overmodulation.

D. Occupied Bandwidth Analysis of Noise Modulation

All of the measured spectra in Figures 9a through 9d meet FCC occupied bandwidth criteria, although 10 dB of preemphasis at 15 kHz would cause the -25 dB FCC limit to be exceeded over a narrow band at the 0 dB noise level or higher. This result substantiates our observation on the non-constraining nature of the FCC limits.

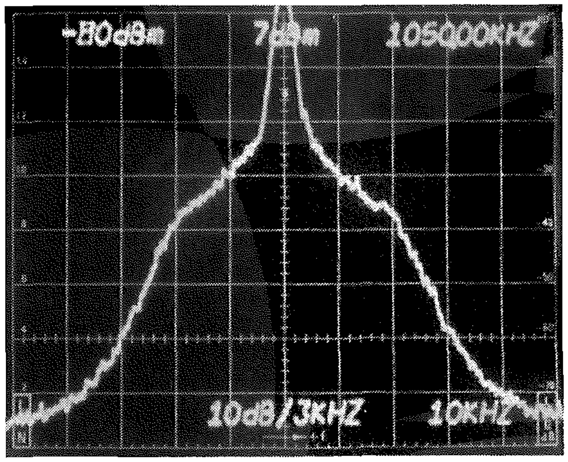
Although only modest audio filtering would be required to meet FCC occupied bandwidth requirements even under these most egregious modulation conditions, it is good engineering practice to seek lower sideband levels at frequencies far from the carrier. As mentioned above, USASI noise contains more energy above 20 kHz than does program audio, so the test spectra appear worse than they would under program conditions. However, to minimize interference to second- and higher-adjacent-channel stations, protective filtering is advised even with program audio. Most popular audio processors contain a low-pass filter with a sharp cutoff characteristic above 12 kHz, so as to be at least 25 dB down at 15 kHz. The National Radio Systems Committee (NRSC) is presently discussing the potential advantages of lower cutoff frequencies in certain allocation circumstances.

It can be concluded from Figure 9 and the discussions in this section that the elimination of overmodulation would have only a modest effect on the present character of the AM band. While overmodulation may create some low-level spurious energy far from the carrier, the primary component of sideband energy in nearby adjacent channels is the energy in the program material. As the high-frequency content of program material has increased, as the amount of preemphasis has increased, as antenna and

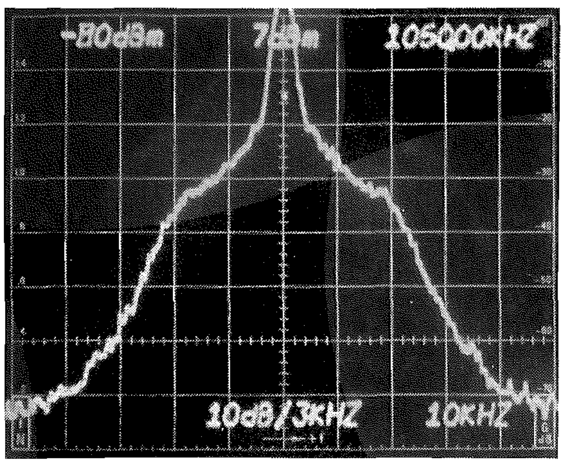
[13] The figures appear to show a carrier amplitude greater than the reference level. This occurs in the MAX HOLD mode because the maximum energy in the 3 kHz resolution bandwidth, which includes the carrier and the lower-frequency sidebands, exceeds the carrier amplitude alone.

[14] It was mentioned in Section VII-A that, for high modulating frequencies, carrier pinchoff produces more splatter than does audio clipping. This is not evident in Figure 9. Therefore, there was some concern that the non-preemphasized noise was not exciting the splatter-generating mechanisms in a realistic way, because the high-frequency components of USASI noise are below the level necessary to cause carrier clipping. However, these figures show that, even with 15 dB preemphasis, a 10 kHz noise component remains below the clipping threshold. It was concluded that non-preemphasized USASI noise was a realistic modulating waveform for this analysis.

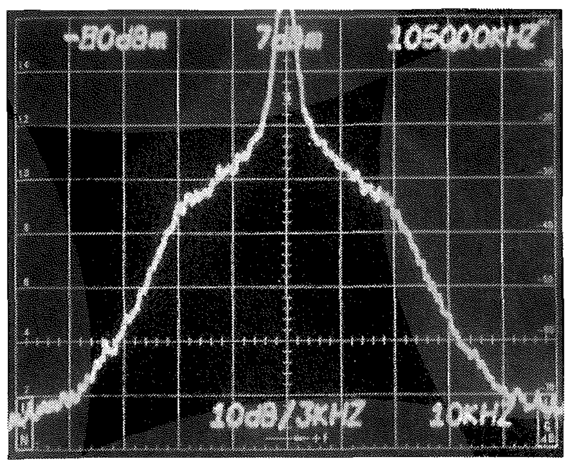
[15] *See Minutes of Subgroup on Methods and Procedures of the National Radio Systems Committee*, Attachment C (May 21, 1986), available from NAB Science and Technology, (202) 429-5346.



a. -3 dB input level (<100% peak modulation)



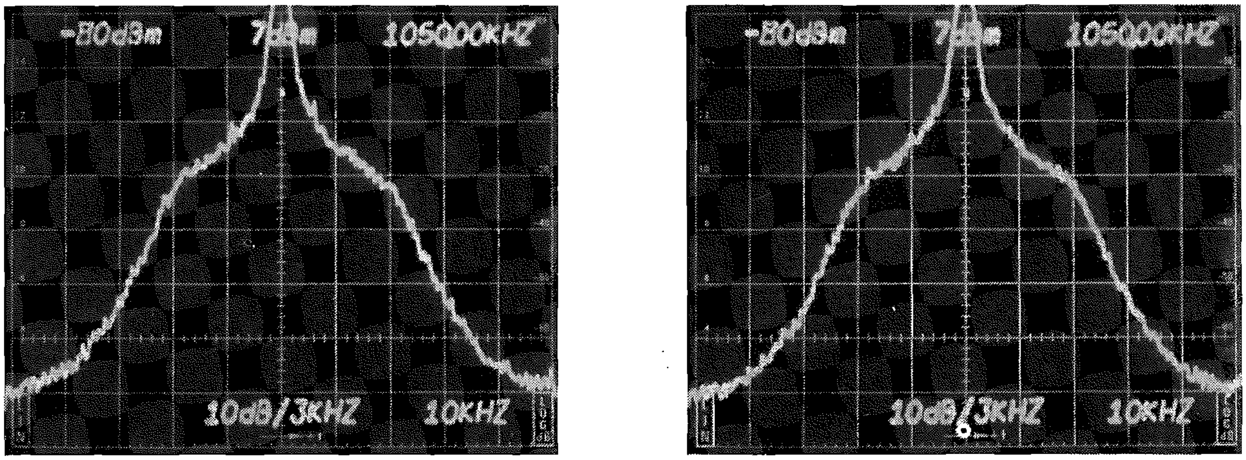
carrier pinchoff only



-95% audio clip (<+125% peak modulation)

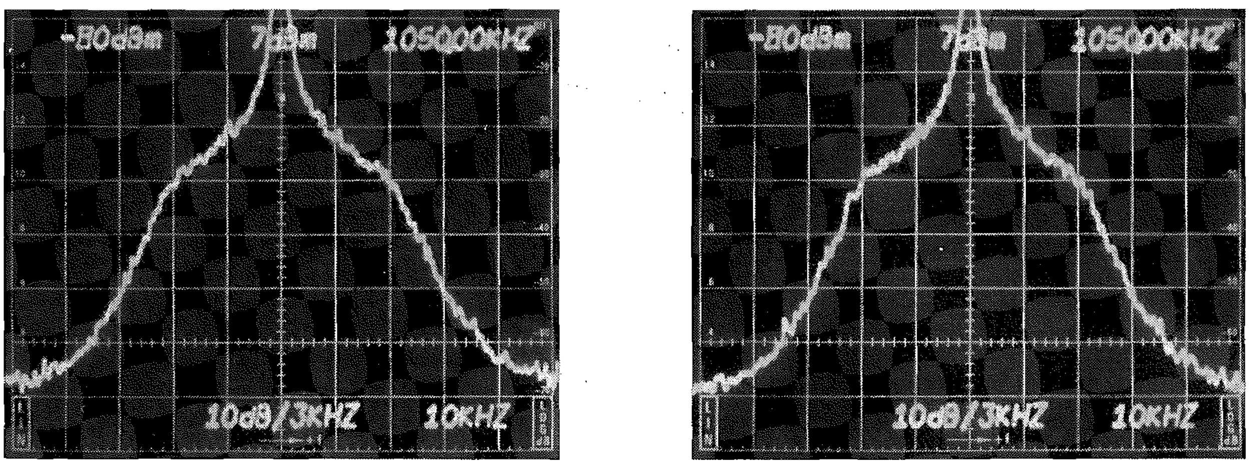
b. 0 dB input level

Figure 9. Spectrum Analyzer Displays, USASI Noise Modulation.



carrier pinchoff only —95%, +125% audio clip

c. +3 dB input level



carrier pinchoff only -95%, +125% audio clip

d. +6 dB input level

Figure 9 (cont.). Spectrum Analyzer Displays, USASI Noise Modulation.

transmitter bandwidths have increased, and as the number of stations has increased, the amount of interference has greatly increased and the limitations of 10 kHz channel spacing have become more apparent. Overmodulation is only a minor component. *Major reductions in splatter interference can only be achieved by reducing the high-frequency energy content of the modulating signal.*

E. DC Level Shift

Overmodulation can be exacerbated by the DC level shift in AC-coupled transmitters. Station engineers often meticulously adjust their processing to reach –98% modulation, only to find that overmodulation occurs with different program material. Even if the peak levels leaving the processor are tightly controlled, AC-coupling can shift them by several percent as the asymmetry of the program material varies. This can be enough to cause overmodulation in an otherwise well-designed audio processing system.

DC level shift is a common problem in television but is rarely addressed in radio. Several stations have modified their AM transmitters to full DC-coupled operation. Other engineers have experimented with a DC clamp that stops the final amplifier of the transmitter from cutting off the carrier; while this does not prevent DC level shift, it eliminates its negative effects. These techniques are effective but of limited use. DC-coupling can only be accomplished in some types of transmitters; plate-modulated transmitters, for example, cannot be DC-coupled. In addition, many engineers are uncomfortable with the existence of a DC path at the audio input.

Fortunately, the problem can be essentially solved, without the necessity for complete DC-coupling, by improving the low-frequency response of the transmitter. It has been suggested that a reduction of the lower 3 dB cutoff frequency to 0.1 Hz should be sufficient.[16] Transmitter engineers should develop field modification kits for such operation and include this feature in newer models.

F. Minimizing Occupied Bandwidth

As a result of the measurements and analysis of sine-wave and noise modulation, the following steps are recommended to minimize excessive occupied bandwidth:

1. All audio processing equipment should contain or be followed by an appropriate overshoot-corrected low-pass filter to minimize spectral components above the desired audio range.
2. Because high-frequency audio clipping produces less splatter than does high-frequency carrier clipping, a protective audio clipper is advised as the last device before the transmitter. Such a clipper is often contained in modern audio processors; if so, it is preferable to a separate device. If a separate device is necessary, it can be a stand-alone unit or may already be built into the transmitter input circuitry. The clipping point of a separate device should be approximately -95%, to insure that isolated high-frequency peaks passed by the audio processing system do not cause carrier pinchoff in the transmitter, and the audio output level from the processor should be just below the level at which clipping occurs. *The protection clipper must not be used to increase loudness.* New clippers should be designed to provide an indication of clipping amount so that excessive clipping can easily be recognized.
3. To minimize DC level shift, which can cause unwanted carrier clipping of audio signals having even well-controlled peak levels, stations should investigate reducing the low-frequency cutoff point of their transmitters to approximately 0.1 Hz, or converting their transmitters to DC-coupled or DC-restored operation. Transmitter engineers should develop field modification kits for such operation for older transmitters and include these features in newer models.

VIII. Amplitude and Phase Errors

A. Envelope Distortion

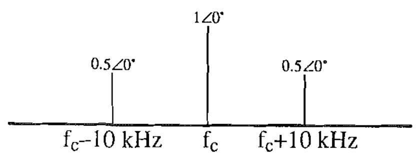
The limited-bandwidth circuitry in a transmitter output network or antenna system can change the amplitude and phase of the spectral components of an amplitude-modulated signal. Because these amplitude and phase errors can affect the apparent modulation level and can cause distortion in receivers, it is easy to mistakenly conclude that antenna systems can cause transmitter-type overmodulation and splatter. This is not correct. Transmitter-type overmodulation is a non-linear process that does not occur in linear electrical networks made up of only inductors, capacitors, and radiation resistance. These elements can affect the amplitude and phase of a spectral component, but they cannot create new frequencies.

Although RF networks do not cause splatter, their effects on the signal can be significant. Unless understood, these effects can lead station engineers to make changes in their audio processing that distort the signal for most listeners. It is important that station engineers understand the actual mechanism by which these effects occur so they can properly make compensating adjustments or can recognize the need for design changes.

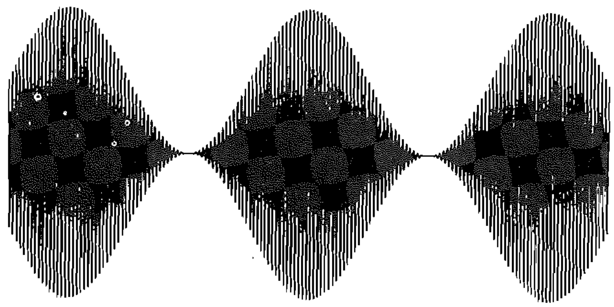
FFT techniques can be used to graphically demonstrate these effects. Figure 10 shows the effect that a limited-bandwidth antenna can have on the spectrum of an AM signal.[17] The first figure shows the theoretical three-component spectrum when the carrier is 100%-modulated with a 10 kHz tone; the sideband components are 6 dB

[16] Personal communication with Robert Orban (id.).

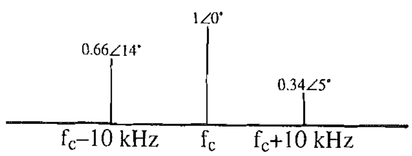
[17] These figures were derived from the measured tower currents and phases of an actual two-tower directional antenna.



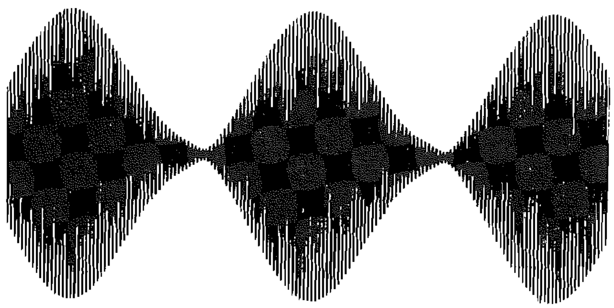
a. Theoretical Spectrum



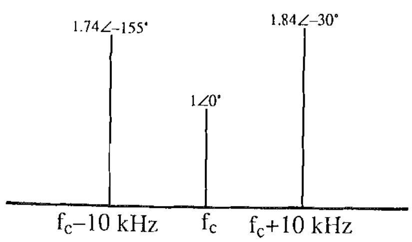
a. Theoretical Waveform



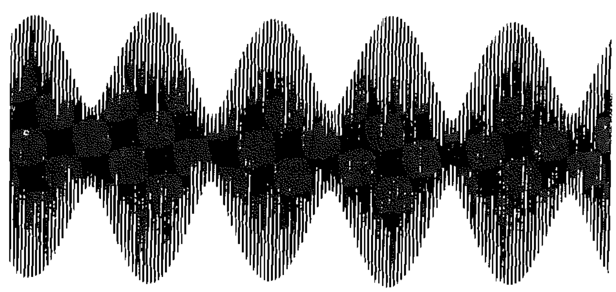
b. Spectrum in Major Lobe



b. Waveform in Major Lobe



c. Spectrum in Pattern Minimum



c. Waveform in Pattern Minimum

Figure 10. Spectrum Effects of Limited-Bandwidth Antennas.

Figure 11. Waveform Effects of Limited-Bandwidth Antennas.

below the carrier and the three components are in phase. The second and third figures show the actual spectra that would be found in the major lobe and in the pattern minimum, respectively. The amplitudes and phases of the sideband components have been normalized for a carrier component of $1\angle 0°$; they show significant differences from the theoretical spectrum.

To show what effect these amplitude and phase distortions have on the received signal, these spectra were converted to their time-domain waveforms with an inverse Fast Fourier Transform. The waveforms, shown in Figure 11, are those that would be detected by an AM receiver in the field. The amplitude and phase distortions in the frequency domain have been converted to envelope distortions in the time domain. In the major lobe, the envelope still approaches 100% modulation but some distortion is visible. In the direction of the array minimum, the fundamental 10 kHz frequency of the modulating waveform is not even discernable; only a distorted second harmonic of reduced modulation percentage is apparent.

This envelope distortion was quantified by digitally synthesizing an envelope detector and using the detected waveforms as inputs to a direct FFT. The output spectrum of each transform, containing the 10 kHz audio fundamental together with the various harmonic distortion components, was converted to a Total Harmonic Distortion (THD) figure. THD in the major lobe is approximately 5%, while in the pattern minimum it is over 1500%![18] Although there were no undesired frequency components created in the transmitted RF spectrum by the amplitude and phase errors in the antenna system, there were many such components created in the envelope-detected audio spectrum. While these distortion components can be harmful to the station's audio quality, and may sound similar to RF splatter, they do not cause interference to other stations.

B. Modulation Nonlinearities

Amplitude and phase distortion can also result in apparent transmitter overmodulation where none actually exists. Figure 12 is an example of this phenomenon. The antenna system is driven with 80% modulation. In this example, the system has an asymmetric impedance characteristic that boosts the upper sideband while leaving the lower sideband unchanged. The result is a distorted envelope as observed in the far field, with an apparent modulation of 100%. THD is approximately 15%. If the modulation level as observed at the transmitter were increased to 100%, the signal as observed in the far field would have the folded-over waveform of multiplier-type overmodulation and would be greatly distorted in an envelope detector. An engineer with such an antenna system would notice excessive distortion in the field even if the transmitter were modulating normally.

Potentially more damaging to other stations is the case where the modulation level in the far-field is less than that at the transmitter, causing a loudness loss in the field. If the transmitter audio input level were raised in an attempt to increase the far-field modulation, the transmitter would overmodulate. There would be no significant peak modulation increase in the field, although the carrier clipping would cause the average level to increase. The net effect would be a minor increase in average loudness accompanied by a large increase in distortion and possibly adjacent-channel interference.

C. Previous Papers

Most of the effects described above have been recognized for many years and were addressed in depth by Doherty, Moulton, and more recently by Bingeman and Clarke. Doherty, in his classic paper attached as Appendix C, showed how the modulation envelope varies depending on the impedance characteristic at the modulation monitoring point and on whether a voltage or a current sample is taken.[19] He found that modulation percentage measurement required different monitoring point criteria than did sideband power or distortion measurement, and he described how to select these points. He also developed the well-known "line-stretching" technique to rotate the antenna impedance characteristic so that the transmitter output tube sees a symmetrical load.

Moulton, whose paper is attached as Appendix D, published a comprehensive collection of data showing how frequency response, distortion, and square-wave response differed at varying azimuths from a directional antenna. He also showed an example of a folded-over multiplier-type waveform in the minimum of an actual directional antenna.

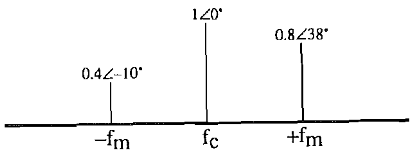
Bingeman and Clarke, in the paper of Appendix E, described a computer technique that quantitatively relates bandwidth to modulation percentage and THD. With this technique, the antenna designer can optimize RF networks to minimize envelope distortion and the difference between transmitter and far-field modulation percentage.

D. Evaluating Station Antenna Performance

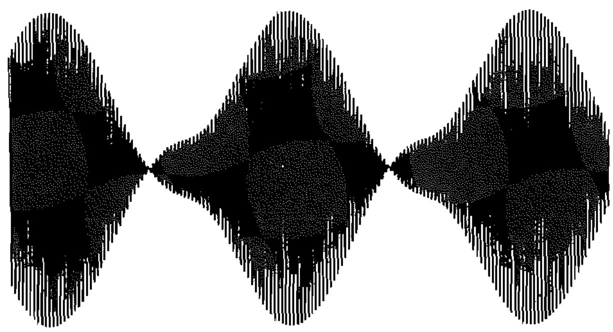
A station that experiences distortion on its signal in the field, even though the modulation monitor at the transmitter produces clean audio, or that notices different modulation levels in the field than at the transmitter, is likely to be suffering from antenna system amplitude or phase errors. These errors cannot always be detected from the common point impedance plot of a directional antenna system. Although amplitude and phase errors will be most significant in a narrowband antenna system, they can still occur even in an antenna that has apparently sufficient

[18] The 10 kHz distortion is extremely high because the fundamental component is more than 20 dB below the second harmonic.

[19] The impedance characteristic differs at different points within the antenna phasing and matching circuitry and along the transmission lines.



a. Spectrum



b. Waveform

Figure 12. Example of Apparent Overmodulation.

bandwidth.[20] For example, Bingeman and Clarke (ibid.) describe a relatively broadband antenna system with poor modulation linearity and THD, which was significantly improved by modifying the input matching network.

There are a number of ways for a station to evaluate its performance in this regard. Using an oscilloscope and a modulation monitor, it can examine its modulated envelope in the field at various azimuths while transmitting a cleanly modulated high-frequency tone. Significant discrepancies between the modulation percentages and waveforms seen at the station and those seen in the field are signs of antenna amplitude and phase errors. The station can also compare the outputs of a synchronous detector with an envelope detector at suspect locations. A problem would be indicated if distortion were present in the envelope detector but not in the synchronous detector. The station could also perform the kind of theoretical analysis described in the Bingeman and Clarke paper for various field locations.

IX. Monitoring of Modulation and Occupied Bandwidth

As described in Section VII, extraneous sideband energy can be minimized through proper audio processing system design, but improperly designed or adjusted equipment can cause interference. Unless stations have the ability to properly monitor their modulation characteristics, they have no way of assuring themselves that they are operating as intended. Accurate monitoring equipment and techniques are essential if occupied bandwidth is to be minimized.

A. Transmitter Monitoring

The modulation monitoring point necessary to adjust peak transmitter audio input levels for less than 100% negative modulation must provide an accurate sample of the modulation envelope voltage at the modulated stage of the transmitter, since carrier clipping occurs there. Monitoring the voltage or current waveshape at the wrong point in the transmitter or antenna tuning networks can give an inaccurate indication of the envelope, perhaps showing a lower modulation level than actually exists at the point of modulation. Splatter and distortion would be generated if the modulation level were increased to compensate.

Most transmitters have a modulation monitoring tap for this purpose, but some may not provide accurate voltage envelope samples. A station can evaluate whether its monitoring tap is appropriate by examining the modulation envelope with an oscilloscope while modulating the transmitter with a sine wave. As the input level is increased to 100% modulation and beyond, the carrier should cleanly disappear during negative peaks. If the negative sine wave peaks distort before reaching 100%, or if the negative peaks fold over and then become distorted, an unsuitable monitoring point is indicated. The techniques in Doherty (ibid.) can then be used to select an alternative point that does provide an accurate envelope voltage sample.

B. Need for Field Monitoring Improvements

A station often needs to accurately monitor its modulation and occupied bandwidth at the studio or another field location. It may also need the ability to monitor the signals of other stations. However, it is very difficult to determine a station's modulation percentage in the field or to determine whether the station is in compliance with FCC modulation and occupied-bandwidth limits. Modulation percentage readings on either an oscilloscope or a conventional modulation monitor can be erroneous due to the envelope distortions previously described. While occupied bandwidth can be accurately measured with a conventional swept-filter RF spectrum analyzer if the modulating waveform is noise, such measurements are inaccurate on program material because the filter may miss the transients that are the primary sideband components. If the signal is envelope-detected and analyzed for extraneous audio components, the splatter that might have been transmitted cannot be differentiated from the distortion components that are generated in the envelope detector.

To help solve these monitoring problems, the broadcast industry should develop a high-quality precision demodulator using a synchronous detector. A precision demodulator would allow a broadcast station to accurately and consistently monitor its modulation characteristics throughout its coverage area. The synchronous detector would eliminate envelope distortion, so the received signal would remain undistorted regardless of the monitoring location or the bandwidth characteristics of the antenna system. Any distortion or other spurious components present in the detected audio would be from the station itself.

C. Accurate Occupied Bandwidth Measurements

A synchronous demodulator would make accurate measurements of occupied bandwidth possible, even in a null of a directional antenna. There have been several suggestions concerning such measurements. One was to construct a filter whose shape is as close as possible to the inverse of the FCC occupied bandwidth curve.[21] The output of the demodulator would pass through this filter. Any filter output of greater than the 0 dB reference level

[20] A narrowband antenna is generally apparent from the shape of the plate load impedance curve or the antenna system input impedance curve; examples are a curve which has a complex shape, whose impedance varies greatly across the channel, or which is asymmetric at the plate of the final amplifier.

[21] Personal communication with Robert Orban (id.). For a station with five kilowatts of power or more, this filter would have ideally infinite attenuation below 15 kHz, then would be –55 dB relative to unmodulated carrier from 15 kHz to 30 kHz, –45 dB from 30 kHz to 75 kHz, and 0 dB above 75 kHz. Although such a filter is not physically realizable with such steep slopes, a reasonable approximation could be designed.

would correspond to an emission greater than the FCC limit.

Another suggestion was to analyze the output of the precision demodulator with an audio-frequency FFT analyzer. Such a device uses the FFT techniques of this report on an actual audio signal to digitally compute the spectrum. The audio input is sampled for a predetermined length of time; the computed transform then shows all frequency components of the signal during that time window. With repeated samplings, the maximum occupied bandwidth is accurately determined. This type of analyzer is superior for audio purposes to a swept-filter spectrum analyzer because it captures transients that the filter would probably miss. At present, FFT analyzers are uncommon and quite expensive, but are expected to become more easily available in the future.

A synchronous demodulator would serve other purposes as well. It could be used as a high-quality audio source for the station's monitor system or could be coupled to an audio analyzer for use during proofs-of-performance or other audio tests. If the demodulator were designed with both in-phase and quadrature ("I" and "Q") outputs, the Q output would provide a convenient indication of incidental phase modulation (IPM) in the transmitter. Minimizing IPM is essential for stereo performance, and stations have also found it to be important in improving monaural audio quality.

Accompanying any new occupied bandwidth measurement capability should be a fundamental review of occupied bandwidth measurement procedure, because it is not now well-defined in the FCC Rules.[22] A modulated spectrum is dynamic, with a constantly changing energy distribution. Occupied bandwidth measurements are indications of energy density, so are dependent on whether peak or average energy is measured and on the bandwidth of the measuring device. If more emphasis is to be placed on occupied bandwidth in the future, more rigorous definitions will be required. An appropriate organization such as the National Radio Systems Committee should begin this review.

D. Modulation Percentage Measurement Limitations

To avoid carrier clipping, it would be useful to have an accurate indication in the far field of modulation percentage as observed at the transmitter. Unfortunately, the synchronous demodulator does not provide this capability. Modulation percentage is strictly an envelope parameter, a function of relative envelope amplitudes. Amplitude and phase errors, which are inevitable in the AM transmission process, distort the envelope and make it essentially impossible to determine the transmitted modulation percentage in the field. Stations should recognize that differences between modulation percentage readings in the field and those at the transmitter are perfectly normal. Only the modulation percentage at the proper transmitter monitoring point, described in Section IX-A above, is significant.

[22] Section 2.202(a) of the FCC Rules defines occupied bandwidth as the "frequency bandwidth such that, below its lower and above its upper frequency limits, the mean powers radiated are each equal to 0.5 percent of the total mean power radiated by a given emission." However, the Rules are silent on the procedure for measuring occupied bandwidth.

X. Conclusions and Recommendations

The material in this report can be used by AM engineers and equipment manufacturers to take actions that will reduce excessive occupied bandwidth in the AM band. By knowing when and how splatter is generated and how to accurately measure modulation conditions, engineers can make appropriate improvements in the design and adjustment of their AM transmitting facilities. These improvements may benefit the station's sound by reducing distortion and listener fatigue. Most importantly, if stations make widespread use of this material to minimize their occupied bandwidth, and if receiver manufacturers become convinced that a high-quality interference-free signal will be increasingly available, it will be in the self-interest of the manufacturers to supply receivers that can accurately reproduce this signal.

The most important conclusions and recommendations of this report are as follows:

1. The primary cause of splatter is excessive high frequency content in modulating audio. This is the single most important conclusion to be drawn from this report. If the modulation contains excessive high frequencies, these will cause splatter. Traditional "overmodulation," meaning the pinchoff of the carrier, is undesirable, but is much less significant than the audio itself. Carrier disappearance, commonly believed to be the significant cause of splatter, has little to do with it.
2. FCC occupied bandwidth limits are bare minimums; they are no guarantee of clean operation. Even egregious modulation practices will normally meet FCC requirements. Stations should strive for occupied bandwidths much narrower than FCC limits. The FCC should consider tightening its limits to improve adjacent-channel interference protection.
3. AM stations can minimize excessive occupied bandwidth through the design of their audio processing systems. All audio processing equipment should contain or be followed by an appropriate overshoot-corrected low-pass filter. A final protective clipper is advised as the last device before the transmitter, preferably as an integral part of the audio processor; this clipper must not be used to increase loudness. To eliminate DC level shift, stations should investigate improving the low-frequency response of their transmitters to approximately 0.1 Hz or less, and transmitter engineers should provide

this feature as a field modification and in new models.

4. Amplitude and phase errors in the transmitter and antenna tuning networks can distort the envelope of an AM signal, changing its apparent modulation percentage. To prevent carrier pinchoff, a station must insure that it can accurately monitor the modulated envelope as it exists at the modulated stage of the transmitter. An appropriate modulation monitoring tap is often provided by the transmitter manufacturer, but it may be necessary to choose a different monitoring point using the techniques of the Doherty paper.
5. Envelope distortion due to amplitude and phase errors can affect the quality of the signal received by listeners. Stations should evaluate antenna system modulation performance through field measurements or analysis. Transmitter modulation adjustments should be tailored accordingly; undermodulation might be required at the transmitter to prevent apparent overmodulation in the field. If poor modulation linearity or high envelope distortion is apparent, antenna system improvements should be investigated.
6. A high-quality synchronous detector AM demodulator is needed by the AM industry. Such a demodulator would avoid the envelope distortion caused by amplitude and phase errors and would permit accurate analysis of modulation characteristics in the field. More accurate occupied bandwidth measurements would be possible using such a demodulator. The National Radio Systems Committee or another appropriate organization should begin to develop an improved definition for occupied bandwidth.

For a number of years, AM broadcasters have focused on audio processing and loudness with little concern for the impact of their actions on other stations or on receiver manufacturers. Combined with some basic misconceptions about the causes of splatter, this loudness race has often resulted in signals which sounded good to the stations themselves but wreaked havoc with receivers and other stations. With the current interest in AM improvement, it seems that the entire industry would benefit if each AM station reviewed its operating practices in light of this report.

ACKNOWLEDGEMENTS

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APPENDIX A

DERIVATION OF SYNCHRONOUS DETECTION CHARACTERISTICS

The expanded equation for an amplitude-modulated carrier, modulated with a single tone and passing through a transmission path containing amplitude and phase errors, is

$$e(t) = \cos \omega_c t + A_L \cos [(\omega_c - \omega_m)t + \phi_L] + A_U \cos [(\omega_c + \omega_m)t + \phi_U],$$

where ω_c = carrier frequency,

ω_m = modulating frequency,

A_L = amplitude of lower sideband,

ϕ_L = phase of lower sideband,

A_U = amplitude of upper sideband,

and ϕ_U = phase of upper sideband.

For simplicity we assume a carrier amplitude of 1.

To synchronously detect this signal, it is multiplied by $\cos \omega_c t$.

Because $e(t)$ is a linear summation of three terms, we can multiply each term by $\cos \omega_c t$, using the trigonometric identity

$$\cos A \cos B = 1/2 \cos (A + B) + 1/2 \cos (A - B),$$

and sum the results. For the carrier term,

$$\cos \omega_c t \cos \omega_c t = 1/2 \cos 2\omega_c t + 1/2 \cos 0 .$$

This contains only high frequency and DC terms, which are removed by receiver filtering so can be ignored.

For the lower sideband,

$$\begin{aligned} &\cos \omega_c t \,(A_L \cos [(\omega_c - \omega_m)t + \phi_L]) \\ &= A_L \cos \omega_c t \cos (\omega_c t - \omega_m t + \phi_L) \\ &= 1/2\, A_L [\cos (\omega_c t + \omega_c t - \omega_m t + \phi_L) + \cos (\omega_c t - \omega_c t + \omega_m t - \phi_L)] \\ &= 1/2\, A_L [\cos (2\omega_c t - \omega_m t + \phi_L) + \cos (\omega_m t - \phi_L)]. \end{aligned}$$

Eliminating the left hand high frequency term, the detected lower sideband is

$$1/2\ A_L \cos(\omega_m t - \phi_L).$$

Repeating the same process for the upper sideband yields

$$1/2\ A_U \cos(\omega_m t + \phi_U).$$

Summing the two terms gives the resultant detected signal

$$1/2\ A_L \cos(\omega_m t - \phi_L) + 1/2\ A_U \cos(\omega_m t + \phi_U),$$

which is a pure cosine wave at the modulating frequency, whose amplitude depends on the amplitudes and phases of the individual sideband components.

The detected wave could disappear with certain combinations of phase shifts, but there are no distortion components present. If one of the two sidebands is filtered out at RF frequencies, the phase sensitivity is eliminated, and only amplitude changes will affect the detected output. For a transmission path with no amplitude or phase errors, $A_L = A_U = 1$ and $\phi_L = \phi_U = 0$, so the detected output is

$$1/2 \cos \omega_m t + 1/2 \cos \omega_m t = \cos \omega_m t ,$$

which is an exact replica of the modulating signal.

APPENDIX B

FAST FOURIER TRANSFORM PARAMETERS

Throughout the report, FFT examples use a carrier frequency of 625 kHz, with a fundamental modulating frequency of 9765.625 Hz. This corresponds to a carrier period of 0.0016 ms, and a modulating period of 0.1024 ms. These numbers were chosen to represent realistic AM frequencies and to facilitate the computation of direct and inverse Fast Fourier Transforms. With a 1024-point transform, each sample corresponds to 0.0001 ms. The full transform period of 0.1024 ms is one complete cycle of the modulating frequency and 64 complete cycles of the carrier frequency. This choice of frequencies is necessary to prevent undesirable transform artifacts from coloring the output data.

The sideband amplitudes determined by the FFT calculations are valid not only for the frequencies used in the calculations, but for all single modulating frequencies of an AM carrier. For example, if a certain modulation condition resulted in an FFT sideband component of –20 dB at $f_c + 4f_m$ (39,062.5 Hz above the carrier, or 664.0625 kHz), the same modulation condition with carrier and modulating frequencies of 1000 kHz and 1000 Hz, respectively, would also yield a sideband component of -20 dB at $f_c + 4f_m$ (4000 Hz above the carrier, or 1004 kHz).

APPENDICES C, D, E are attached.

Appendix C

Doherty, W.H., "Operation of AM Broadcast Transmitters into Sharply Tuned Antenna Systems," Proceedings of the I.R.E., July 1949, pgs. 729-734.

Operation of AM Broadcast Transmitters into Sharply Tuned Antenna Systems*

W. H. DOHERTY†, FELLOW, IRE

Summary—**The impedance of some broadcast antenna arrays varies so much over the transmitted band as to impair the performance of the radio transmitter. The impairment consists in clipping of sidebands and distortion of the envelope at high modulation frequencies. This paper reports on an experimental determination of the nature and magnitude of this impairment and on its substantial reduction by suitable coupling methods.**

Introduction

THE IMPEDANCE of a broadcast antenna, and particularly the common-point impedance of an array, often varies widely over the transmission band. When this is the case, the frequency response, amplitude linearity, and modulation capability of the broadcast transmitter can be adversely affected. Recognition of this difficulty by transmitter manufacturers has led to the formulation of an RMA specification for the "normal load" into which a transmitter should operate and meet its performance requirements. This is a load whose resistance does not depart more than 5 per cent from its midband or carrier-frequency value at ±5 kc or 10 per cent at ±10 kc, and whose reactance, which is zero at midband, does not exceed 18 per cent of the midband resistance at ±5 kc or 35 per cent at ±10 kc.

A study of the effect of frequency-sensitive loads from the viewpoint of the transmitter designer has been carried out by engineers of the broadcast transmitter development group of Bell Telephone Laboratories under the supervision of J. B. Bishop. Extensive data and oscillograms have been taken which indicate the extent of the impairment of transmitter performance under a variety of conditions, and the effectiveness of corrective methods which are to be described.

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I. Monitoring Methods

The first phase of the study necessarily involved determination of proper monitoring conditions whereby actual sideband power, effective percentage modulation, and true distortion in the signal delivered to the antenna system can be measured. A preliminary discussion of the monitoring problem appeared in a previous publication,[1] in which it was shown that the appearance of the modulation envelope is greatly different at different points in a coupling circuit or at different points along a transmission line when the impedance of the termination at the side frequencies differs substantially from the impedance at the carrier frequency. This is illustrated by the oscillograms of Fig. 1, which show the voltage envelope for a modulation frequency of 7500 cps as observed at three points along a transmission line or coupling network whose termination, for example, is equivalent to a series-tuned circuit resonant at the carrier frequency. At the termination, or at points removed therefrom by an even number of quarter wavelengths, a fully modulated voltage wave may appear (Fig. 1(a)), and the amplitude of each of the two side-frequency voltages accordingly will be one-half the amplitude of the carrier voltage. But, because the impedance rises on either side of the carrier frequency (Fig. 1(d)), the side-

[1] W. H. Doherty, "Notes on modulation of AM transmitters," *The Oscillator*, p. 22, no. 5; October, 1946.

frequency *currents* will each be less than one-half the carrier current, and an inspection of the current envelope would show substantially less than 100 per cent modulation. On the other hand, at points removed from the termination by odd quarter wavelengths, the impedance will correspond to that of a parallel-tuned circuit (Fig. 1(f)), decreasing on either side of the carrier frequency; and, although the current envelope if observed would show a fully modulated wave, the voltage

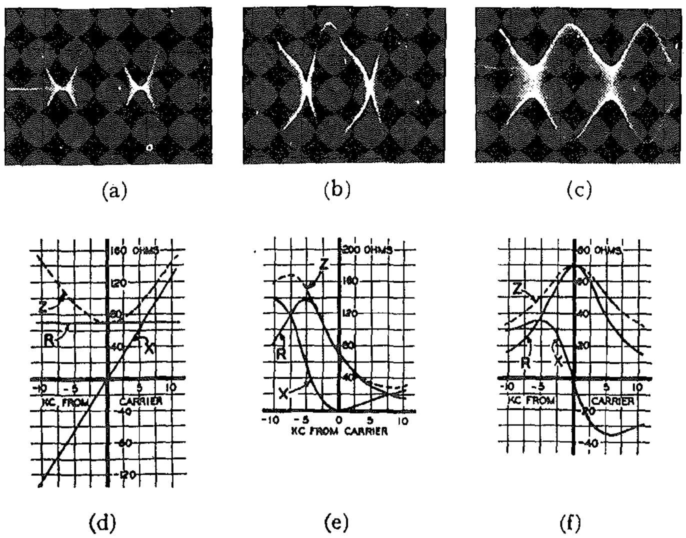


Fig. 1—Voltage envelopes and impedance versus frequency relations at three points in an output circuit or transmission line with narrow-band termination.

envelope (Fig. 1(c)) shows considerably less than 100 per cent modulation—in the case illustrated, only 60 per cent, since the impedance at the side frequencies is only 60 per cent of the impedance at the carrier frequency (the inverse of the situation of Fig. 1(d)). Finally, at odd eighth-wavelength points, where the impedance versus frequency curve is dissymmetrical (Fig. 1(e)), the voltage envelope has the distorted appearance indicated in Fig. 1(b). The current envelope at this point would also be badly distorted. A monitoring rectifier and conventional distortion-measuring instrument would register a high percentage of distortion for this wave, yet there are no extraneous side frequencies being radiated, the envelope distortion being entirely due to the inequality and phase dissymmetry of the two desired side-frequency voltages at this particular point in the line or coupling circuit. It is obvious that if, in addition, the operator were to raise the audio input level, endeavoring to bring about apparent full modulation at this point, much more severe distortion would be registered, because the wave at other points would then be over-modulated.

The impedance versus frequency relations indicated are those corresponding to a simple resonant circuit in which the ratio of reactive volt-amperes to watts at 550 kc, for example, is approximately fifty to one. The impedance curve of a simple resonant circuit appears on a Smith chart[2] as a circle. Fig. 2 shows impedance circles D, E, and F for points along a transmission line

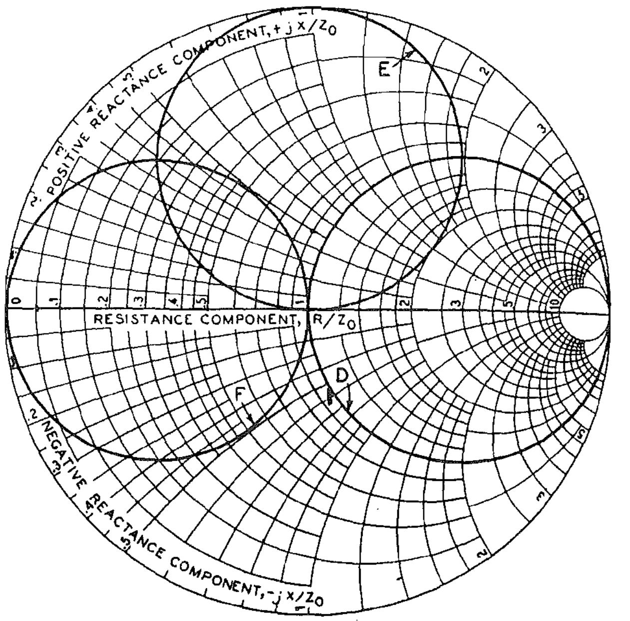


Fig. 2—Smith-chart impedance diagrams for a simple resonant circuit at points along a transmission line corresponding to D, E, and F, of Fig. 1.

or coupling circuit where the impedance versus frequency relations would correspond to D, E, and F of Fig. 1. The frequency dependence in the case shown is several times more severe than the RMA standard for artificial antennas for transmitter testing, but is comparable with that frequently found in actual broadcast installations.

Measurement of Effective Modulation and Distortion

If one wishes to determine actual delivered sideband power by measurement of a sample of the current envelope, it is necessary to make the inspection and measurement at a point in the circuit corresponding to Fig. 1(d), where the series resistance is independent of frequency over the band transmitted, since it is the current squared times the series resistance that determines power. We may refer to such points as D points for convenience. In order that the measurement may include true radiated distortion power, i.e., power in extraneous sidebands, the constancy should hold over a correspondingly wider band. Now, when measuring at such points, one should not try to adjust the audio input for 100 per cent modulation of the current envelope observed on the oscilloscope, but for a certain lower percentage—60 per cent, in the case considered in Fig. 1—since the voltage envelope (not being observed by the operator, but shown in Fig. 1(a)) will then have reached full modulation, and any further increase

[2] P. H. Smith, "An improved transmission line calculator," *Electronics*, vol. 17, pp. 130–134; January, 1944.

would necessarily entail severe distortion in both the voltage wave and the current wave.

If, on the other hand, it is the voltage envelope rather than the current envelope that is to be monitored, the point of measurement should be one where the *parallel* rather than the series resistance is constant over the transmitted band, since the power is the voltage squared divided by the parallel resistance. It can be shown[3] that this constancy will be found only at points corresponding to Fig. 1(f), one-quarter wavelength removed from D points. We may label these as F points. At these F points one must not look for a fully modulated wave, since the *current* wave (not being observed) will have reached full modulation, in the case considered, when the voltage wave, seen in Fig. 1(c), is only 60 per cent modulated.

Thus we have the curious situation that, with narrow-band antennas, and when modulating at high audio frequencies, only certain points D in the output circuit of the transmitter are suitable for determination of effective sideband power and distortion power when a sample of the current wave is being analyzed, and only certain other points F when a voltage sample is being analyzed; while the maximum permissible modulation of the wave being analyzed has to be set in the reverse manner, i.e., by inspection of the voltage envelope at points D or the current envelope at points F, or by calculation from the impedance versus frequency curve.

It should be noted that at D points, where the series resistance is independent of frequency, the parallel resistance is greater for the sidebands than for the carrier frequency, and a distortion measurement on the voltage envelope would be pessimistic since a given distortion power will be represented by a disproportionately high sideband voltage. Similarly, at F points, where the parallel resistance is independent of frequency but the series resistance is lower for the sidebands than for the carrier, a distortion measurement on the current envelope will be pessimistic. Thus the point where minimum distortion is registered is the correct monitoring point for the ideal case described, and will also, in general, afford the most reliable measure of distortion power in cases where the impedance diagram is irregular.

In order to permit the plotting of conventional curves of distortion versus modulation frequency for particular percentages of modulation, it is necessary to express the percentage modulation in terms of the quantity—current or voltage—which can be allowed to attain full modulation at the monitoring point, even though the true sideband power, with sharply tuned loads, does not correspond to full modulation. There is, moreover, justification for this in the fact that, when the maximum permissible modulation (without overmodulation) is reached, the power amplifier tubes are being required to deliver either full peak current or full peak voltage to the load, even if not both.

While it is only in certain types of programs that a broadcast transmitter is subjected to heavy modulation at high audio frequencies, the application of a test tone and measurement of harmonic distortion at such frequencies is a part of regular testing routine carried out with standard station equipment, and, when correctly done, provides valuable information both in the initial tune-up of a transmitter and in maintenance. However, an equally important application of the monitoring technique just described is in connection with the over-all frequency characteristic of the transmitter from audio input to sideband power output. This will be discussed in Section II. With the recent establishment by the Federal Communications Commission of a requirement for submission of performance data prior to renewal of licenses, it is important to be able to verify delivery of appropriate energy to the array from the transmitter for all modulation frequencies.

II. Operation of the Power Amplifier

With the impedance-versus-frequency characteristic varying widely from point to point in a coupling circuit, it is scarcely necessary to state that the characteristic as found at the particular point where the power-amplifier tubes are connected is of profound importance in the performance of the amplifier, regardless of the type of circuit or modulation method used. In particular, reverting to Fig. 1, if the impedance seen by the tubes were to vary with frequency in the manner given by Fig. 1(e), the tubes would *have* to impress on the circuit a voltage envelope resembling Fig. 1(b) in order that the voltage envelope at a proper voltage monitoring point (an F point) might be free of distortion. Indeed, since the impedance dissymmetry of Fig. 1(e) necessarily entails considerable phase modulation in the radio-frequency wave which does not show up in the oscillogram, it would be necessary for the tubes to introduce a corresponding phase modulation as well as to deliver a voltage envelope distorted in amplitude. Since one can ask only that a transmitter deliver to its load a voltage wave or a current wave free of phase modulation and having its envelope identical in shape to the audio input wave, it is apparent that, when the load is frequency-sensitive, only a point of impedance symmetry such as a D point or an F point is appropriate for making connection to the tubes. In the former case, when a high-frequency test tone is applied, the tubes will be asked to deliver full rated voltage and less than full rated current at the peak of the envelope; in the latter case, full rated current at less than full rated voltage. In either case, undistorted voltage and current envelopes are desired.

[3] The well-known impedance-inverting property of quarter-wave lines and their equivalent networks can be expressed in a form which shows, interestingly, that if one speaks of parallel components at one end and series components at the other end, the inversion holds for the resistances and also, *independently*, for the reactances; hence, the constant series resistance at D points necessarily means constant parallel resistance at points one-quarter wavelength therefrom, irrespective of reactance values.

Tests on several transmitters at powers from 1 to 10 kw, with very sharply tuned artificial antennas, have confirmed that when arrangements are made to connect the tubes to a *D* or *F* point the distortion at high modulation frequencies, when properly measured, differs very little from the distortion measured with a flat antenna, the slight increase observed being attributable to the effect of the sharp antenna on the width of the band over which negative feedback is effective. Fig. 3 shows the harmonic distortion in an experimental 10-kw broadcast transmitter at 95 per cent modulation with (1) a flat dummy antenna, and (2) a dummy antenna having a ratio of kva to kw of 25 to 1 at 550 kc, giving an impedance-versus-frequency characteristic about half as severe as that of Fig. 1. The tubes were connected at an *F* point and the distortion and per cent modulation were determined in the manner described. In contrast, with the tubes connected at a point of impedance dissymmetry and the distortion and per cent modulation improperly monitored, apparent distortions as high as 20 and 30 per cent were recorded. Fig. 4 gives an example of the kind of envelope shape that was observed under such conditions. A distortion-measuring instrument registered 21 per cent for this wave.

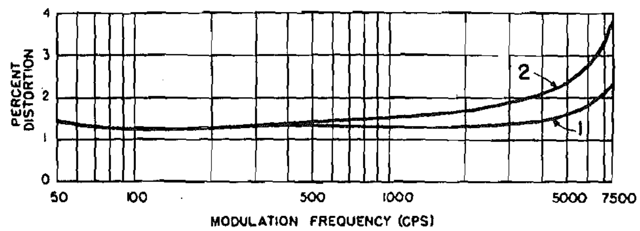


Fig. 3—Distortion curves for an experimental 10-kw transmitter at 95 per cent modulation with flat antenna (curve 1) and frequency-sensitive antenna (curve 2).

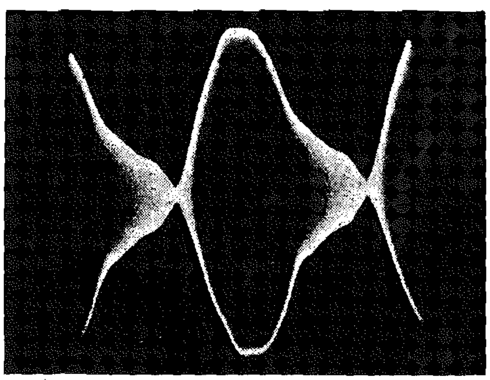


Fig. 4—Modulation envelope of a 10-kw transmitter for frequency-sensitive load as seen under improper operating and monitoring conditions.

The transmitter used for this test employed the high-efficiency circuit[4] devised by the author, with grid-bias modulation[5] of the final stage and employing "envelope" feedback, in which a sample of the radio-frequency output is detected and fed back to the audio input circuits.

From a distortion standpoint, the *D* and *F* connections are found to be about equally satisfactory. For transmitters employing envelope feedback, the *F* connection offers an important advantage in that it provides automatic compensation for the sideband-clipping tendency of the antenna. This comes about from the fact that with envelope feedback it is desirable for reasons of bandwidth to "pick off" a radio-frequency voltage sample directly at the plate of the final power tube; and if this point is an *F* point, the sample will then represent the voltage across a parallel resistance that is the same for the sidebands as for the carrier, and consequently the feedback will act to maintain a flat frequency characteristic in the radiated sideband power. When the impedance-versus-frequency characteristic of an antenna differs from that of a simple tuned circuit (i.e., exhibits in the transmitted band a curvature differing from that of the circles in Fig. 2), the corrective action of the feedback is less complete but still substantial. To achieve equivalent compensation for a narrow-band antenna characteristic by the use of high-kva coupling meshes in the output circuits would be unduly expensive in apparatus and would involve critical tuning and considerable radio-frequency power loss.

The potency of feedback derived from the radio-frequency envelope in bringing about delivery of the desired sideband energy to the antenna system is shown in Fig. 5, which pertains to the same 10-kw transmitter and the same types of loads to which the distortion curves of Fig. 3 apply. Curve 1 of Fig. 5 gives the overall frequency characteristic of the transmitter at 50 per cent modulation with no feedback when operating into a broad-band resistance load. The departures from flatness at the low and high ends arise mainly in the audio circuits in the transmitter. Curve 2, still without feedback, includes the additional loss at high modulation frequencies due to the frequency-sensitive load when the tubes are connected at an *F* point, the monitor being also connected at this point. But with envelope feedback applied, as derived from the voltage at this *F* point where the parallel resistance is constant over the band, the improved performance indicated by curve 3 is ob-

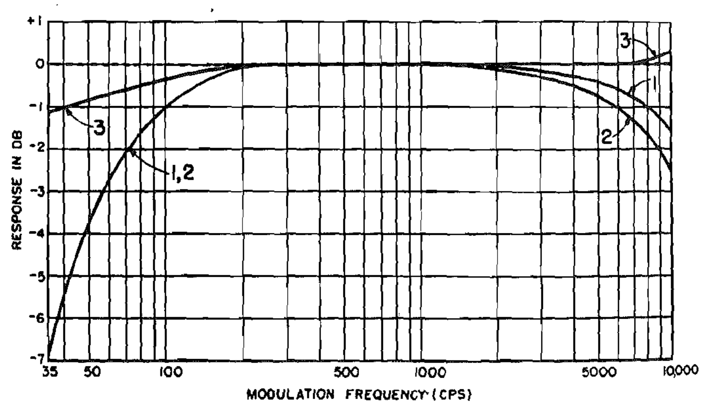


Fig. 5—Frequency characteristic curves for a 10-kw transmitter.

[4] U. S. Patent No. 2,210,028, W. H. Doherty.
[5] U. S. Patent No. 2,226,258, H. A. Reise and A. A. Skene.

tained. Fig. 6 shows, in contrast, corresponding curves for connection of the power amplifier tubes at a D point, where the parallel resistance is higher for the sidebands than for the carrier (but with the monitor still connected at an F point, since it is only here that a voltage-operated monitor will give a true indication of sideband power). The feedback in this case, being derived from

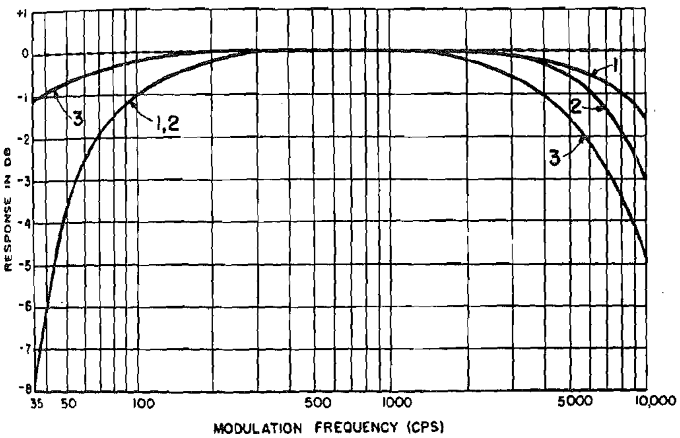
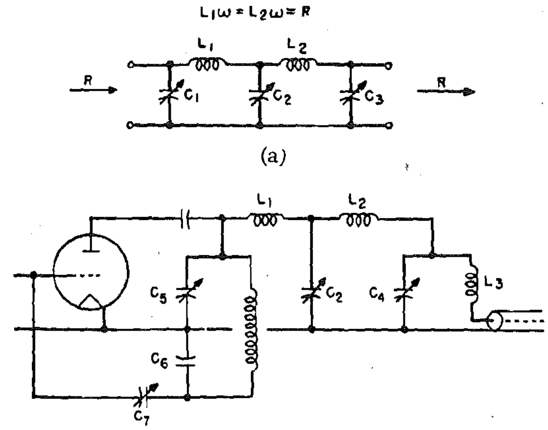


Fig. 6—Effect of unfavorable feedback connection on the frequency characteristic.

the voltage at the D point where the tubes are connected, actually aggravates the sideband-clipping action of the sharp antenna, as seen in curve 3 of Fig. 6. This is because, while the power tubes tend to impress higher than normal sideband voltages on the circuit on account of higher impedance to sideband frequencies, and thus partially compensate for the clipping action of the circuit, the feedback acts to prevent this compensation.

III. Transmitter Output-Circuit Design

To incorporate in broadcast transmitters the facilities for reorientation of the impedance-versus-frequency characteristic of any sharply tuned load that may be encountered, a variable phase shifter is required, equivalent to a "line stretcher," with a total range approaching 180 degrees to cover all cases. Such a phase shifter, as built for the experimental 10-kw transmitter on which these tests were conducted, is shown in Fig. 7(a). With the recent commercial availability of variable vacuum capacitors of wide capacitance range, linear calibration, and high voltage rating, it was most practical to incorporate this phasing device in the high-impedance output circuit of the transmitter prior to transforming down to the transmission line. The transformation ratio of the phase shifter shown is unity, and the coils L_1 and L_2 have reactances equal to the terminating resistance R regardless of the phase shift desired. For the minimum phase shift of 90 degrees, capacitor C_2 likewise has a reactance of R ohms, and C_1 and C_3 have zero capacitance. For greater phase shifts, capacitor C_2 is increased, and capacitors C_1 and C_3 come into play. By proper choice of these three capacitances, the phase shift can be increased to any value up to 270 degrees (or more) with no change required in the inductances, the input impedance remaining a pure resistance of R ohms through-



(b)

Fig. 7—Incorporation of a phase shifter in the transmitter output circuit.

out the range. The required admittances for these capacitances are:

$$C_1\omega = C_3\omega = \frac{1}{R}\,(1 - \cot \tfrac{1}{2}\Phi) \qquad (1)$$

$$C_2\omega = \frac{2}{R}\,(1 - \tfrac{1}{2}\sin \Phi) \qquad (2)$$

where Φ is the phase shift desired. These relations are shown in Fig. 8.

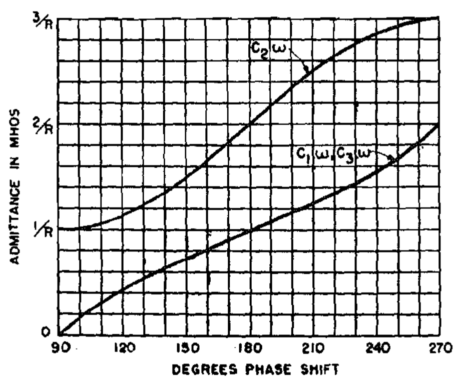


Fig. 8—Capacitive admittance required for the phase shifter of Fig. 7.

This phasing device is inserted, as shown in Fig. 7(b), between the radio-frequency plate terminal of the power amplifier and a load circuit consisting of C_4 and L_3, normally tuned to resonance and matching the amplifier to the transmission line. The built-out neutralizing circuit shown contains a "tank" capacitor C_5 with which the power tube is tuned in the conventional manner. With this arrangement, when a phase shift other than 90 degrees is required for improving performance with a frequency-sensitive antenna, the capacitances C_1 and C_3 of Fig. 7(a) are obtained by simply increasing the values of C_4 and C_5, which, like C_2, are variable vacuum capacitors.

In stations whose radiation patterns are different for day and night operation, the daytime pattern usually involves a simpler excitation of the antenna array, giving a flatter impedance characteristic. The phase-shifter adjustment chosen would, accordingly, be that best fitted to the nighttime impedance curve.

In the 10-kw transmitter built for testing these principles, the tube shown in Fig. 7(b) was the No. 2 or "peak" tube of a high-efficiency amplifier operating at a plate potential of 10,000 volts. The desired load impedance for the amplifier was 720 ohms. Coils L_1 and L_2 were accordingly made 720 ohms each and coil L_3 was adjusted for 186 ohms to obtain, in combination with C_4 a transformation from the 51.5-ohm coaxial-line impedance to 720 ohms. Capacitor C_2 is made 720 ohms in all cases where the antenna presents no bandwidth problem. When a narrow-band antenna is encountered, the required phase shift for best operation is determined by plotting the impedance characteristic at the input terminals of the transmission line on a Smith chart with an added peripheral scale, as shown in Fig. 9. Recalling

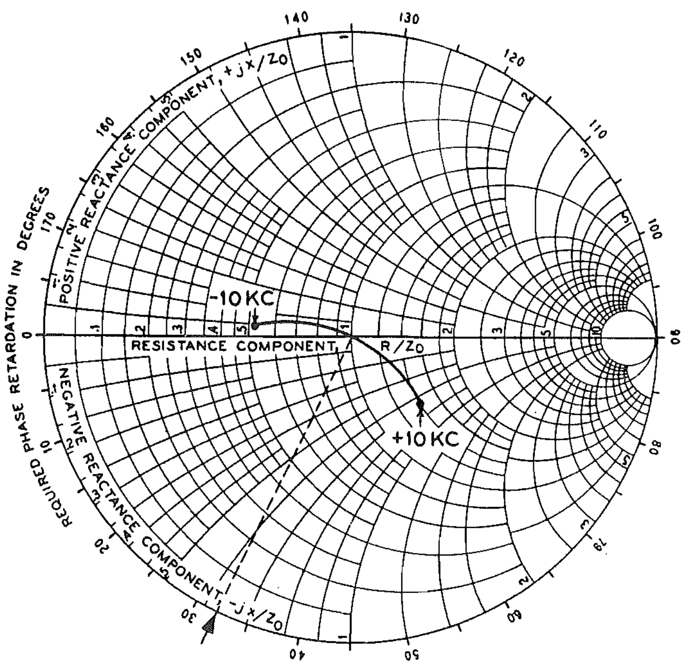


Fig. 9—Determination of the required phase shift in a typical case from the Smith chart.

that the desired orientation of the characteristic at the plate of the power-amplifier tube is that of circle F of Fig. 2, it is seen from Fig. 9 that the total phase retardation desired between the transmission line and the power tube for the case illustrated is either 32 degrees, or 180 plus 32 degrees. Since the phase retardation introduced by coil L_3 is $\tan^{-1} 186/51.5$ or 75 degrees, it is necessary to adjust the phase shifter to 137 degrees to obtain the total of 212 degrees required orientation. Fig. 8 then gives the values of C_2 and for the increments to be made in C_4 and C_5 to constitute effectively the capacitances C_1 and C_3 of Fig. 7(a). The actual final adjustment of C_5 is, of course, that which gives a unity power-factor load at the amplifier tube.

The radio-frequency plate terminal, being an F point, is used as the source of energy for the feedback rectifier and monitoring rectifier, due consideration being given to the fact that with a narrow-band antenna the voltage envelope observed when modulation is applied at high audio frequencies will not indicate 100 per cent modulation when the current delivered by the tubes is fully modulated.

With some antenna arrays the impedance may vary so irregularly with frequency as to call for a compromise adjustment which is difficult of prediction from the impedance diagram. In such cases, experimental determination of optimum phase shift is desirable by direct observation of the envelope shape at the plates of the tubes. The type of phase shifter described is especially well adapted to this procedure because of the wide range of adjustment possible without removal of power and the constancy of carrier-frequency impedance as the phase shift is varied.

Because of the extra harmonic suppression provided by the phase-shifting network, the usual harmonic filter connected at the input to the transmission line and employing mica capacitors (due to the low impedance) is no longer required. In cases where an unusually high degree of suppression is needed for one harmonic, a small fixed vacuum capacitor paralleling L_3 will provide a substantial increase in suppression. The circuit described thus combines with its property of handling frequency-sensitive loads the features of high harmonic suppression and long-life components.

In most cases, a station with a new antenna and transmitter would operate initially with the phase shifter set for its minimum shift of 90 degrees. After the completion of all antenna adjustments and the establishment of the final radiation pattern, the transmission-line input impedance would be measured over a wide band, and the desirability of phase correction determined. By merely increasing the capacitances of three variable capacitors, any additional phase shift desired can then be introduced to permit optimum performance of the transmitter and provide a monitoring point where the most reliable measurements of this performance can be made.

ACKNOWLEDGMENT

The experimental transmitters used in the investigations reported were built by H. A. Reise and C. W. Norwood of Bell Telephone Laboratories, who were also responsible for all of the performance data obtained. The contributions of these engineers and their colleagues are gratefully acknowledged.

Appendix D

Moulton, Clifford H., "Signal Distortion by Directional Broadcast Antennas," Proceedings of the I.R.E., May 1952, pgs. 595-600.

Signal Distortion by Directional Broadcast Antennas*

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Summary—Directional broadcast antenna systems are inherently capable of producing signal distortion. One type of distortion, described by Doherty,[1] involves the frequency sensitivity of the antenna input impedance and relationships between this impedance and the transmitter and transmission lines. A second type of distortion results from the directional radiation characteristics of the array, and is caused by differences in the response conditions of a directional array for the carrier and each sideband frequency.

Sources of Signal Distortion

THE WAVEFORM of the modulation envelope is responsible for the receiver audio-signal waveform in a double sideband AM system. The modulation envelope must therefore remain unchanged during transmission if signal distortion is to be avoided. Any change in the relative phase or amplitude of a signal component may result in a change in the waveform of the modulation envelope, which may cause audio distortion.

The directional pattern of an array is a function of the transmitted frequency, and is necessarily different at the sideband frequencies from that at the carrier frequency. These radiation-pattern differences cause changes in the amplitudes and relative phases of the signal components arriving at the receiving point and hence changes in the modulation envelope. The effect of altering the phases or amplitudes of the signal components in any particular manner may be determined by adding the components vectorially and obtaining the distorted modulation envelope.

Certain combinations of phase shifts and amplitudes result in large amounts of modulation-envelope distortion. One such condition occurs in directions where the high-frequency sideband amplitudes are increased with respect to the carrier amplitude. When the transmitter is modulated 100 per cent with a high-frequency tone, the sideband amplitudes are then greater than required for 100-per cent modulation of the carrier, producing a type of overmodulation.

At low audio modulating frequencies the response conditions for the carrier and sideband frequency components are essentially identical. With high audio modulating frequencies and relatively low carrier frequencies, however, the antenna bandwidth may be sufficiently low to result in severe changes in the relative phases and amplitudes of the signal components. In antenna-pattern null directions the strong carrier-frequency radiation fields of the antennas may almost completely cancel, but the cancellation may rapidly approach reinforcement for a sideband component as its frequency is removed from the carrier frequency. Carrier-frequency null directions are therefore most likely to be accompanied by increased high-frequency sideband power. In directions of maximum carrier power the converse conditions are likely to occur, resulting in reduced high-frequency sideband power. The audio-frequency response characteristics of a directional array will therefore be a function of the receiving direction.

It is significant that although audio distortion components may be found in the receiver output when modulation-envelope distortion exists the array itself does not introduce new frequency components in the transmitted signal but merely alters the amplitudes or phases of existing components. The receiver second detector is responsible for the addition of the large number of distortion components found in the receiver output.

Experimental Measurements

The simplest method of evaluating the audio distortion and frequency-response changes produced by directional broadcast arrays appears to be by direct measurement rather than by calculations or by the use of scale models. Two standard broadcast stations with directional antenna systems were therefore selected as test stations. One of these, station A, employed a two-tower 550-kc array with shunt feed. The other, station B, employed a three-tower 1,280-kc array with series feed.

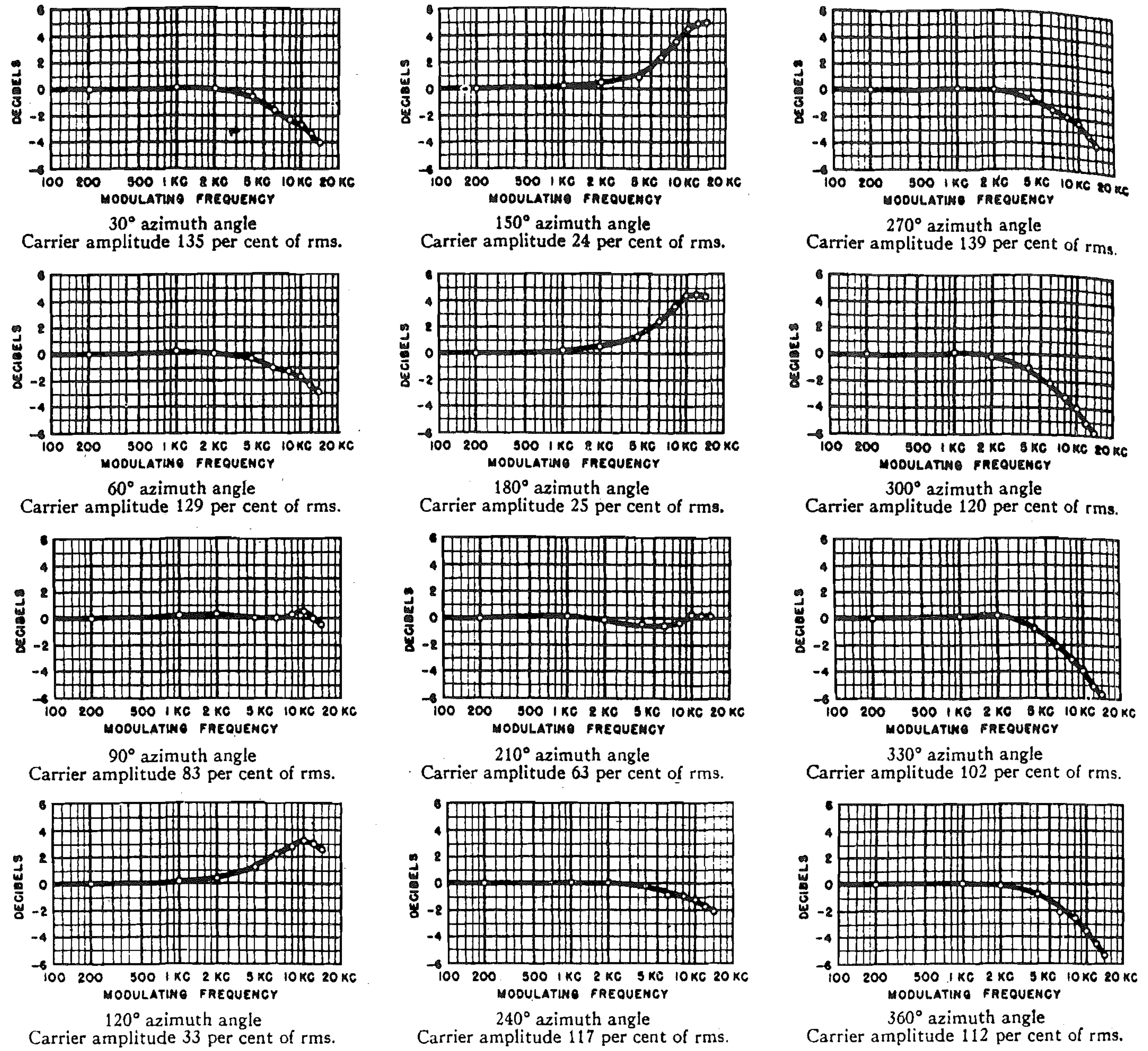
The signal distortion and frequency-response data which follow were obtained during the test period from 1 to 6 A.M. A battery amplifier with a whip antenna and ground rod was carried into completely open spaces. The received test signal was amplified at carrier frequency by broad-band amplifiers, and sent to a mobile unit by coaxial cable. After further amplification at the truck, the signal was then distributed to a Tektronix model 511-AD oscilloscope for photographing waveforms and to a Hewlett Packard model 330-B noise and distortion meter for distortion- and frequency-response measurements. Receiving locations were chosen outside of the antenna induction-field regions and at random distances from the antenna systems.

Frequency-response data are for 50-per cent modulation at the transmitter. This modulation percentage allows a few decibels of sideband power increase before the point of severe distortion, and yet prevents noise and interference from becoming too objectionable. Two-

* Decimal classification: R326.4×R148.11. Original manuscript received by the Institute, January 22, 1951; revised manuscript received, October 1, 1951.

† Tektronix, Inc., P. O. Box 831, Portland 7, Ore.

[1] W. H. Doherty, "Operation of AM broadcast transmitters into sharply tuned antenna systems," Proc. I.R.E., vol. 37, p. 729; July, 1949.



30° azimuth angle
Carrier amplitude 135 per cent of rms.

150° azimuth angle
Carrier amplitude 24 per cent of rms.

270° azimuth angle
Carrier amplitude 139 per cent of rms.

60° azimuth angle
Carrier amplitude 129 per cent of rms.

180° azimuth angle
Carrier amplitude 25 per cent of rms.

300° azimuth angle
Carrier amplitude 120 per cent of rms.

90° azimuth angle
Carrier amplitude 83 per cent of rms.

210° azimuth angle
Carrier amplitude 63 per cent of rms.

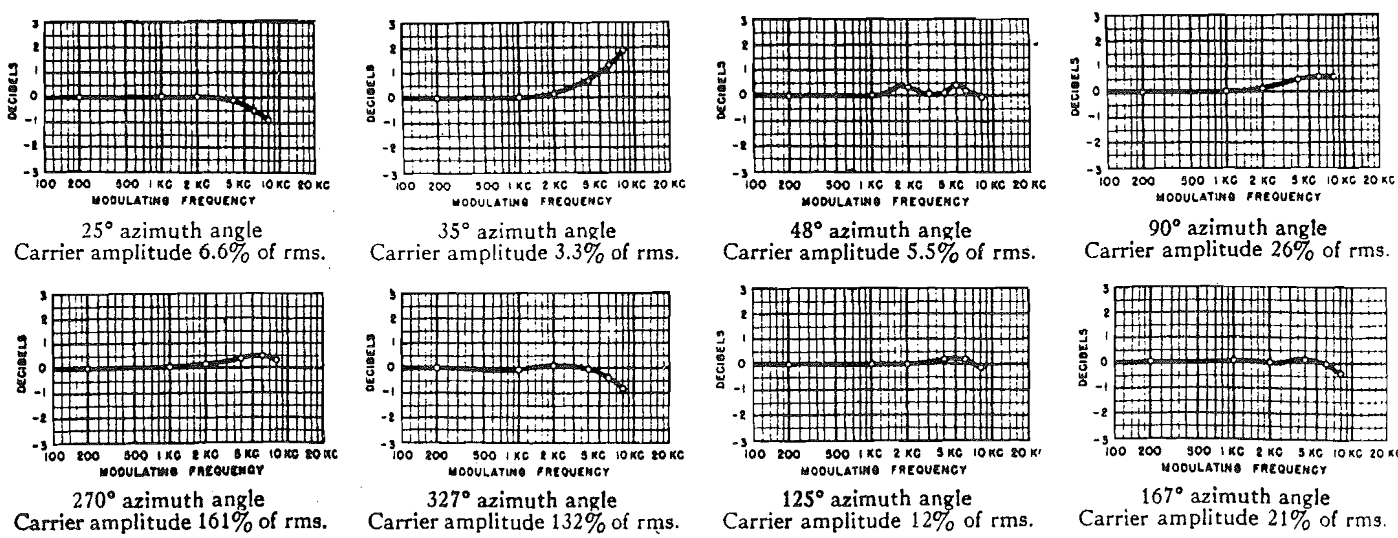
330° azimuth angle
Carrier amplitude 102 per cent of rms.

120° azimuth angle
Carrier amplitude 33 per cent of rms.

240° azimuth angle
Carrier amplitude 117 per cent of rms.

360° azimuth angle
Carrier amplitude 112 per cent of rms.

Fig. 1—Audio-frequency response curves for station A at various azimuth angles. Data are for 50-per cent modulation at the transmitter and random distances from the antenna system.



25° azimuth angle
Carrier amplitude 6.6% of rms.

35° azimuth angle
Carrier amplitude 3.3% of rms.

48° azimuth angle
Carrier amplitude 5.5% of rms.

90° azimuth angle
Carrier amplitude 26% of rms.

270° azimuth angle
Carrier amplitude 161% of rms.

327° azimuth angle
Carrier amplitude 132% of rms.

125° azimuth angle
Carrier amplitude 12% of rms.

167° azimuth angle
Carrier amplitude 21% of rms.

Fig. 2—Audio-frequency response curves for station B at various azimuth angles. Data are for 50-per cent modulation at the transmitter and random distances from the antenna system.

kc, square-wave audio modulation was employed for a series of waveform photographs at various receiving locations because the phases and amplitudes of a large number of audio-frequency components could be observed simultaneously. Test data at station B were limited to 8 kc because of modulator overload.

Results

Audio-frequency response curves at various azimuth angles are shown for station A and station B in Figs. 1 and 2, respectively. The high-frequency audio response is greater in the null directions than in the maximum directions in both cases. The zero decibel level for each direction is taken as the received audio level at 200 cycles for 50-per cent modulation at the transmitter output. The differences in the response curves at station B are much less than at station A, partly because station B operates at a carrier frequency 2.33 times as high as station A. The audio spectrum represents a much smaller percentage of the carrier frequency at 1280 kc than at 550 kc. It is necessary to correct the data of station B to 550-kc conditions, in order to compare them directly with the data of station A, by dividing the modulating frequencies of station B by 2.33. Tests at station B extend to 8 kc, which becomes 3.44 kc when corrected.

The maximum spread in response curves, Δ, expressed in decibels, is plotted as a function of audio modulating frequency for both stations in Fig. 3. Another curve of corrected Δ for station B is given in Fig. 3 after correcting the audio frequencies to 550 kc. From the latter curve it is seen that Δ would be higher for station B than

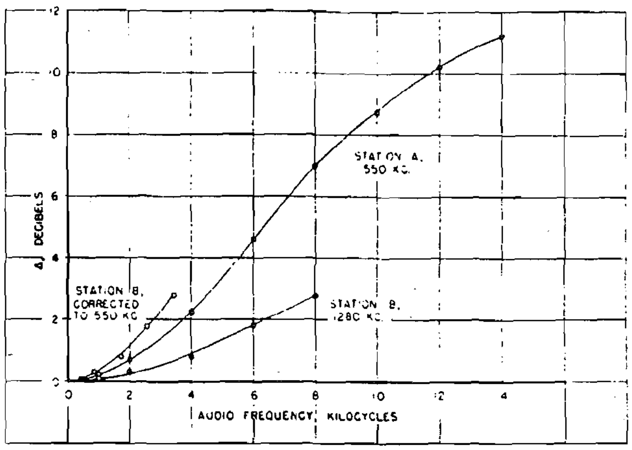
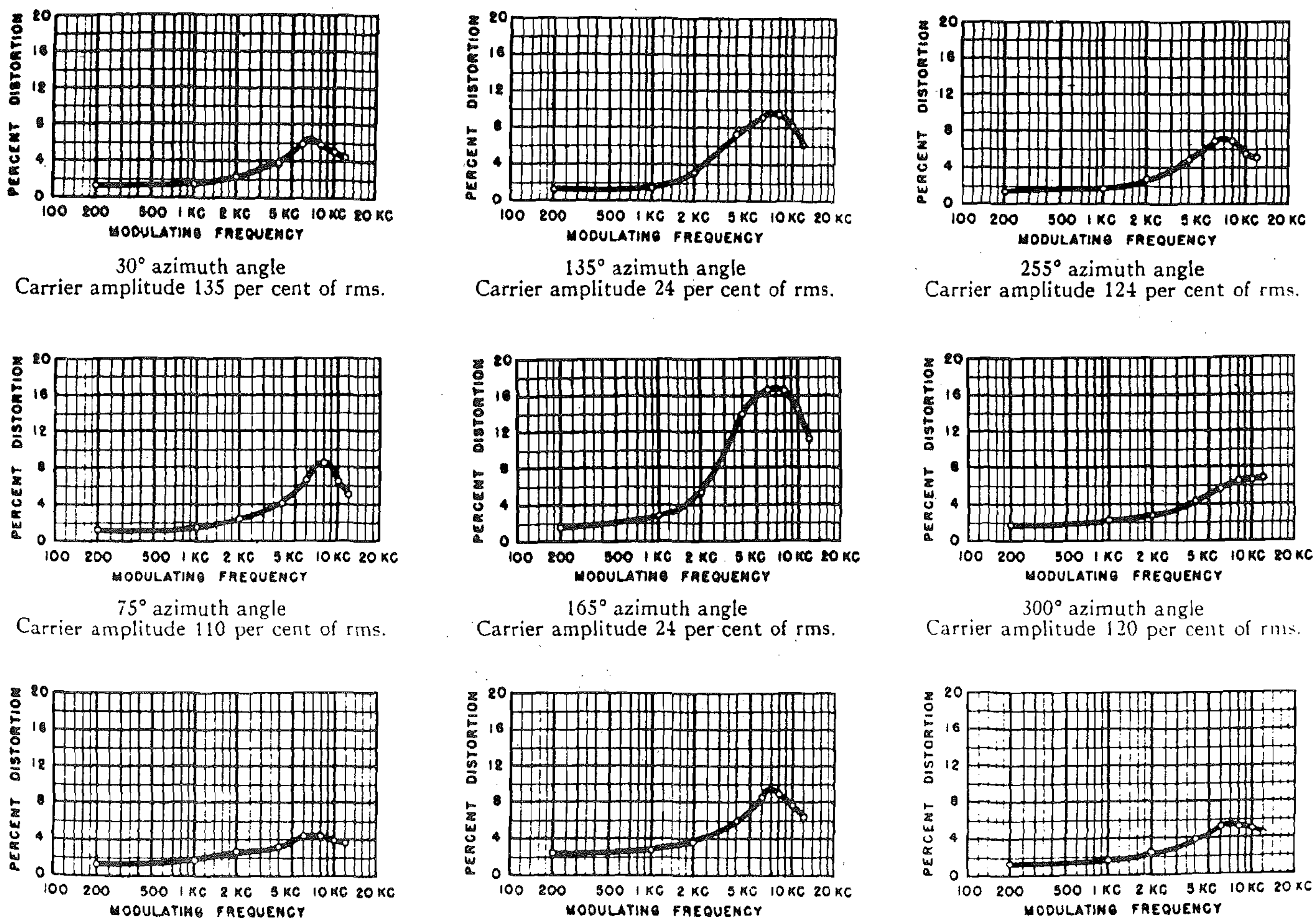


Fig. 3—Maximum response-curve differences, Δ, versus modulating frequency for 50-per cent modulation at the test stations.



30° azimuth angle
Carrier amplitude 135 per cent of rms.

135° azimuth angle
Carrier amplitude 24 per cent of rms.

255° azimuth angle
Carrier amplitude 124 per cent of rms.

75° azimuth angle
Carrier amplitude 110 per cent of rms.

165° azimuth angle
Carrier amplitude 24 per cent of rms.

300° azimuth angle
Carrier amplitude 120 per cent of rms.

120° azimuth angle
Carrier amplitude 33 per cent of rms.

210° azimuth angle
Carrier amplitude 63 per cent of rms.

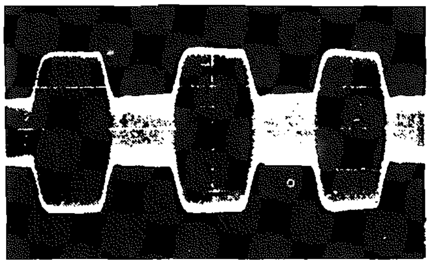
345° azimuth angle
Carrier amplitude 104 per cent of rms.

Fig. 4—Station A audio-frequency harmonic distortion as a function of modulating frequency at various azimuth angles. The transmitter was modulated 50 per cent.

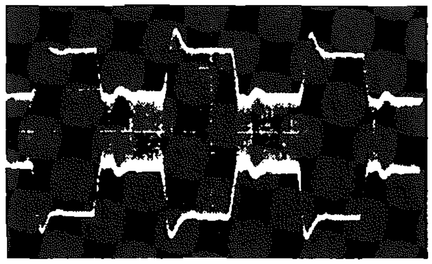
station A if both were to be scaled to operate at 550 kc.

Harmonic distortion of the received signal with 50-per cent sinusoidal modulation at the transmitter output is shown in Fig. 4 as a function of modulating frequency and azimuth angle at station A. The highest distortion percentages are found in the null directions. Similar data could not be taken at station B because of interference and noise in the null directions.

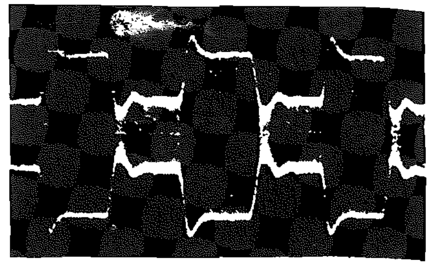
The responses of both stations to square-wave audio modulation are seen in Figs. 5 and 6. The waveforms of station A are markedly different in the null and maximum directions. The slow rises of the waveforms in the maximum directions show reduced high-frequency sideband power while the leading- and trailing-edge overshoots in the null directions show increased high-frequency sidebands. The peculiar waveform in the leading



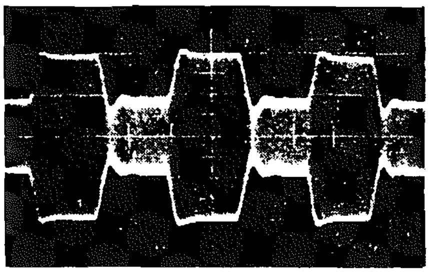
45° azimuth angle
Carrier amplitude 139 per cent of rms.



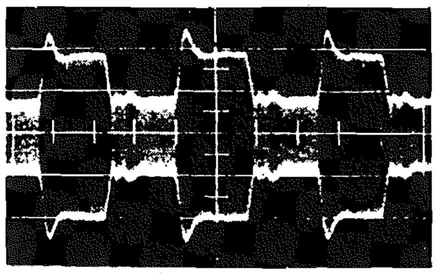
135° azimuth angle
Carrier amplitude 24 per cent of rms.



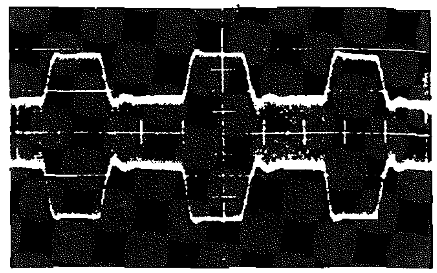
180° azimuth angle
Carrier amplitude 25 per cent of rms.



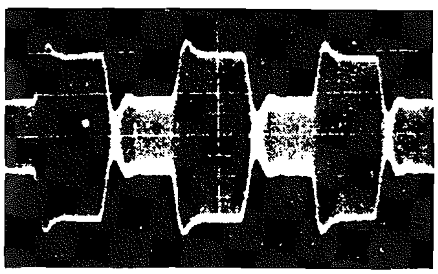
90° azimuth angle
Carrier amplitude 83 per cent of rms.



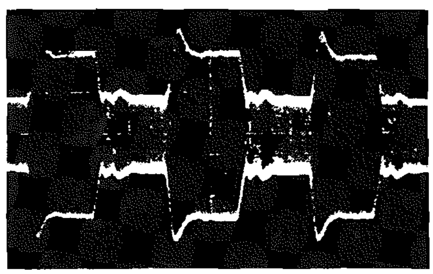
150° azimuth angle
Carrier amplitude 24 per cent of rms.



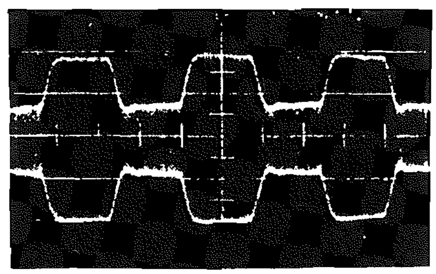
210° azimuth angle
Carrier amplitude 63 per cent of rms.



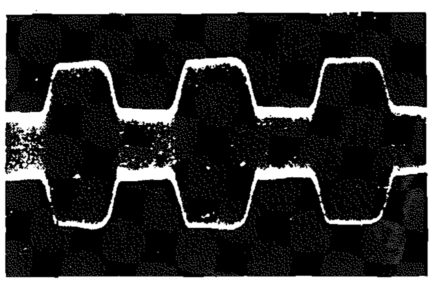
120° azimuth angle
Carrier amplitude 33 per cent of rms.



165° azimuth angle
Carrier amplitude 24 per cent of rms.

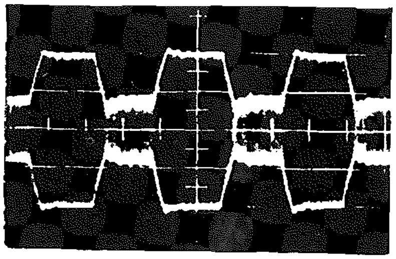


270° azimuth angle
Carrier amplitude 139 per cent of rms

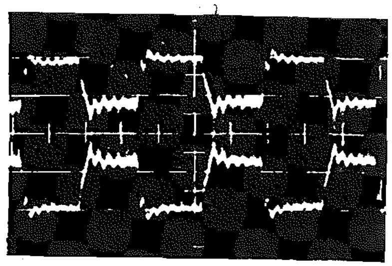


345° azimuth angle
Carrier amplitude 104 per cent of rms.

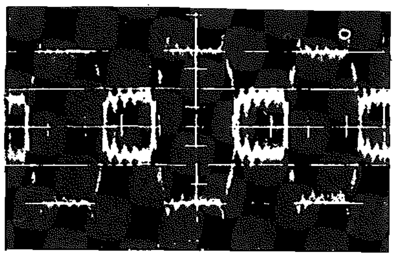
Fig. 5—Photographs of station A modulation-envelope oscilloscope traces at various azimuth angles for 2-kc square-wave modulation. The transmitter was modulated 50 per cent.



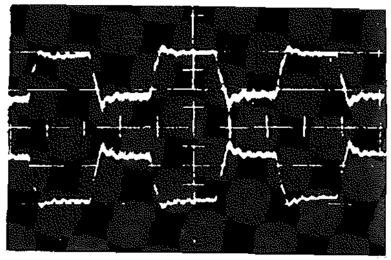
25° azimuth angle
Carrier amplitude 6.6 per cent of rms.



48° azimuth angle
Carrier amplitude 5.5 per cent of rms.

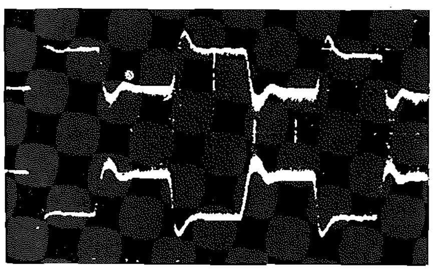


35° azimuth angle
Carrier amplitude 3.3 per cent of rms.

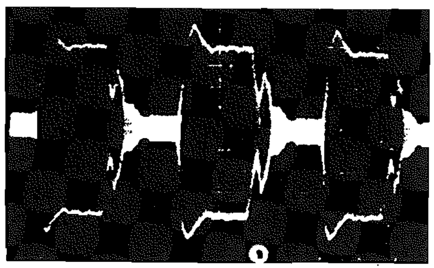


270° azimuth angle
Carrier amplitude 161 per cent of rms.

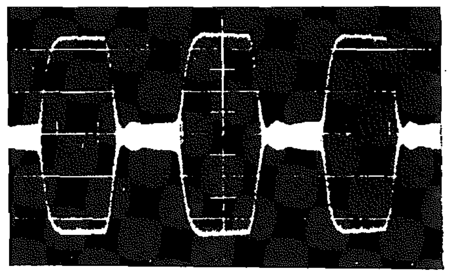
Fig. 6—Station B square-wave modulation photographs for various azimuth angles. The transmitter was modulated 50 per cent with a 2-kc square wave.



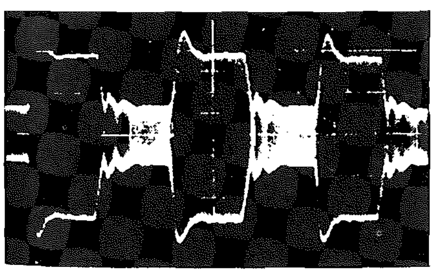
(a)
Modulation 40 per cent.



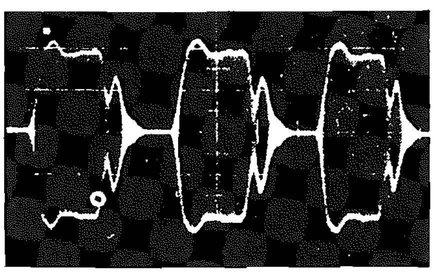
(c)
Modulation 85 per cent.



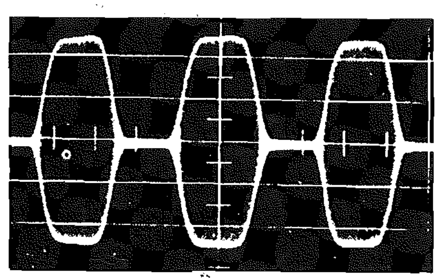
(e)
Modulation 85 per cent.



(b)
Modulation 60 per cent.



(d)
Modulation 98 per cent.



(f)
Modulation 95 per cent.

Fig. 7(a–d)—Station A square-wave response for various modulation percentages at an azimuth angle of 135° (carrier amplitude 23.8 per cent of rms).
Fig. 7(e and f)—Same at an azimuth angle of 255° (carrier amplitude 133.5 per cent of rms).

edge of the downward modulation half of the square wave in the null region of station A should be noted. The damped oscillation on each waveform of station B in Fig. 6 was present with a dummy antenna and was produced in the transmitter audio system. Noise and interference are responsible for the fuzziness of the null direction waveforms.

Fig. 7 shows the effect of modulation percentage on the transient response characteristics of station A for two receiving directions. A given percentage of modula-

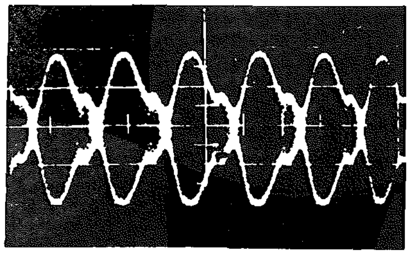
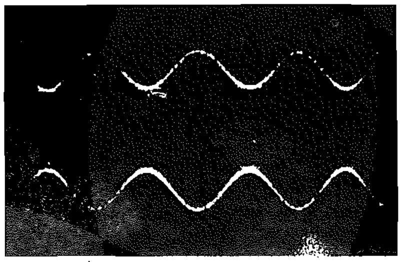
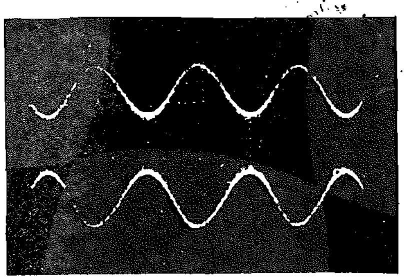


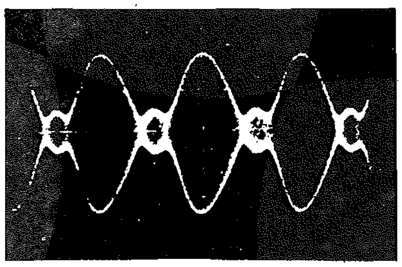
Fig. 8—Modulation envelope waveform at station A for 70-per cent modulation at 4 kc at an azimuth angle of 165° (carrier amplitude 23.8 per cent of rms).



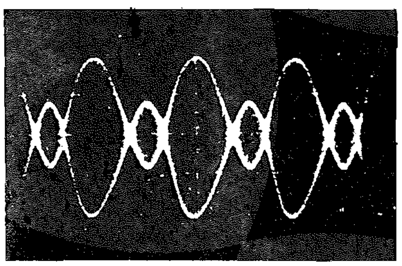
Transmitter modulation 20 per cent.



Transmitter modulation 40 per cent.



Transmitter modulation 60 per cent.



Transmitter modulation 80 per cent.

Fig. 9—Station A modulation-envelope photographs for 10-kc sinusoidal modulation at various modulation percentages. The receiving location azimuth angle was 135° (carrier amplitude 23.8 per cent of rms).

tion results in entirely different modulation envelopes in the two receiving directions. High percentages of modulation are accompanied by severe distortion in the null region.

An example of serious distortion in the null direction at 4 kc is shown in Fig. 8. Analysis of this waveform shows that the two sidebands are not of the same amplitude and have been shifted slightly in time phase with respect to the carrier. Waveforms of the type shown in Fig. 9 occur when the sidebands have similar amplitudes and are shifted from the normal phase position. The effect of modulation percentage on the modulation envelope of station A in the null direction with 10-kc sinusoidal modulation is seen from Fig. 9.

Discussion

Arrays with many elements, nigh-Q tuning networks, negative power elements, and deep nulls are more likely to have severe antenna distortion than simpler, lower Q systems. Stations operating at the low-frequency end of the broadcast band are much more subject to this distortion than those operating at the high-frequency end of the band.

Deep nulls should be avoided from a standpoint of signal distortion if service is to be rendered in the null directions.

Directional signal distortion in the horizontal plane would not be expected for single-element vertical antennas.

While this article has dealt principally with transmitting antenna signal distortion at broadcast frequencies, it is apparent that somewhat similar effects should also be expected for receiving antennas, for other radio frequencies, and for other types of modulation.

Conclusions

1. Signal distortion is observed in directional broadcast antennas, and is found to be a function of receiving direction.
2. Signal distortion results from changes produced by the directional antenna system in the magnitudes or relative phases of the signal components.
3. Directional signal distortion is accentuated by deep nulls, low-percentage antenna bandwidth, high audio-modulating frequency, and high percentage of modulation.

Acknowledgment

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Appendix E

Bingeman, Grant W. and C.V. Clarke, "AM Antenna System Bandwidth versus Harmonic Distortion," IEEE Transactions on Broadcasting, June 1977, pgs. 50-55.

AM ANTENNA SYSTEM BANDWIDTH VERSUS HARMONIC DISTORTION

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With the advent of AM stereo, the growing concern for high-fidelity, and the race for loudness, bandwidth has become a popular topic.[1] We all know that the state-of-the-art transmitter has solved the transmitter bandwidth problem, and that the weakest link in the chain is the antenna system, now more than ever. This paper will show that most AM antenna systems can be designed and adjusted to have a bandwidth compatible with today's transmitters.

There are two critical monitoring points in the AM RF chain that are affected by antenna system bandwidth: the final stage of the transmitter, and the far-field detector. Since the need for an optimized bandwidth at the transmitter has been recognized for several decades,[2] this paper emphasizes antenna system bandwidth as it relates to distortion at the far-field detector.

This analysis includes several assumptions. These are:

1. The receiving point is located in the major lobe of the pattern, and in the far-field;
2. The detector responds to the peak of the envelope only;
3. Antenna radiation efficiency is constant over the band of interest;
4. Antenna current distribution is sinusoidal;
5. The transmitter is ideal and is a separate, distortionless voltage generator for each frequency

   $(f_o,\ f_o - f_m,\ f_o + f_m)$;

6. The transmission lines are lossless;
7. The transmitter is being modulated by a 10 kHz tone;
8. The center frequency (carrier) is 750 kHz.

In order to meaningfully analyze the data, a computer program was developed by C. V. Clarke (Senior Engineer, Harris Broadcast Products Division), which quantitatively relates bandwidth to total harmonic distortion (THD) and modulation index. This program uses a numerical Fourier series approach[3] to calculate the amplitude of the first five harmonics produced at the detector due to an imperfect sideband pair.

The approach to data collection was to model several variations of a representative phasor on the computer (the results of this analysis also apply to non-directional systems). The complex sideband currents in each radiator were determined, and then combined at the receiving point (P, fig. 1). From this superposition the r.m.s. THD and the degrees of positive and negative modulation at the far-field detector were determined for a given modulation index at the transmitter. The idea was to minimize the difference between the modulation index at the transmitter and the modulation index in the far-field, and to minimize the harmonic distortion in the far-field. The modulation index and THD seen at the transmitter will rarely be the same as those seen in the far-field.

The following parameters (and all the others they imply) were modeled as being frequency dependent:

1. All the reactances;
2. Radiator self impedances;
3. Radiator mutual impedance;
4. Transmission line electrical length;
5. Far-field angular distance, Ø.

The radiators were modeled as a tee network (fig. 2) in order to eliminate the problem caused by interdependence of radiator currents and base impedance (eqn. 1 and 2).

$$Z_1 = Z_{11} + Z_{12}I_2/I_1 \qquad \text{eqn. 1}$$

$$Z_2 = Z_{22} + Z_{21}I_1/I_2 \qquad \text{eqn. 2}$$

For the phasor, a typical two-tower (fig. 3) array producing a cardioid pattern (fig. 4), using steel-guyed, uniform cross-section, series-fed towers of 90 degrees electrical height, was chosen.[4] The spacing and field ratios and angles were chosen so as to produce good final base impedances, within the constraints of the pattern. The tower self impedances were taken from empirical data. It was assumed that any impedance-transforming antenna accessories (e.g.: static drain chokes) or stray reactances within the system had already been taken into account, that all components were lossless, and that there was no mutual coupling between coils (even in the power divider).

An input tee network to match the transmitter to the power divider was chosen, as opposed to an input 'L' network, in order to allow optimization of the sideband response seen by the transmitter. It should be noted, however, that the Smith chart model of this 'line stretcher' is not exact if all the sideband impedance points are lumped together on the same chart. This is because the lower sideband wavelength we are dealing with is significantly longer than the upper sideband wavelength, but the angle scale around the circumference of the Smith chart is valid at one frequency only. More importantly, a tee network model of a transmission line is valid at only one frequency. Therefore, computer iteration was used to obtain exact and conclusive results.

It was determined that minimum far-field distortion is obtained when the sideband impedance, as seen by the transmitter, approaches symmetry; that is, when the sideband reactances are equal and opposite, and when the sideband resistances are equal.

This was not a particularly surprising result, and tends to support the validity of the analysis up to this point.[5] As a matter of fact, it looks like there is no reason why a well-designed phasor cannot be adjusted to introduce much less harmonic distortion (for a given, single tone) than the rest of the system building-blocks.

But we have yet to deal with far-field modulation index (defined in fig. 5).

The following are the original sideband impedances at the input to the phasor, after the system has been adjusted to produce the desired pattern:

$$Z\ (f_0 - 10\ \text{kHz}) = 57.1 + j13.4$$

$$Z\ (f_0) = 50.0 + j0$$

$$Z\ (f_0 + 10\ \text{kHz}) = 36.2 + j2.77$$

This is a fairly typical sideband pair. The phase shift across the input matching network is -90 degrees.

Figures 6 and 7 show what happens to the envelopes when we try for 100% negative modulation at the transmitter. The far-field envelope exists along the locus of points equidistant from the two towers. In the absence of any non-linearities, distortion occurs because of the unequal propagation characteristics across the frequency band of interest. Since the transmitter is an ideal voltage generator, a current pick-up was used for the envelope in Figure 6. A voltage pick-up would not indicate any distortion at this point in the circuit, except during negative overmodulation.

Figure 8 displays the harmonic distortion at the two monitoring points as a function of modulation index seen at the transmitter point. It is considerably worse for the listening audience in the far-field, and this is by no means the worst listening point in the pattern for the present unsymmetrical sideband situation.[6]

Figure 9 compares the modulation indexes at the transmitter with those in the far-field. There is almost a one-to-one relationship between the positive indexes, but the negative index relationship is markedly non-linear. The reference line is for the case of a perfectly flat, resistive load across the band of interest.

After iterating a while by altering the phase shift across the input matching network, a relatively symmetrical, realizable sideband pair at the transmitter for a phase shift of -155 degrees was obtained:

$$Z\ (f_0 - 10\ \text{kHz}) = 58.6 - j25.0$$

$$Z\ (f_0) = 50.0 + j0$$

$$Z\ (f_0 + 10\ \text{kHz}) = 56.6 + j20.1$$

Incidentally, when the phase shift across the input tee is changed like this, the radiator sideband voltages and currents change, but the radiator sideband impedances remain fixed, since we are dealing with a linear system.

With 100% modulation at the transmitter (both positive and negative), the THD becomes a mere 0.5%. The corresponding far-field distortion is only 0.1%. The reason for the smaller far-field THD is the fact that the far-field modulation index is only 66%, for a transmitter index of 100%. The envelopes are shown in Figures 10 and 11.

Figure 12 displays the modulation index relationship for the improved sideband pair. As you can see, the relationship is linear, but the far-field loudness seems to have dropped. Distortion is way down, but so is the far-field modulation index. This could get someone in a bit of trouble when the next rating period rolls around, but fortunately there is a solution.

There is another point of sideband symmetry available when the input matching network is adjusted to have a phase shift near -55 degrees. When this is done, the sideband impedances are as follows:

$$Z\ (f_0 - 10\ \text{kHz}) = 41.9 + j9.43$$

$$Z\ (f_0) = 50.0 + j.0$$

$$Z\ (f_0 + 10\ \text{kHz}) = 40.8 - j11.2$$

Figure 13 displays the new modulation index comparison; still linear, but now the far-field index is greater than its transmitter counterpart. Figure 14 shows the transmitter envelope when 100% modulation exists in the far-field. This new situation requires that the transmitter modulation index be limited to less than 85%, in order not to overmodulate in the far-field with a 10 kHz tone. One might conclude from all this that a real-world transmitter can now operate in a region where its inherent distortion is lower, its power consumption is lower, its tube life is extended, and component reliability is raised. Real transmitters, however, are designed to work best into a flat resistive load. As antenna system bandwidth is increased, the impedances of the balanced sideband pair will approach the center frequency impedance.

As the frequency of the modulating tone is decreased, line A in Figure 13 will move closer to the reference line, in most cases. For example, a 5 kHz tone will require a modulation index greater than 85% to produce 100% modulation at the far-field point. Since normal programming energy is concentrated closer to the center frequency than 10 kHz, the deleterious effects of bandwidth will not be as pronounced with average program material. But, of course, the FCC proof of performance is not done with average program material.

If we move the far-field monitoring point a full 90 degrees of azimuth to point P_2 (fig. 4), we find that the modulation index is about five percent higher, and the THD is about the same (compared with point P). Similar results are obtained for a 45 degree change in azimuth, at point P_3 (fig. 4). It is fortunate that this array produces only a five percent variation across the major listening area; if the variation were large, there would be little to gain from optimization of the sideband pair back at the transmitter.

As for the rest of the phasor, we can decrease the 'Q' (Q = X/R) of the common-point center frequency impedance by increasing the power divider inductance. Some people say that this will help the system bandwidth. But first, let's look at the sideband impedances at the input to the power divider, for our second design (input tee -155 degrees):

$$Z\ (f_0 - 10\ \text{kHz}) = 22.7 + j4.28$$

$$Z\ (f_0) = 26.9 + j9.12$$

$$Z\ (f_0 + 10\ \text{kHz}) = 36.2 + j6.62$$

This is about what would appear at the input to a 3/8 wavelength transmission line terminated in a series resonant circuit.

But when we increase the power divider inductance in an attempt to reduce 'Q', we also increase the impedance looking into, and the lagging phase shift across, the power divider. Therefore, we must change the phase shift of the phase-adjusting network (in this case, resulting in a reduction in component reactances), and adjust the input matching network (an increase in component reactances). These necessary additional changes may actually negate any improvement obtained at the power divider, so it is important to treat each phasor individually.

We must keep in mind that a bandwidth 'bottleneck' may exist elsewhere in the phasor. If we treat anything other than this bottleneck first, there may not be an appreciable improvement in the overall system bandwidth.

There are several potential bottlenecks in a phasor:

1. The input matching network,
2. Power divider,
3. Phase-adjusting network(s),
4. Transmission lines,
5. Antenna coupling networks,
6. And the radiators.

The networks (1,2,3,5) can be optimized by keeping the phase shifts across them as low as practical (there are trade-offs with stability[7] and ease of adjustment), and by using the least number of components possible, including zero. A transmission line can be modeled as a series of tee networks (one for each frequency), and thus fits in with Item 3. The radiator bandwidth is a bit more complicated, being a function of effective tower diameter, electrical height, spacing, system base currents, the ground system, and nearby structures. In most cases, the radiators are the bottleneck.

The approach to optimization of AM antenna system bandwidth is to first select the radiator configuration that provides the most cost-effective bandwidth for a given pattern requirement, using the tee-network model (fig. 2). Once this is done, it is a simple matter to optimize the bandwidth of the rest of the system, if one is methodical in developing the specific computer model, which should also include the exact output network of the transmitter.

In conclusion, we have found that the far-field harmonic distortion is minimized when the THD at the final stage of the transmitter is minimized (when sideband symmetry is approached), and that it is not necessary to get especially close to symmetry to obtain excellent distortion figures. More importantly, we have found that an improvement in far-field distortion may be accompanied by a considerable decrease in the far-field modulation index at all levels of modulation (for a single tone). We have also been reminded that a single Smith chart is not an exact model for the behavior of the sideband pair in a lumped-parameter system, and that an iteration technique can overcome this problem. It has been pointed out that sideband symmetry is not a panacea for real transmitters; thus an attempt to improve the system bandwidth should also be made. Finally, we have seen that it is possible to have our cake and eat it, too: vanishing distortion, perhaps a decreased workload for the transmitter, and loudness can all be obtained by following the methods outlined in this paper.

FOOTNOTES

1. Since this paper deals with harmonic distortion only, the reader interested in phase and intermodulation distortion is referred to:

   Stanford Goldman, Frequency Analysis, Modulation and Noise, Chapter entitled "Modulation", 1948, McGraw-Hill.
2. W. H. Doherty, "Operation of AM Broadcast Transmitters Into Sharply Tuned Antenna Systems", July, 1949, Proceedings of the I.R.E.
3. Skilling's Electrical Engineering Circuits, 1952, John Wiley and Sons, NY, Pages 439 - 455.
4. Preliminary design for a 50 kW phasor for Dr. Ivo Facca, Alvorada, Brazil.
5. Bill McCarren, Associate Director of AM Transmission Systems, CBS Radio, Paper presented at 1976 NAB Show in Chicago, "Antenna Q vs. Audio Response".
6. Clifford H. Moulton, "Signal Distortion by Directional Broadcast Antennas", May, 1952, Proceedings of the I.R.E.
7. R. S. Bush, "Phasing System Network Sensitivities", Available from Harris Broadcast Products Division, Quincy, Illinois.

FIG. 1. FAR-FIELD CURRENT SUPERPOSITION

$I_p = I_1 e^{j\phi_1} + I_2 e^{j\phi_2}$

$I_i = I_i e^{j\theta}$ = tower$_i$ base current

ϕ_i = angular distance

lines ϕ_1 and ϕ_2 are parallel, for P in the far-field

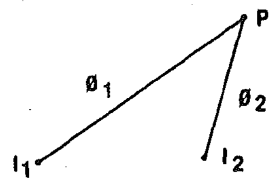
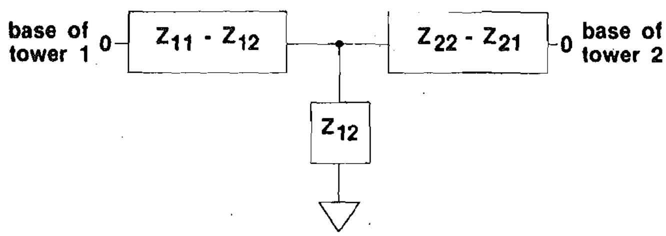


FIG. 2. MODEL OF RADIATORS



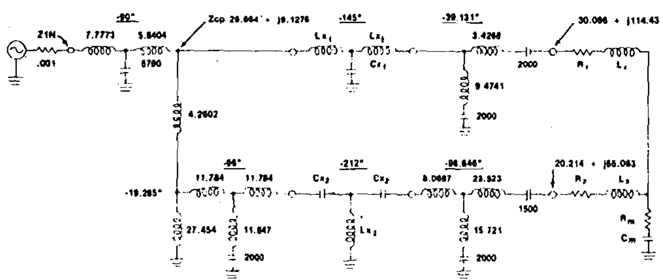
Z_{11} = self-impedance of radiator 1

Z_{22} = self-impedance of radiator 2

$Z_{12} = Z_{21}$ = mutual impedance between radiators 1 and 2

For this analysis, $Z_{11} = Z_{22}$.

FIG. 3. PHASOR R.F. DRAWING



$L_1 = L_2$

$R_1 = R_2$

$R_1 + jWL_1 = Z_{11} - Z_{12}$

$R_2 + jWL_2 = Z_{22} - Z_{21}$

COIL UNITS — uH

CAP UNITS — pF

FIG. 4. RADIATION PATTERN

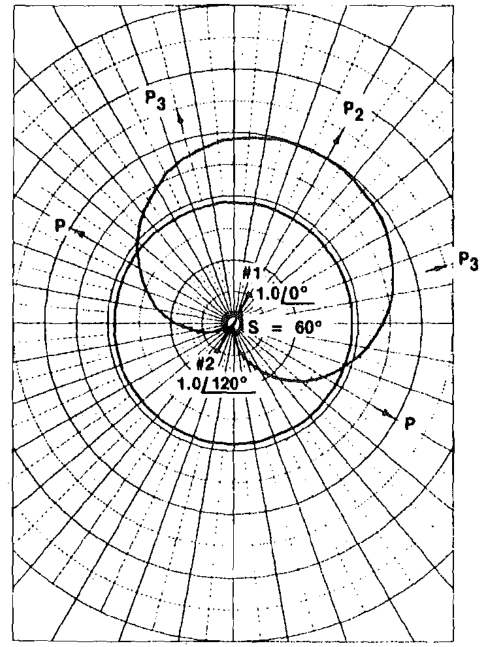
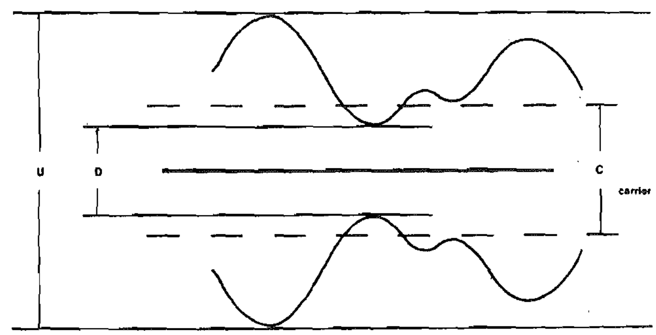


FIG. 5. MODULATION INDEX DEFINED



positive modulation = (U - C)/C
negative modulation = (C - D)/C

FIG. 6. ENVELOPE AT TRANSMITTER - DESIGN #1

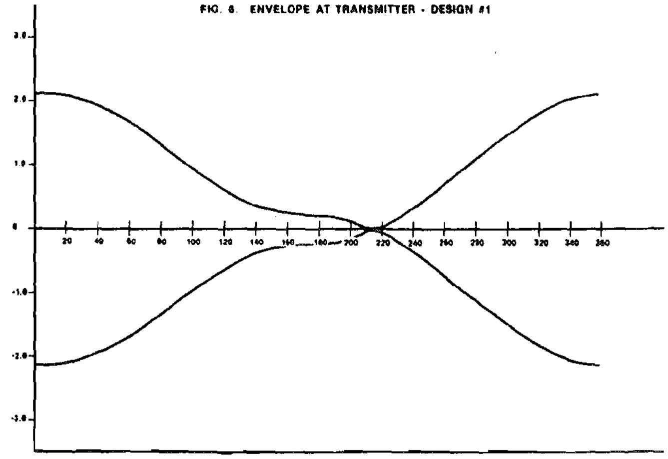


FIG. 7. ENVELOPE AT FAR FIELD - DESIGN #1

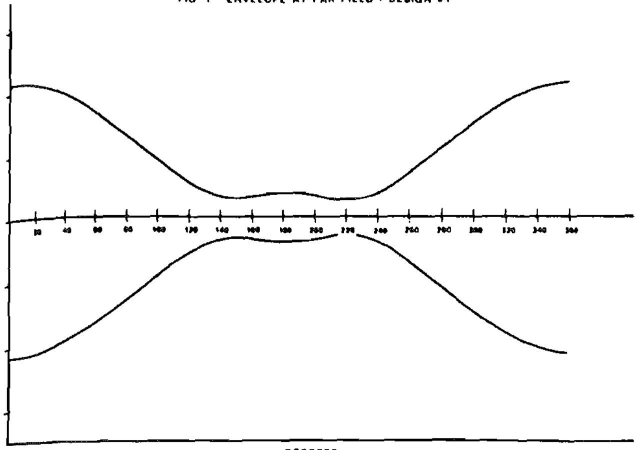


FIG. 8. THD - DESIGN #1

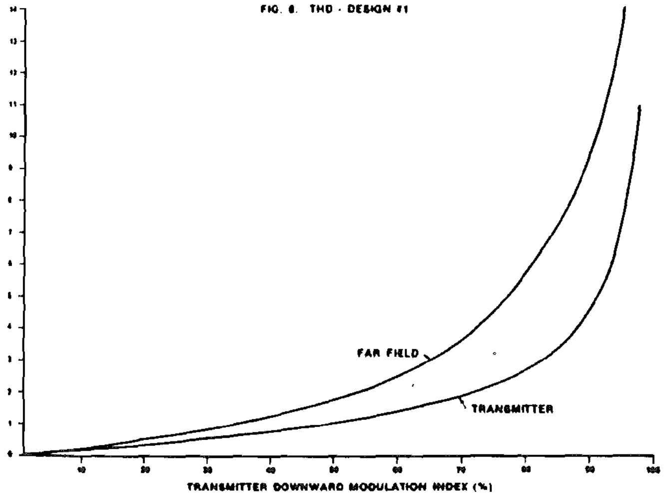


FIG. 9. MODULATION INDEX COMPARISON - DESIGN #1

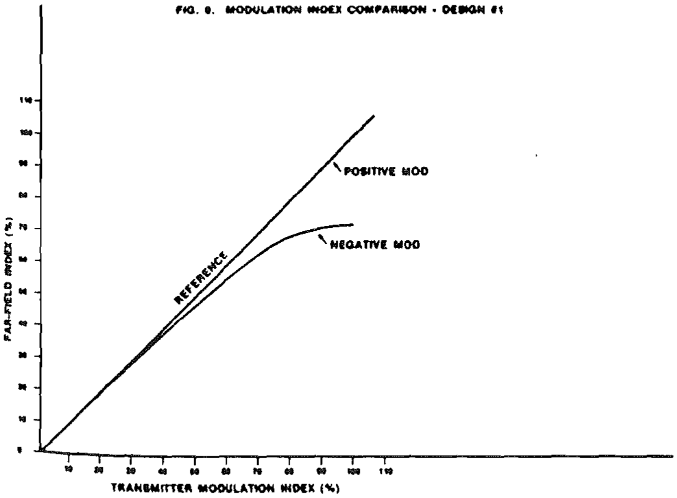


FIG. 10. ENVELOPE AT TRANSMITTER - DESIGN #2

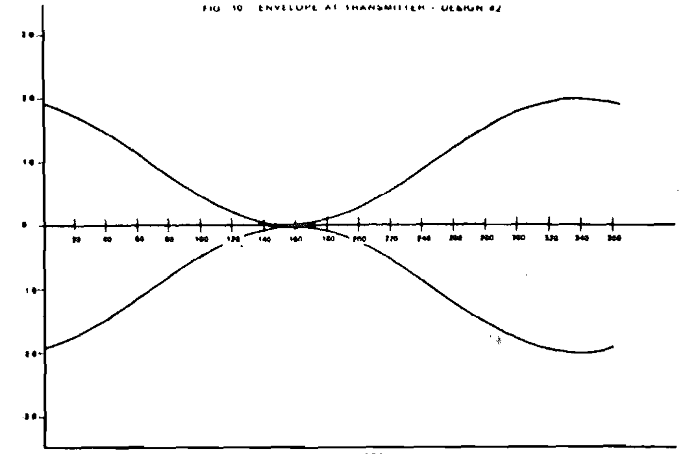


FIG. 11. ENVELOPE AT FAR FIELD DESIGN #2

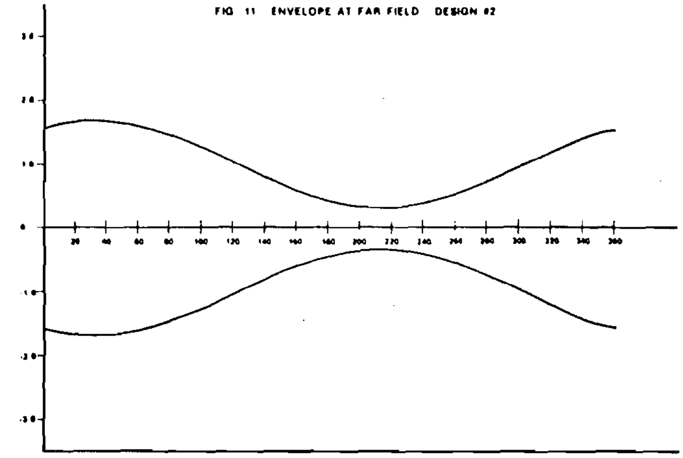


FIG. 12. MODULATION INDEX COMPARISON - DESIGN #2

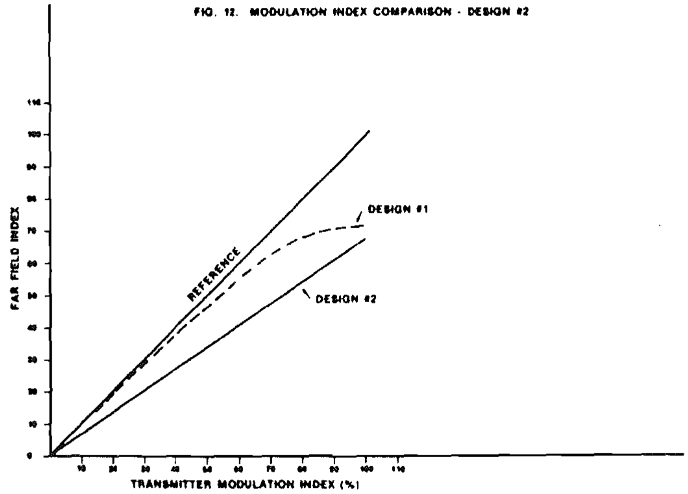


FIG 13 MODULATION INDEX COMPARISON - DESIGN #3

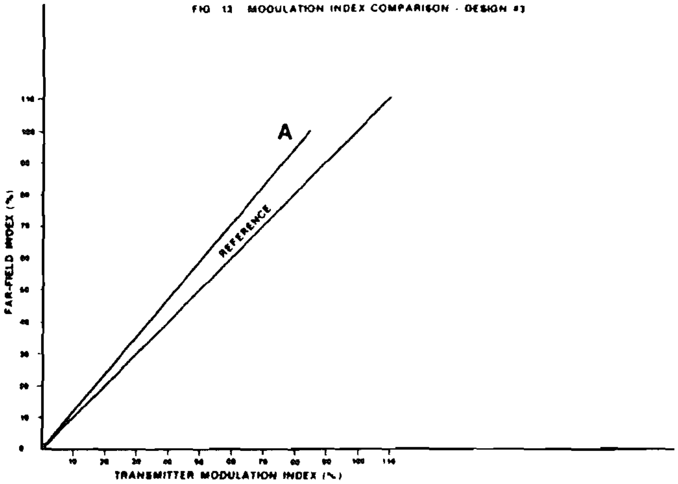


FIG. 14. ENVELOPE AT TRANSMITTER - DESIGN #3