DESIGN OF TUBE AMPLIFIERS FOR OPTIMUM FM TRANSMITTER PERFORMANCE

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INTRODUCTION

Tube amplifiers are widely used in frequency modulation (FM) broadcast transmitters to increase the level of the FM signal at the wideband exciter or Intermediate Power Amplifier (IPA) output to higher power output levels. Tube amplifiers are more efficient and cost effective at high power levels than a combination of several low power solid-state amplifiers in the 88-108 MHz FM broadcast band.

The power amplifier (PA) is typically a high gain single-tube type operated as a tuned class "C" radio frequency (RF) amplifier. The PA design goal is to deliver the authorized power output to the antenna with high efficiency and reliability while providing excellent modulation performance.

This paper discusses various topologies of the input and output circuits of a vacuum tube power amplifier and analyzes their effects on the transmitter amplitude and group delay responses. Results of computer circuit analysis and actual measured data of a typical transmitter with two different topologies are compared. Design considerations for optimum transmitter performance to achieve desired level of transparency to a wideband FM broadcast signal is also discussed including recommendations for compensating the group delay of the transmission system. The contents of the paper are divided into the following main headings:

1. Frequency Modulated Signal And Effects Of Bandwidth Limitation On The Transmitter Performance.

2. Power Amplifier Design Considerations.
   - Primary Design Factors.
   - Input Circuit Configurations And Their Effects On The Transmitter Bandwidth.
   - Output Circuit Configurations And Their Effects On The Transmitter Bandwidth.
   - Computed/Measured Amplitude And Group Delay Responses.

3. Modulation Performance Of A Typical 20 kW Single-Tube FM Transmitter.

Frequency Modulated Signal

A Frequency Modulated RF Signal with modulation index "m", carrier frequency "fc" and single-tone modulation frequency "fm" can be represented by the following mathematical expression [1-6]:

$$E(t) = Ec \cdot Cos[\omega c \cdot t + m \cdot Sin(\omega m \cdot t)],$$

Where:

Ec = The unmodulated carrier amplitude constant
$\omega c = 2\pi \cdot fc$ (carrier frequency)
$\omega m = 2\pi \cdot fm$ (modulating frequency)
$m = \Delta f/fm$ = frequency deviation/modulating frequency

In an FM signal, the deviation "Δf" of the instantaneous frequency from the average (or the carrier frequency) is directly proportional to the instantaneous amplitude of the modulating signal. The rate of frequency deviation is the modulating signal frequency.

The above FM Signal can be expressed as an infinite series of discrete spectral components using trigonometric expansions and series representations of Bessel functions [1-5].

$$E(t) = Ec \cdot \sum_{n=-\infty}^{\infty} Jn(m) \cdot Cos[(\omega c + n \cdot \omega m) \cdot t],$$

Where Jn(m) are Bessel functions of the first kind and nth order with argument "m". The numeric values of the Bessel functions Jn(m) for different "n" express the amplitudes of the various frequency components relative to the unmodulated carrier amplitude. The values of Jn(m) depend on the argument "m" and the order "n". These can be found from the mathematical tables.

For example:

If n = 0, Jn(m) = Jo(m), which is the amplitude of the carrier component.

If n = ±1, Jn(m) = J1(m) and J-1(m), which are the amplitudes of the first order sideband components.

If n = ±2, Jn(m) = J2(m) and J-2(m), which are the amplitudes of the second order sideband components.

Figure 1 shows a graphical representation of how the Bessel function values for the carrier and the first twelve pairs of sidebands vary with the modulation index [1-4]. Bessel function value diminishes as the order of the function "n" gets larger. As the modulation index varies, the carrier or the sideband frequency pair may vanish completely. Representative spectrum plots for six different values of modulation index are shown in Figures 2 and 3 [4]. In FM the energy that goes into the sideband frequencies is taken from the carrier; the total power in the overall composite signal remains the same regardless of the modulation index.

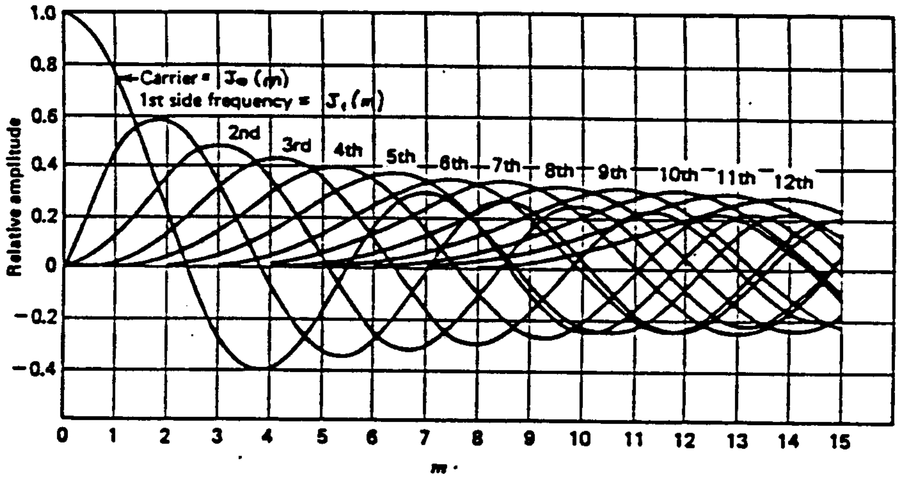


FIGURE 1. PLOT OF BESSEL FUNCTION OF FIRST KIND AS A FUNCTION OF ARGUMENT m.

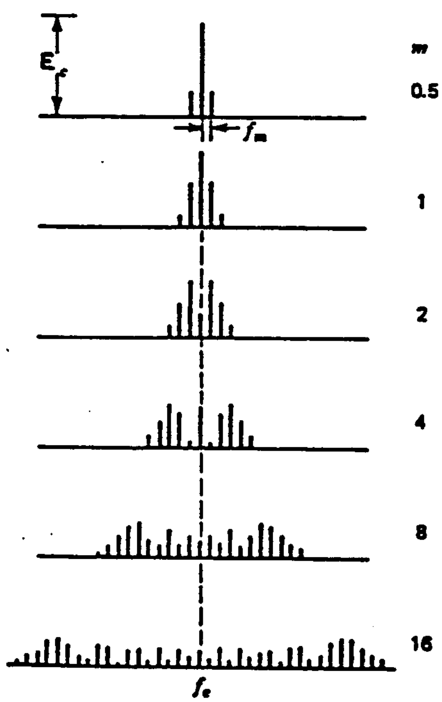


FIGURE 2. FREQUENCY SPECTRUM PLOTS FOR SINUSOIDAL MODULATION, WITH VARIOUS VALUES OF "m" ("fm" IS CONSTANT AND "△f" IS VARIED).

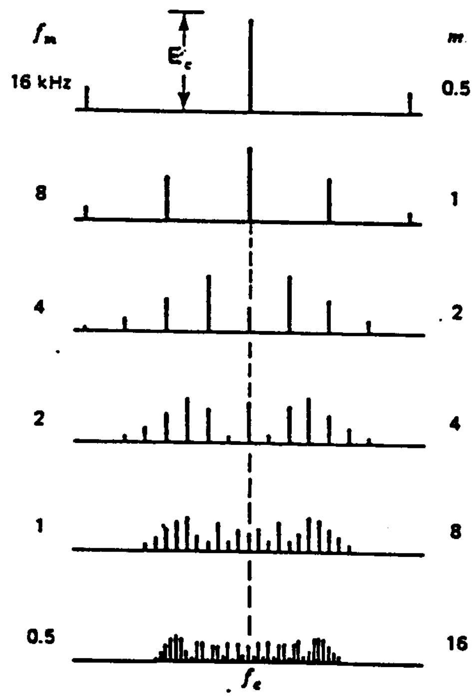


FIGURE 3. FREQUENCY SPECTRUM PLOTS FOR SINUSOIDAL MODULATION, WITH VARIOUS VALUES OF "m" ("△f" IS CONSTANT AND "fm" IS VARIED).

Occupied Signal Bandwidth

Occupied Signal Bandwidth "BW" of an FM signal can be calculated for a single tone modulation by the following formula:

$$BW = 2 \cdot n \cdot fm.$$

Where "n" is the number of significant sideband components which depends on the value of Jn(m) and changes with the modulation index "m". "fm" is the modulating frequency.

The value of "n" can be accurately found from the mathematical tables by ignoring sideband components with amplitudes Jn(m) less than a certain desired number. The maximum value of "n" which need be considered for a given "m" may be found from the following empirical expression [5]:

$$n = m + k \cdot (m)^{0.27}$$

where "k" is 2.4 for Jn(m) = 0.01, and 3.5 for Jn(m) = 0.001.

For a single tone 15 kHz modulation with 75 kHz deviation, the modulation index is "5". If we ignore components with amplitudes less than 1% ($Jn(5) < 0.01$), the number "n" is 9. The bandwidth required is 270 kHz.

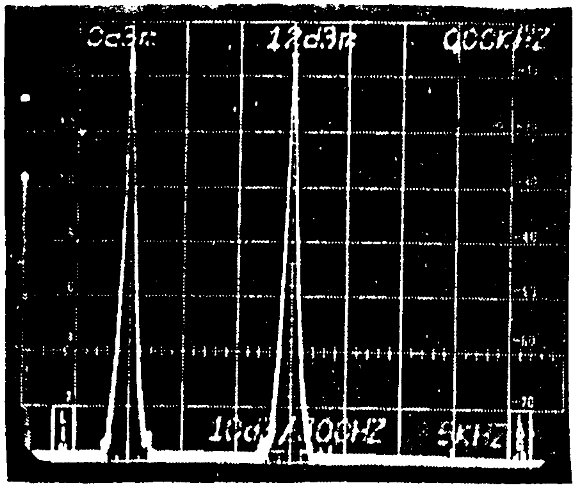
The signal bandwidth for stereo Left or Right only single tone 15 kHz modulation is typically less than that for monaural modulation. This is due to the reduction in modulation index. The frequency deviation is held constant at 75 kHz but the composite baseband spectral components comprise of modulation frequencies at 15 kHz ("L + R" Main Channel), 19 kHz (Pilot), 23 kHz and 53 kHz ("L - R" Subcarrier Channel). The bandwidth calculation for multiple tone or stereo FM modulation is quite complex because several combination frequencies must be accounted for. This is caused by the nonlinear process inherent to frequency modulation.

Effects Of Bandwidth Limitations

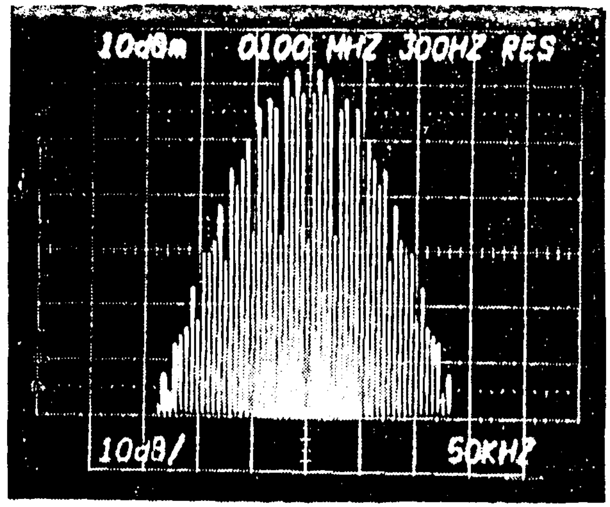
Typically, bandwidth limitations occur when the FM signal passes through the RF path comprising the transmitter, filterplexer/combiner, antenna system and the particularly in the receiver.

Figure 4 shows the effect of passing a twin-tone (10 kHz/25 kHz) modulated FM signal through a wideband and a narrowband (-3 dB bandwidth of 400 kHz) RF path [7]. The top three pictures show that the wideband RF path has negligible effect on the demodulated audio signal. But the bottom three pictures show the distortion products caused by bandwidth restriction.

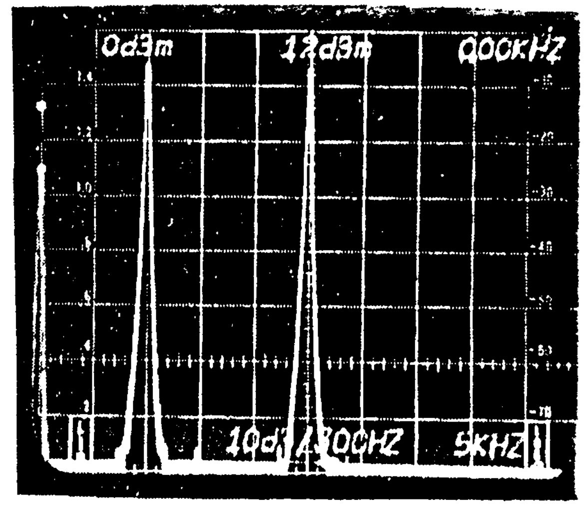
TWO TONE (10KHz & 25KHz) MODULATION THRU WIDEBAND RF PATH



BASEBAND SPECTRUM TO BEI MODEL FX-30 MODULATOR

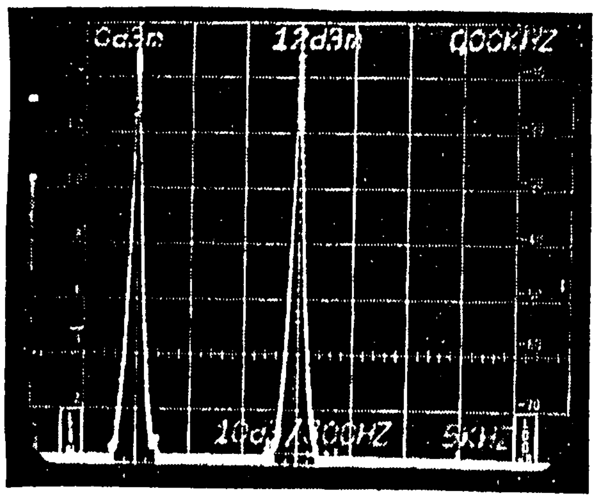


RF SPECTRUM TO DEMODULATOR

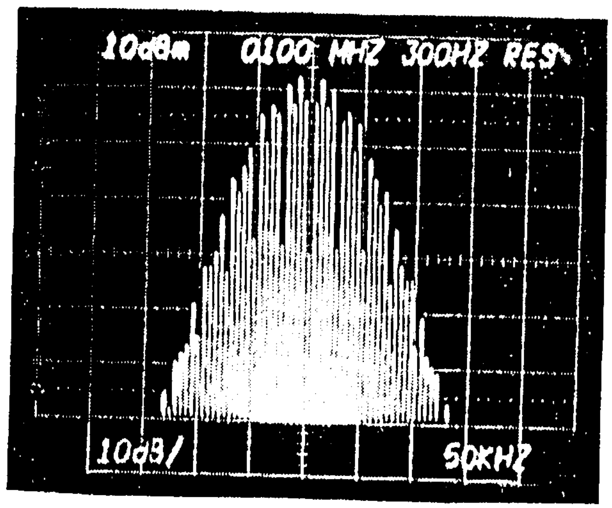


DEMODULATED BASEBAND SPECTRUM FROM BOONTON MODEL 82AD

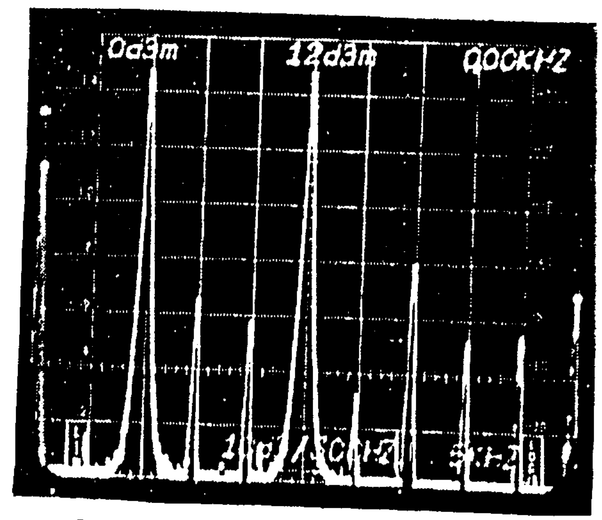
TWO TONE (10KHz & 25KHz) MODULATION THRU NARROWBAND RF PATH



BASEBAND SPECTRUM TO BEI MODEL FX-30 MODULATOR



BANDWIDTH LIMITED RF SPECTRUM TO DEMODULATOR



DEMODULATED BASEBAND SPECTRUM FROM BOONTON MODEL 82AD

FIGURE 4. THE EFFECT OF PASSING A TWIN-TONE (10 kHz/25 kHz) MODULATED FM SIGNAL THROUGH A WIDEBAND AND A NARROWBAND RF PATH.

Effects On The Transmitter Modulation Performance

The following is a summary of results based on a study done by Broadcast Electronics to determine how much bandwidth is required for low distortion FM transmission, and at what bandwidth the point of diminishing return regarding the audio performance improvement is reached. This information has been published by Broadcast Electronics [8].

Composite Total Harmonic Distortion And Noise (THD+N) At 15 kHz

THD is better than 0.1% with more than 600 kHz bandwidth. There is no improvement in performance above 1.5 MHz as shown in Figure 5.

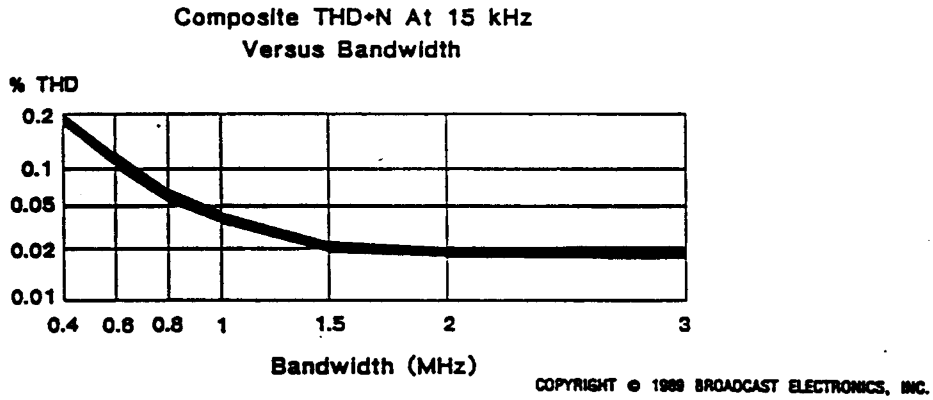


FIGURE 5. COMPOSITE THD+N AT 15 kHz VERSUS BANDWIDTH

Composite SMPTE Intermodulation Distortion

SMPTE IMD (60Hz/7kHz 1:1 Ratio) is better than 0.1% with 400 kHz bandwidth and crosses 0.01% with 1 MHz bandwidth as shown in Figure 6.

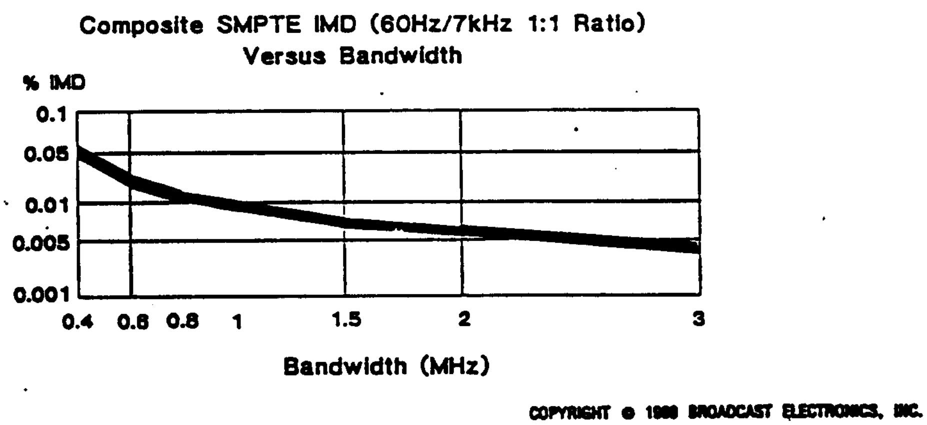


FIGURE 6. COMPOSITE SMPTE IMD (60 Hz/7 kHz 1:1 RATIO) VERSUS BANDWIDTH

Stereophonic Total Harmonic Distortion And Noise (THD+N) At 15 kHz

Stereophonic THD at 15 kHz modulation is better than 0.1% with 600 kHz bandwidth. There is no improvement above 1 MHz as shown in Figure 7.

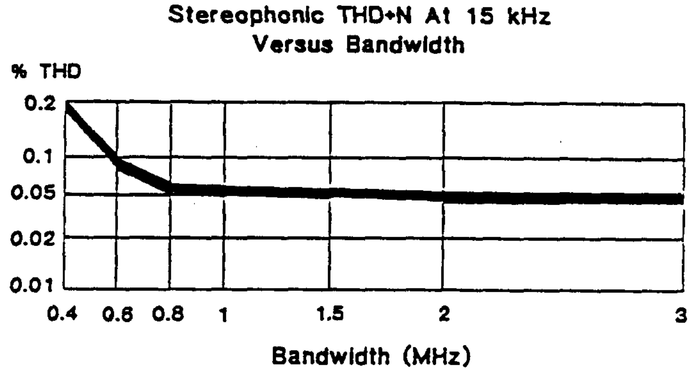


FIGURE 7. STEREOPHONIC THD+N AT 15 kHz VERSUS BANDWIDTH

Stereophonic SMPTE Intermodulation Distortion

Figure 8 shows that the Stereophonic IMD is better than 0.05% with 600 kHz bandwidth. There is very little improvement above 1 MHz.

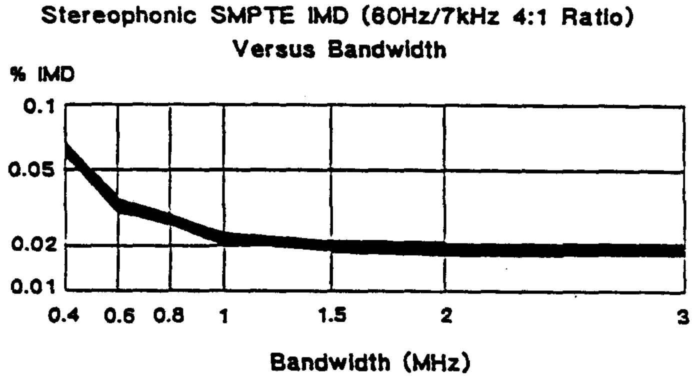


FIGURE 8. STEREOPHONIC SMPTE IMD (60 Hz/7 kHz 4:1 RATIO) VERSUS BANDWIDTH

Stereo Separation

Stereo Separation is about 40 dB with 400 kHz bandwidth, 50 dB with 700 kHz and 60 dB with 1.5 MHz as shown in Figure 9.

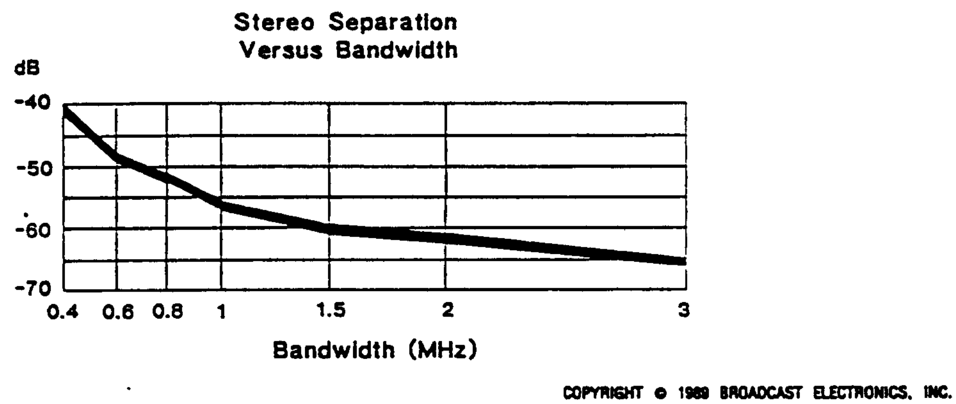


FIGURE 9. STEREO SEPARATION VERSUS BANDWIDTH

Synchronous Amplitude Modulation

Figure 10 shows that Synchronous AM is better than 40 dB with 800 kHz bandwidth and better than 50 dB with 1.5 MHz.

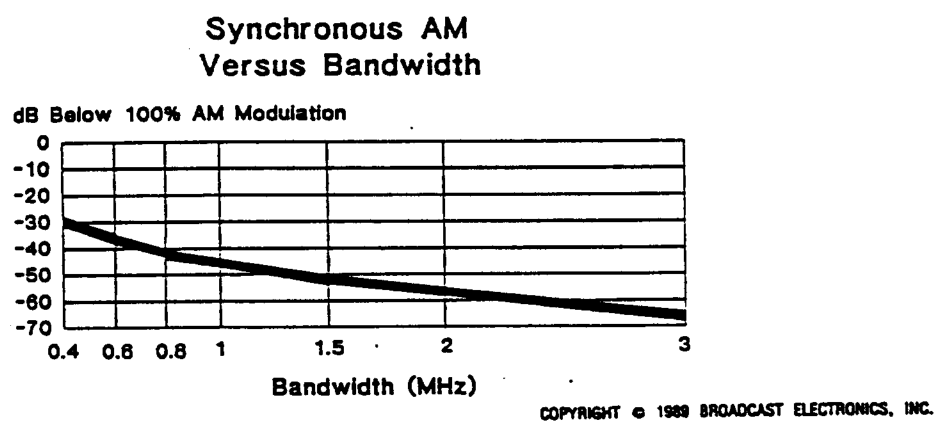


FIGURE 10. SYNCHRONOUS AM VERSUS BANDWIDTH

The conclusion drawn from the above study is that a minimum -3 dB bandwidth of 800 kHz is required for good audio performance and that excellent performance can be achieved with 1 to 1.5 MHz bandwidth [8].

Effects On The Transmitter RF Intermodulation

Frequency Spectrum is a very limited natural resource. The FM broadcast band is shared by several users at the same location. When multiple signals are present, any non-linear device such as tube in the transmitter power amplifier will generate RF intermodulation products due to mixing of these multiple signals. This mixing will have some conversion loss called "turn-around-loss". The degree of intermodulation interference generated within a given system can be accurately predicted when the turn-around-loss of the transmitter is available.

Turn-Around-Loss depends on three factors [9].

- In-band Conversion Loss
- Interfering Signal Attenuation due to PA Output Selectivity
- Attenuation of Resulting IM products due to PA Output Selectivity.

The transmitter with a narrower bandwidth will have higher selectivity thereby making it more immune to RF intermodulation. Therefore, there is certainly a trade-off between modulation performance and immunity from RF intermodulation.

POWER AMPLIFIER DESIGN CONSIDERATIONS

Primary Design Factors

The primary factors which should be considered in Power Amplifier design are:

- Desired Power Output
- Optimum Modulation Performance
- High Efficiency and Reliability
- Best Value for the Cost

This paper will focus its discussion on the second item - the design considerations necessary to achieve the optimum modulation performance.

Power Amplifier Bandwidth

The transmitter power amplifier bandwidth affects the modulation performance. Available bandwidth determines the amplitude response, phase response, and group delay response. There is a trade-off involved between the bandwidth, gain, and efficiency in the design of a power amplifier.

Power amplifier bandwidth is restricted by the equivalent load resistance across parallel tuned circuits. Tuned circuits are necessary to cancel low reactive impedance presented by relatively high input and output capacitance of the amplifying device such as a vacuum tube.

The bandwidth for a single tuned circuit is proportional to the ratio of capacitive reactance, Xc to load resistance, R_L (appearing across the tuned circuit) [10]:

$$BW \cong \frac{K}{2 \cdot \pi \cdot fc \cdot R_L \cdot C} \cong \frac{K(Xc)}{R_L}$$

Where: BW = bandwidth between half-power (or -3 dB) points
K = proportionality constant
R_L = load resistance (appearing across tuned circuit)
C = total capacitance of tuned circuit (includes stray capacitances and output or input capacitances of the tube)
Xc = capacitive reactance of C
fc = carrier frequency

The RF voltage swing across the tuned circuit also depends on the load resistance. For the same power and efficiency, the bandwidth can be increased if the capacitance is reduced.

Input Circuit Configurations And Their Effects On The Transmitter Bandwidth

Newer transmitter designs utilize solid-state intermediate power amplifiers to provide necessary RF drive level to operate the tube in the class C mode. The output load impedance is typically 50 Ohms. It is, therefore, necessary to design a matching network to transform a high grid input impedance to 50 Ohms at the PA input. The following three types of input matching circuit configurations are used:

- Single Element Capacitive (C) Input Match
- Single Element Inductive (L) Input Match
- Broadband (L-C) Input Match.

The first two are the popular ones. These matching circuits have different effects on the PA amplitude and group delay responses.

Capacitive Input Match

Single Element Capacitive Input Matching Circuit is shown in Figure 11 [6,10]. This is the simplest in design as well as implementation. Variable inductor "Lin" tunes out the tube input capacitance past parallel resonance to make the input impedance slightly inductive. This is necessary to transform the high grid impedance to an equivalent series 50 Ohms resistance and some inductive reactance. The reactance is then tuned out with a series variable capacitor, "Cs". Interactive adjustment of "Lin" and "Cs" is required. This configuration has the characteristics of a high pass filter.

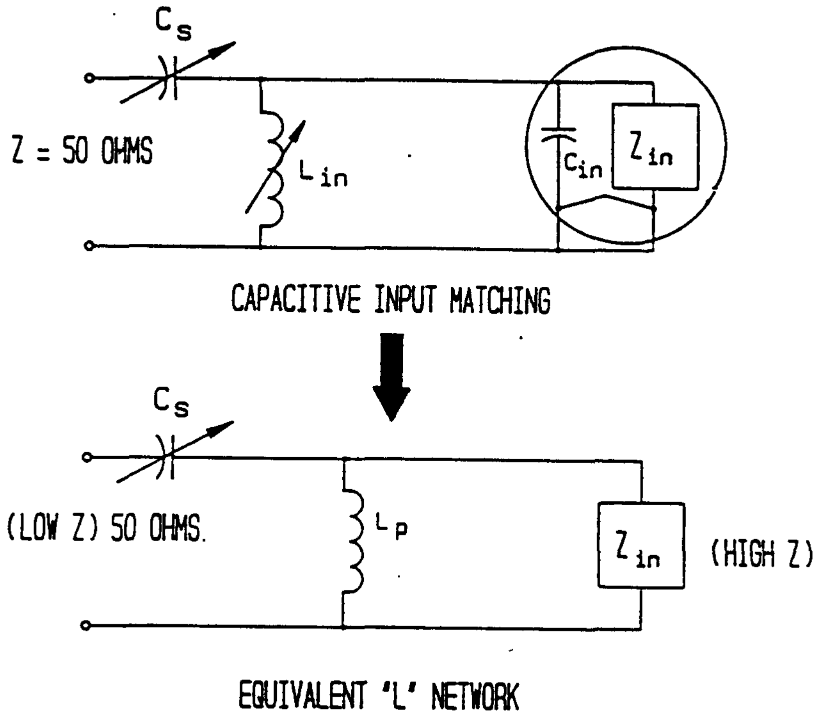


FIGURE 11. CAPACITIVE INPUT MATCHING CIRCUIT

Inductive Input Match

Single Element Inductive Input Matching Circuit is shown in Figure 12 [6,10]. This is the next most popular method for input matching. The variable inductor "Lin" is used to tune out the input capacitance slightly before the parallel resonance is reached, on the capacitive side. This is necessary to transform the high grid impedance to an equivalent series 50 Ohms resistance with some capacitive reactance. The reactance is then tuned out with a series variable inductor "Ls". Interactive adjustment of "Lin" and "Ls" is required. In this circuit configuration, a part of the input capacitance is used for impedance transformation and the equivalent capacitance across the tuned circuit becomes less, thereby increasing the bandwidth. This configuration has the characteristics of a low pass filter.

Broadband L-C Input Match

Broadband L-C Input Matching Circuit is shown in Figure 13 [10]. This is an extension of L-Match circuit. It utilizes multiple L-C sections with each section providing a small step in the total impedance transformation. This technique provides a broadband impedance match without interactive adjustment and improves the transmitter operation. This configuration also utilizes a part of the input capacitance for impedance transformation and has the characteristics of a multiple section low pass filter.

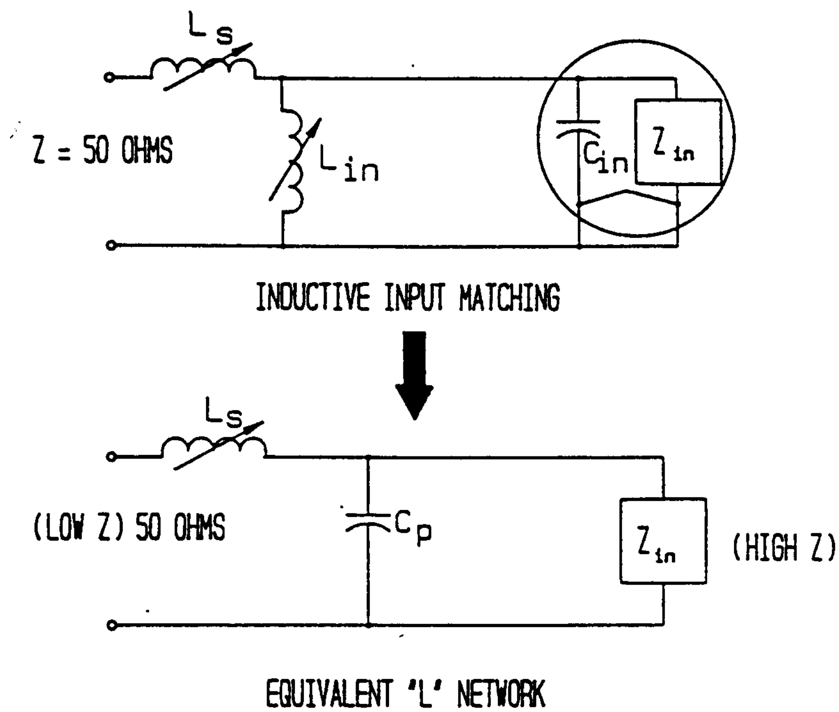


FIGURE 12. INDUCTIVE INPUT MATCHING CIRCUIT

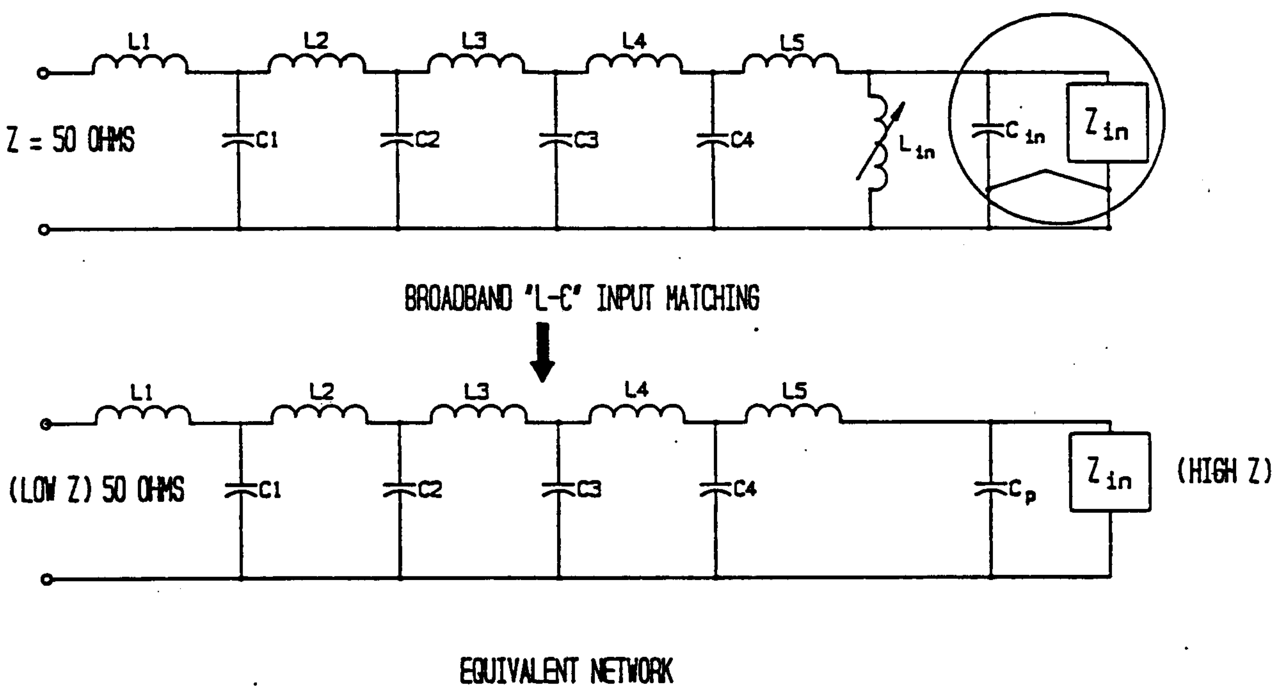


FIGURE 13. BROADBAND L-C INPUT MATCHING CIRCUIT

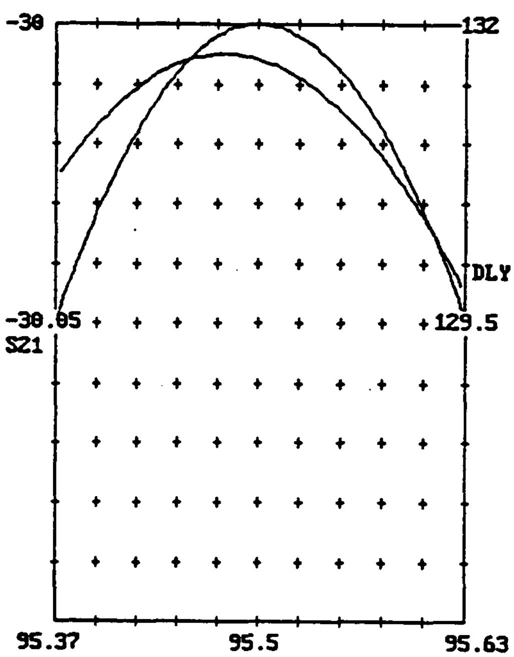
Computed And Measured Amplitude/Group Delay Responses Of Capacitive And Broadband Input Matching Circuits

The results of computer analysis and actual measurements made with two different input circuit configurations, C-Match and Broadband L-C Match, in a real transmitter operating at 20 kW RF power are presented below.

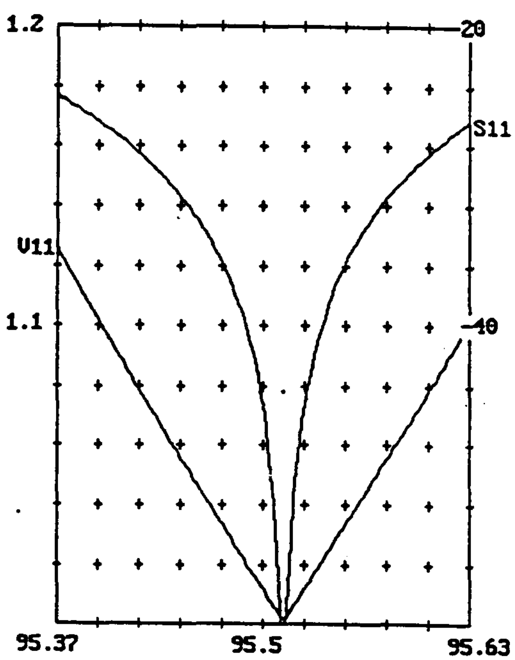
Two input matching circuits were designed to transform the high impedance of PA input tuned circuit to 50 Ohms resistive impedance. For the purpose of illustration it was assumed that a total capacitance of 165 picofarads in parallel with 375 Ohms load resistance appeared across the tuned circuit. This represents the input impedance of a typical 20 kW FM transmitter power amplifier input circuit utilizing an Eimac 8989 tetrode.

Computer software programs ECA [11] and =SuperStar= [12] were used to analyze the circuit designs and obtain the response plots. Hewlett Packard Network Analyzer Model 3577A was used to obtain the measured response plots.

Figure 14A shows the computed amplitude response (S21) and group delay (DLY) response plots for capacitive input matching circuit. Figure 14B shows the input VSWR (V11) and input return loss (S11) plots at -0.05 dB response points. The peaks of amplitude and group delay plots do not coincide in the case of capacitive input matching circuit. Figure 15 shows the measured amplitude response, group delay response and input return loss plots of a typical 20 kW FM transmitter at the tube grid ring for capacitive input matching circuit.



14A. AMPLITUDE (S21) AND GROUP DELAY (DLY)



14B. INPUT VSWR (V11) AND RETURN LOSS (S11) RESPONSES

FIGURE 14. COMPUTED AMPLITUDE, GROUP DELAY, VSWR, AND RETURN LOSS RESPONSES OF A CAPACITIVE INPUT MATCHING CIRCUIT

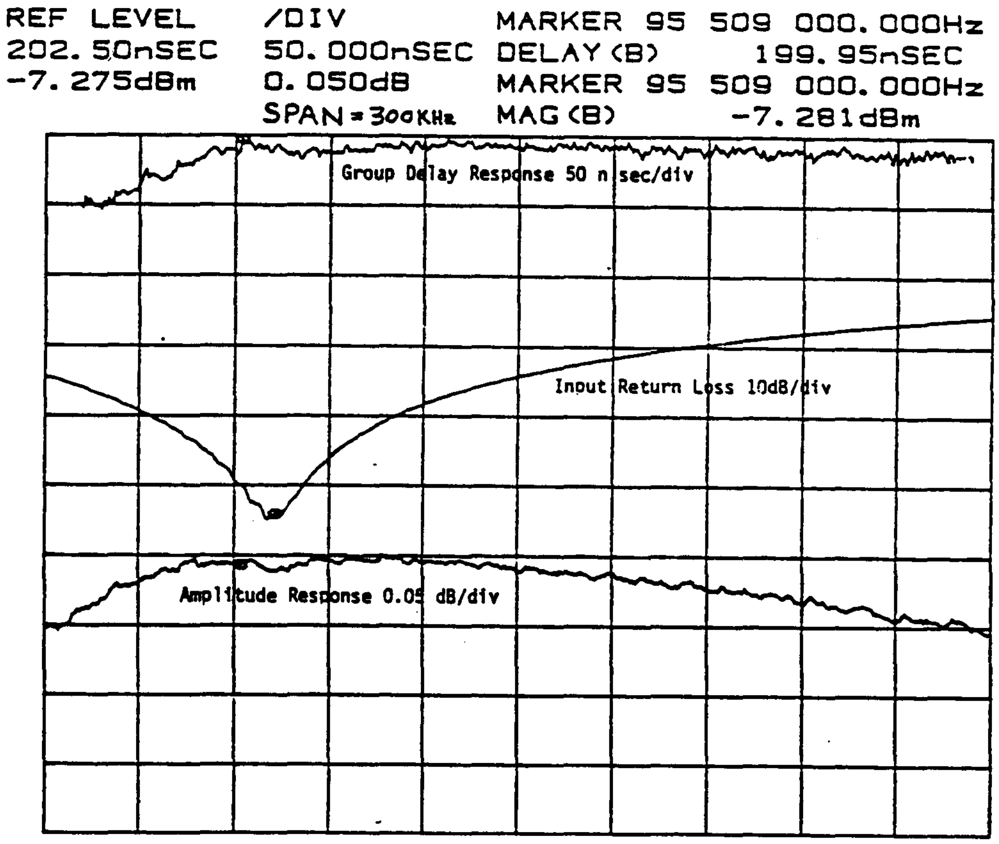
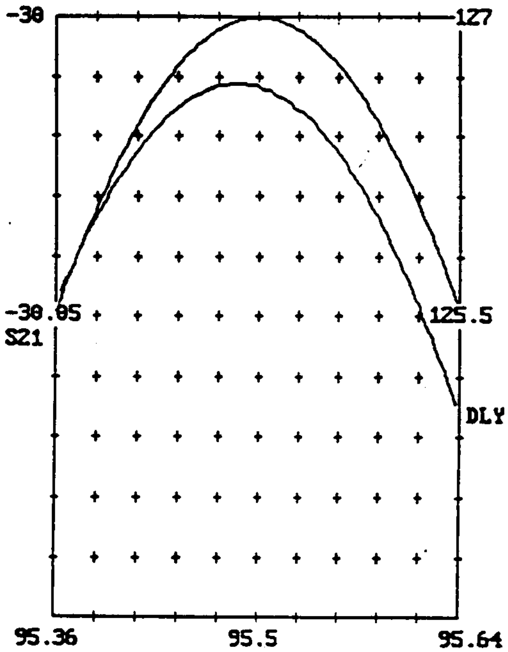
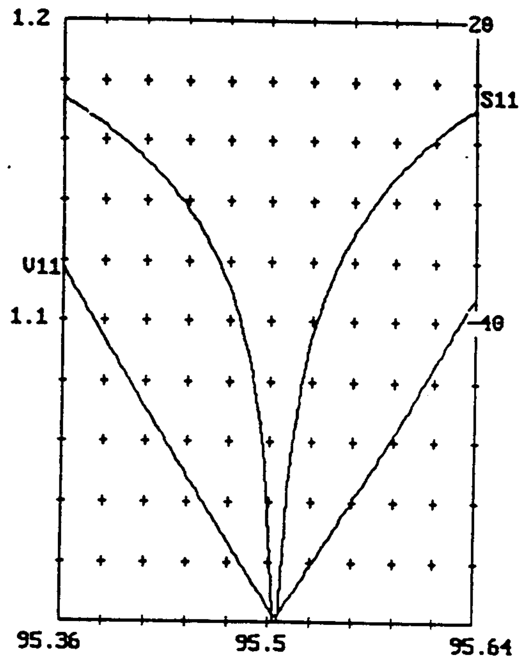


FIGURE 15. MEASURED AMPLITUDE, GROUP DELAY, AND INPUT RETURN LOSS RESPONSES OF A CAPACITIVE INPUT MATCHING CIRCUIT IN A 20 kW TRANSMITTER

Figure 16A shows the computed amplitude response (S21) and group delay (DLY) response plots for broadband input matching circuit. Figure 16B shows the input VSWR (V11) and input return loss (S11) plots at -0.05 dB response points. The peaks of amplitude and group delay plots coincide closely to provide a symmetrical group delay response. Figure 17 shows the measured amplitude response, group delay response and input return loss plots of a typical 20 kW FM transmitter at the tube grid ring for broadband input matching circuit.



16A. AMPLITUDE (S21) AND GROUP DELAY (DLY) RESPONSES



16B. INPUT VSWR (V11) AND RETURN LOSS (S11) RESPONSES

FIGURE 16. COMPUTED AMPLITUDE, GROUP DELAY, VSWR, AND RETURN LOSS RESPONSES OF A BROADBAND L-C INPUT MATCHING CIRCUIT

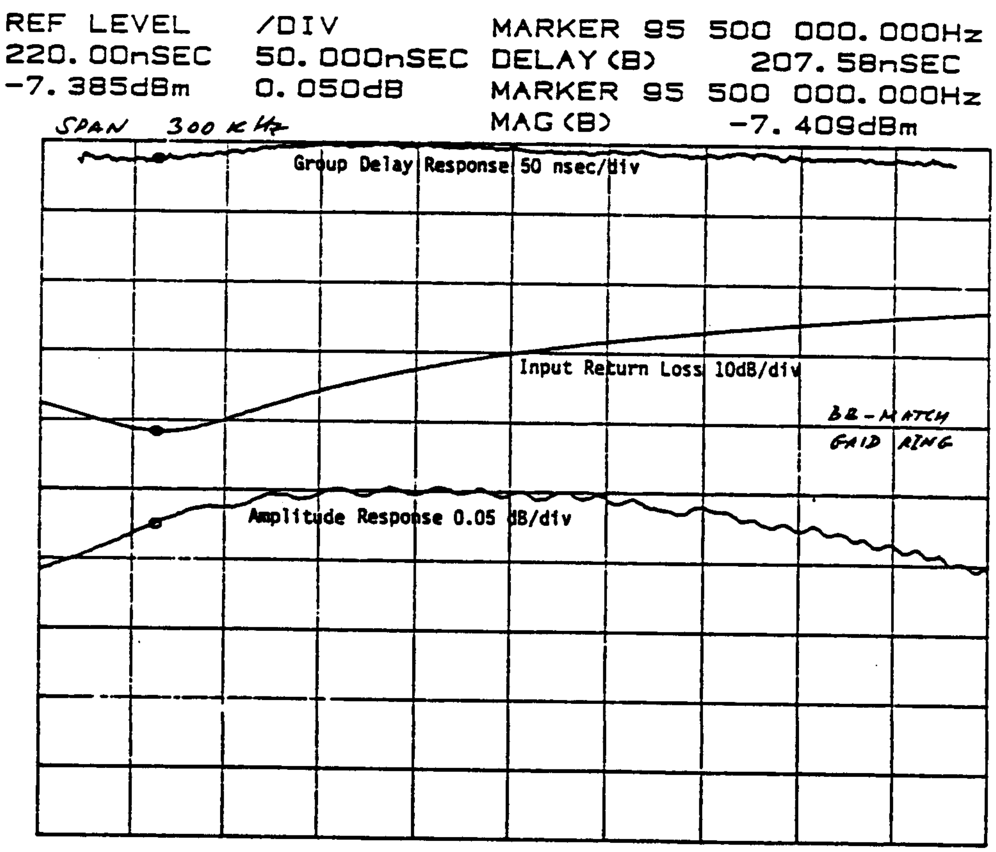


FIGURE 17. MEASURED AMPLITUDE, GROUP DELAY, AND INPUT RETURN LOSS RESPONSES OF A BROADBAND INPUT MATCHING CIRCUIT IN A 20 kW TRANSMITTER

Figure 18 shows the comparison of computed and measured bandwidth of a typical 20 kW FM transmitter at tube grid ring. Broadband L-C input matching circuit has a higher bandwidth than the capacitive input matching circuit.

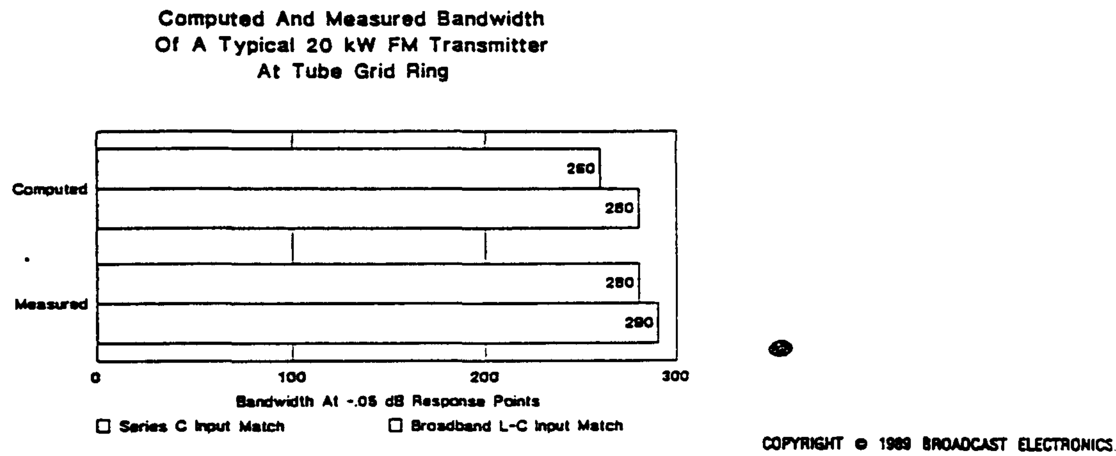


FIGURE 18. COMPUTED AND MEASURED BANDWIDTH OF A TYPICAL 20 kW FM TRANSMITTER AT TUBE GRID RING

Figure 19 shows the comparison of computed and measured asymmetry of amplitude/delay response peaks of a typical 20 kW FM transmitter at tube grid ring. Capacitive input matching circuit has a greater group delay asymmetry than either the inductive or broadband L-C matching circuit.

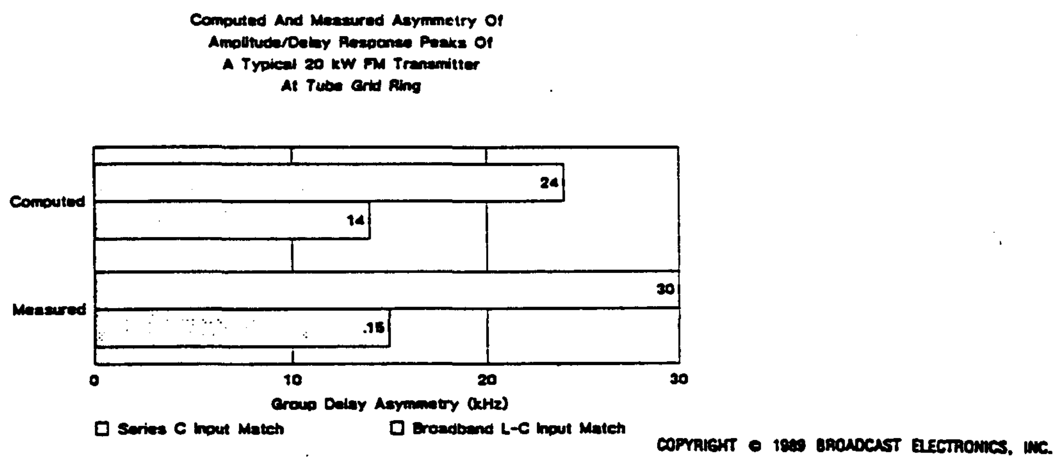


FIGURE 19. COMPUTED AND MEASURED ASYMMETRY OF AMPLITUDE/DELAY RESPONSE PEAKS OF A TYPICAL 20 kW FM TRANSMITTER AT TUBE GRID RING

Output Circuit Configurations And Their Effects On The Transmitter Bandwidth

Tuned circuits are required to resonate the output capacitance of the vacuum tube and to present a fairly high impedance between anode and cathode (screen grid in case of tetrodes) to the fundamental carrier frequency component and present a low impedance to the harmonics. The output circuit may utilize a lumped inductor, a strip line, or higher "Q" (low loss) transmission line section to resonate the tube capacitance.

In newer design transmitters, the tube power amplifier is typically constructed in a cavity enclosure utilizing (larger physical size) coaxial transmission line section of either quarter-wavelength, or half-wavelength to increase the unloaded "Q" and minimize losses.

The Power Amplifier efficiency depends on the RF plate voltage swing developed across the load resistance, the plate current conduction angle and the cavity efficiency. The PA cavity efficiency is related to the ratio of loaded to unloaded "Q" as follows [10]:

$$N = 1 - \frac{QL}{QU} \times 100$$

Where: N = efficiency in percent
QL = loaded "Q" of cavity
QU = unloaded "Q" of cavity

The loaded "Q" is dependent on the equivalent plate load resistance impedance presented across the tuned circuit and output circuit capacitance. Unloaded "Q" depends on the cavity volume and the RF resistivity of the conductors due to skin effects. A high unloaded "Q" is desirable, as is a low loaded "Q", for best efficiency. As the loaded "Q" goes up the bandwidth decreases. For a given tube output capacitance and power level, loaded "Q" decreases with decreasing plate voltage or with increasing plate current. The increase in bandwidth at reduced plate voltage occurs because the load resistance is directly related to the RF voltage swing on the tube element.

For the same power and efficiency, the bandwidth can also be increased if the output capacitance is reduced. Power tube selection and minimization of stray capacitance are areas of particular concern in PA design for maximum bandwidth. Bandwidth can be further improved by minimizing added tuning capacitance and by properly selecting the output coupling method.

The following are the two popular methods of output coupling circuits used:

- Series Capacitive Output Coupling
- Magnetic Output Coupling Loop.

These circuits have different effects on the PA amplitude and group delay responses.

Series Capacitive Output Coupling

Figure 20 shows a schematic of a tetrode power amplifier with a capacitive output tuning and capacitive output coupling circuit [6,10].

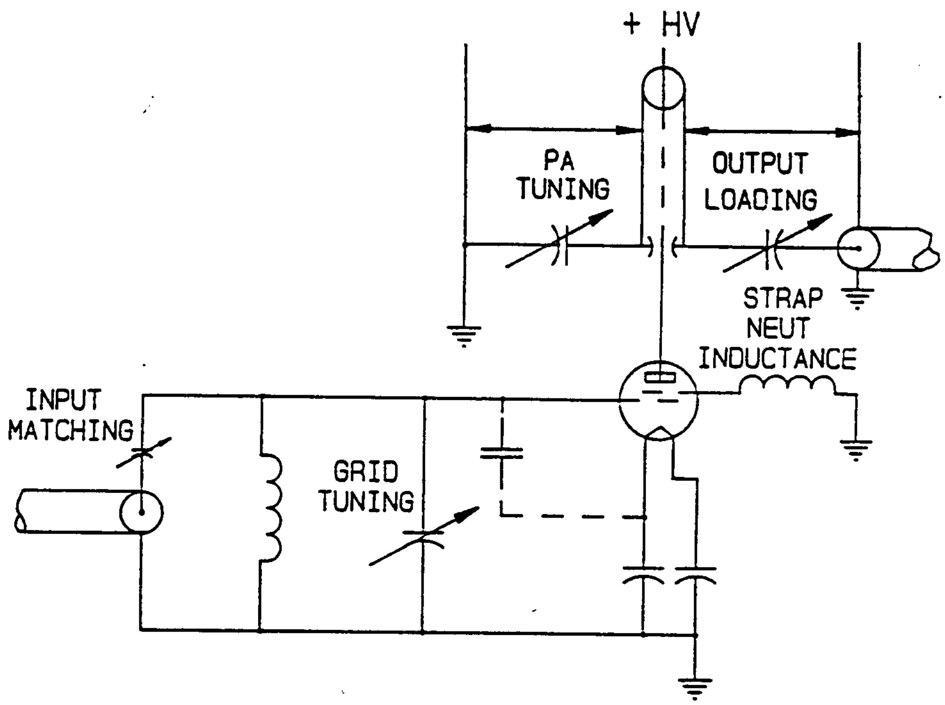


FIGURE 20. TETRODE POWER AMPLIFIER CIRCUIT WITH CAPACITIVE OUTPUT TUNING AND CAPACITIVE OUTPUT COUPLING

Figure 21 is a simplified diagram showing the construction of this type of output circuit in a quarter-wave length coaxial cavity [6,10]. The tube anode is coupled through a DC blocking capacitor to the transmission line. The tube's output capacitance is brought to resonance by the inductive component of the transmission line that is physically less than a quarter-wave length long. The coarse output tuning is done by moving the ground plane at low impedance end of the line. Variable capacitors are used to fine-tune the output resonant circuit and to couple the power from the high RF voltage point located at the anode end of the quarter-wave line to the load. An interactive adjustment of tuning and loading is required in this type of power amplifier cavity design. This type of output coupling is similar to a series capacitive input matching circuit discussed above.

A capacitive output coupling circuit was designed to transform the high impedance of the PA output tuned circuit (with total capacitance of 20 picofarads and 1590 Ohms load resistance across it) to 50 Ohms. This illustrates the output circuit of a typical 20 kW FM transmitter PA utilizing an Eimac 8989 tetrode. Computer software programs ECA [11] and =SuperStar= [12] were used to analyze the circuit design and obtain the response plots.

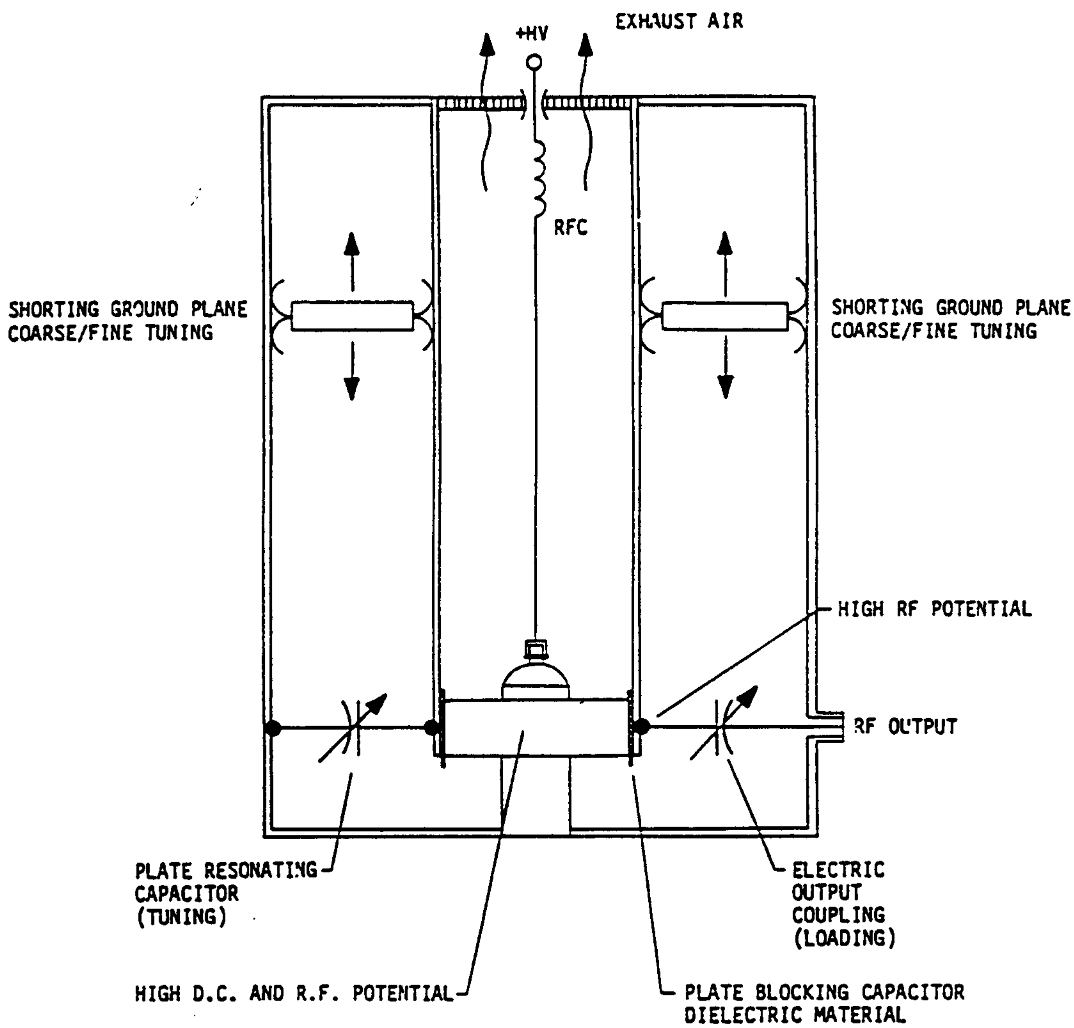
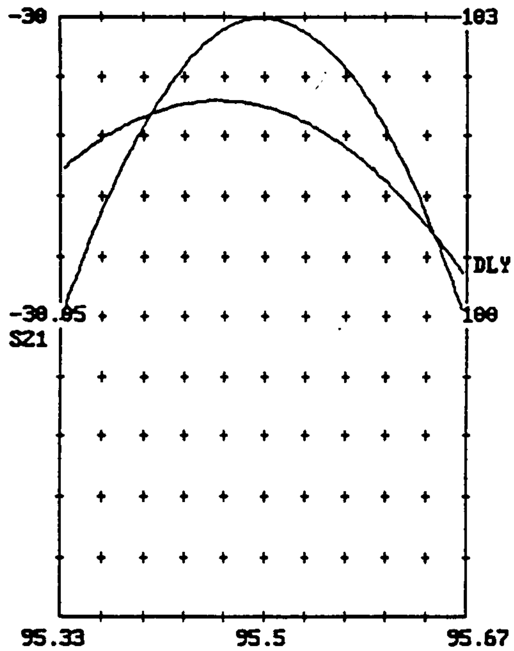
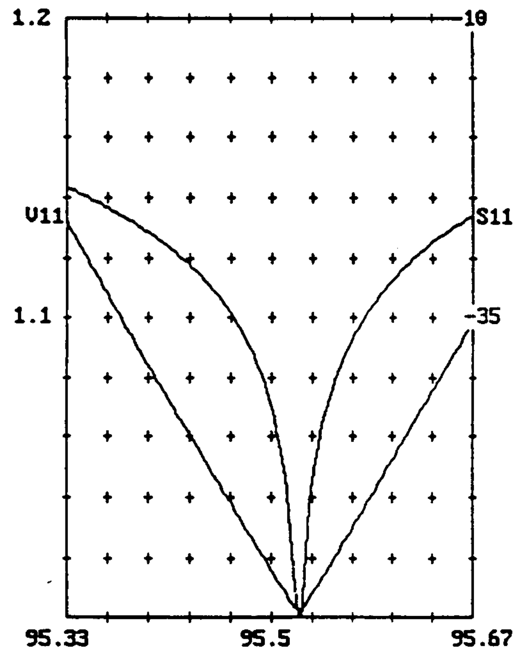


FIGURE 21. QUARTER-WAVE COAXIAL CAVITY WITH CAPACITIVE TUNING AND CAPACITIVE OUTPUT COUPLING

Figure 22A shows the computed amplitude response (S21) and group delay (DLY) response plots. Figure 22B shows the output VSWR (V11) and output return loss (S11) plots at -0.05 dB response points. The peaks of amplitude and group delay plots do not coincide in the case of capacitive output coupling circuit.



22A. AMPLITUDE (S21) AND
GROUP DELAY (DLY) RESPONSES



22B. INPUT VSWR (V11) AND
RETURN LOSS (S11) RESPONSES

FIGURE 22. COMPUTED AMPLITUDE, GROUP DELAY, VSWR, AND RETURN LOSS RESPONSES OF A QUARTER-WAVE CAVITY WITH CAPACITIVE OUTPUT COUPLING CIRCUIT

Magnetic Output Coupling Loop

Figure 23 shows the simplified diagram showing the construction of a magnetic output coupling loop circuit in a folded half-wave coaxial cavity with a secondary re-entrant line section [5,9]. The tube's output capacitance is brought to resonance by the inductive component of the transmission line that is physically less than a half-wave length long. The coarse output tuning is done by presetting the depth of the secondary line into the cavity. The PA output fine tuning is accomplished by varying the physical length of a flexible extension (bellows) on the end of the secondary transmission line. This type of output circuit does not require a plate blocking capacitor in the anode. The output power is coupled to the load by means of a magnetic loop positioned near the RF voltage null point located near the center of the primary transmission line where the RF current is maximum.

The coupling to the cavity varies as the square of the effective loop area and inversely as the square of the distance of the loop center from the cavity center axis. There is a minimal interaction between the tuning and loading adjustments in this type of power amplifier cavity design.

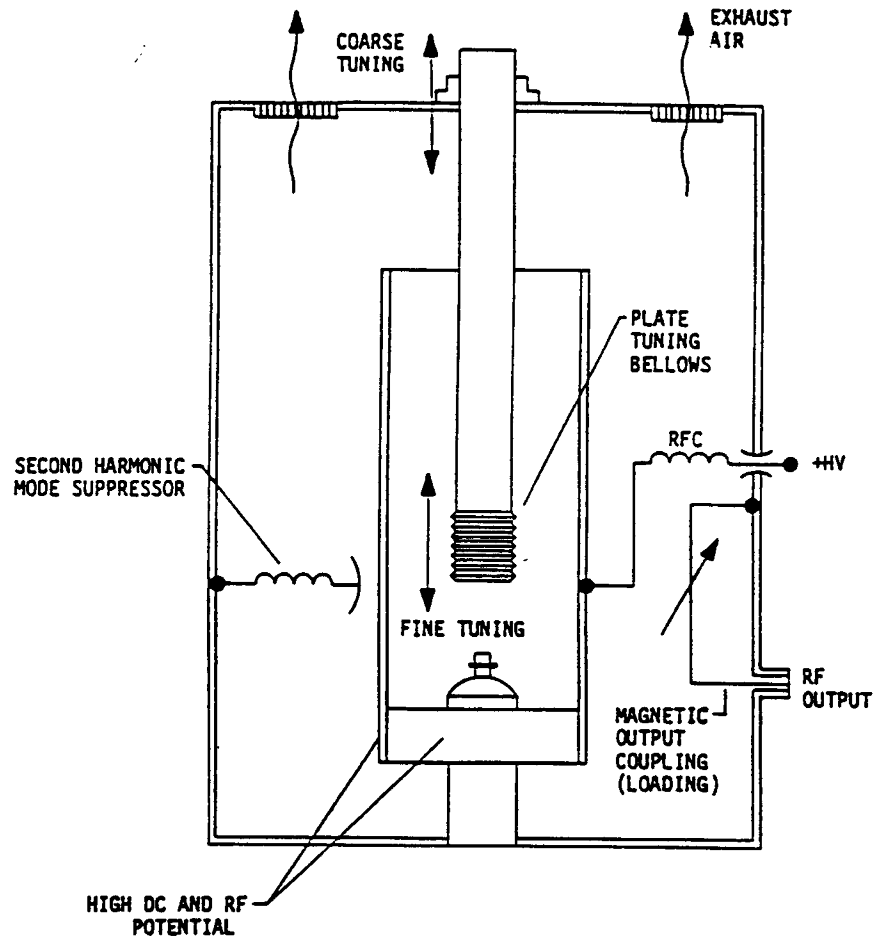
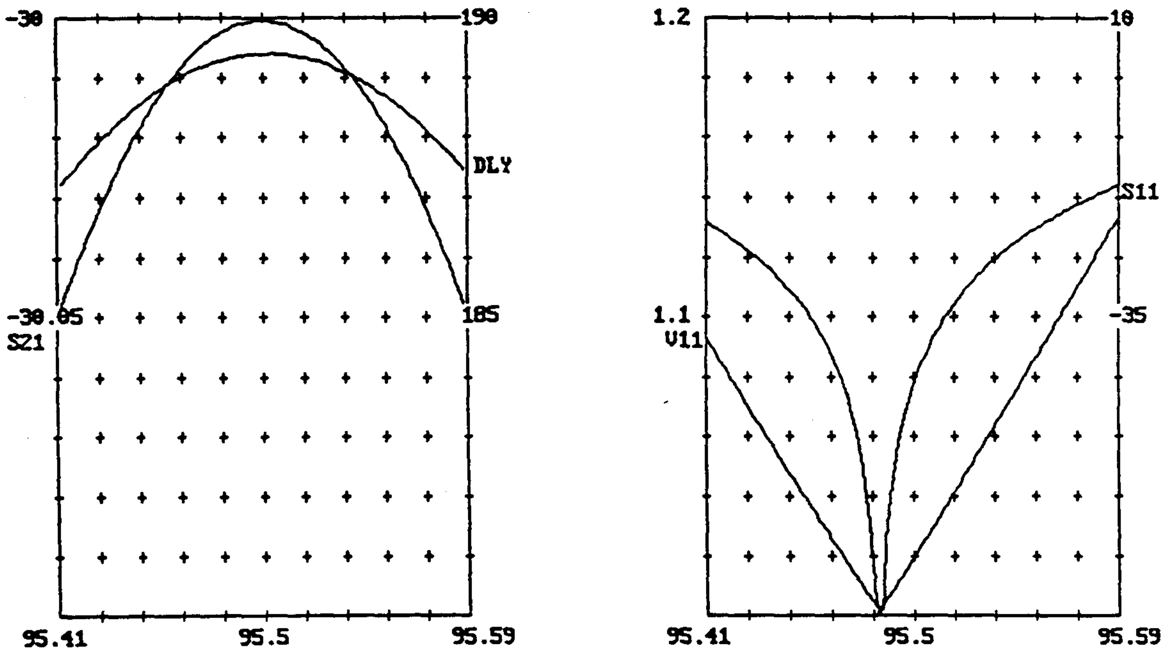


FIGURE 23. FOLDED HALF-WAVE COAXIAL CAVITY WITH INDUCTIVE TUNING AND MAGNETIC OUTPUT COUPLING LOOP

Furthermore, this type of cavity has a greater turn-around-loss due to its higher selectivity and provides better protection against RF interference. An additional 10 dB protection is available for an intermodulation product due to an interfering signal frequency separated from the carrier by 4 MHz.

A magnetic output coupling loop circuit was designed to transform the high impedance of the PA output tuned circuit (with total capacitance of 20 picofarads and 1590 Ohms load resistance across it) to 50 Ohms. This illustrates the output circuit of a typical 20 kW FM transmitter PA utilizing an Eimac 8989 tetrode. Computer software programs ECA [11] and =SuperStar= [12] were used to analyze the circuit design and obtain the response plots.

Figure 24A shows the computed amplitude response (S21) and group delay (DLY) response plots. Figure 24B shows the output VSWR (V11) and output return loss (S11) plots at -0.05 dB response points. The peaks of amplitude and group delay plots coincide in this case providing a symmetrical group delay response.



24A. AMPLITUDE (S21) AND GROUP DELAY (DLY) RESPONSES

24B. INPUT VSWR (V11) AND RETURN LOSS (S11) RESPONSES

FIGURE 24. COMPUTED AMPLITUDE, GROUP DELAY, VSWR, AND RETURN LOSS RESPONSES OF A FOLDED HALF-WAVE CAVITY WITH INDUCTIVE TUNING AND MAGNETIC OUTPUT COUPLING LOOP

MODULATION PERFORMANCE DATA OF A TYPICAL 20 KW FM TRANSMITTER

The following is the measured modulation performance data of Broadcast Electronics Model FM-20B 20 kW FM transmitter taken at the input and output circuits of the power amplifier. Three RF signal samples, one each from Broadcast Electronics Model FX-50 exciter output, the FM-20B PA tube grid ring, and the FM-20B PA output were taken to measure the modulation performance and to see the effects of input and output circuits on the transmitter performance.

The FM-20B transmitter was used to illustrate an example of a real world tube power amplifier, designed for optimum performance. The power amplifier uses an Eimac 8989/ 4CX12,000A high gain tetrode tube in the folded half-wave cavity output circuit with magnetic coupling loop. The FM-20B also uses a broadband L-C matching circuit at the tube grid input to minimize signal degradation.

An RF sample from either the exciter output, or the PA tube grid ring, or the PA output was connected to the FMM-2. The deemphasized audio and wideband composite outputs were used for composite tests. The composite baseband was used to drive the FMS-2. The decoded left and right outputs of the FMS-2 were used for the stereo performance tests.

Figure 25 shows the test setup used to measure the modulation performance of the FM-20B transmitter. The Audio Precision System One test set audio oscillator fed either the FS-30 stereo generator for stereo performance tests, or the FX-50 composite input directly for baseband composite performance tests.

Figure 26 shows the MVDS (Microprocessor Video Diagnostics System) screen printout of FM-20B transmitter meter readings operating at 20 kW RF output power.

Measured Bandwidth Of FM-20B Transmitter

Figure 27 shows a plot of the measured overall amplitude response, group delay response and input return loss response of FM-20B transmitter. It shows a -3 dB bandwidth of 1.45 MHz.

FM-20B MODULATION PERFORMANCE TEST SETUP

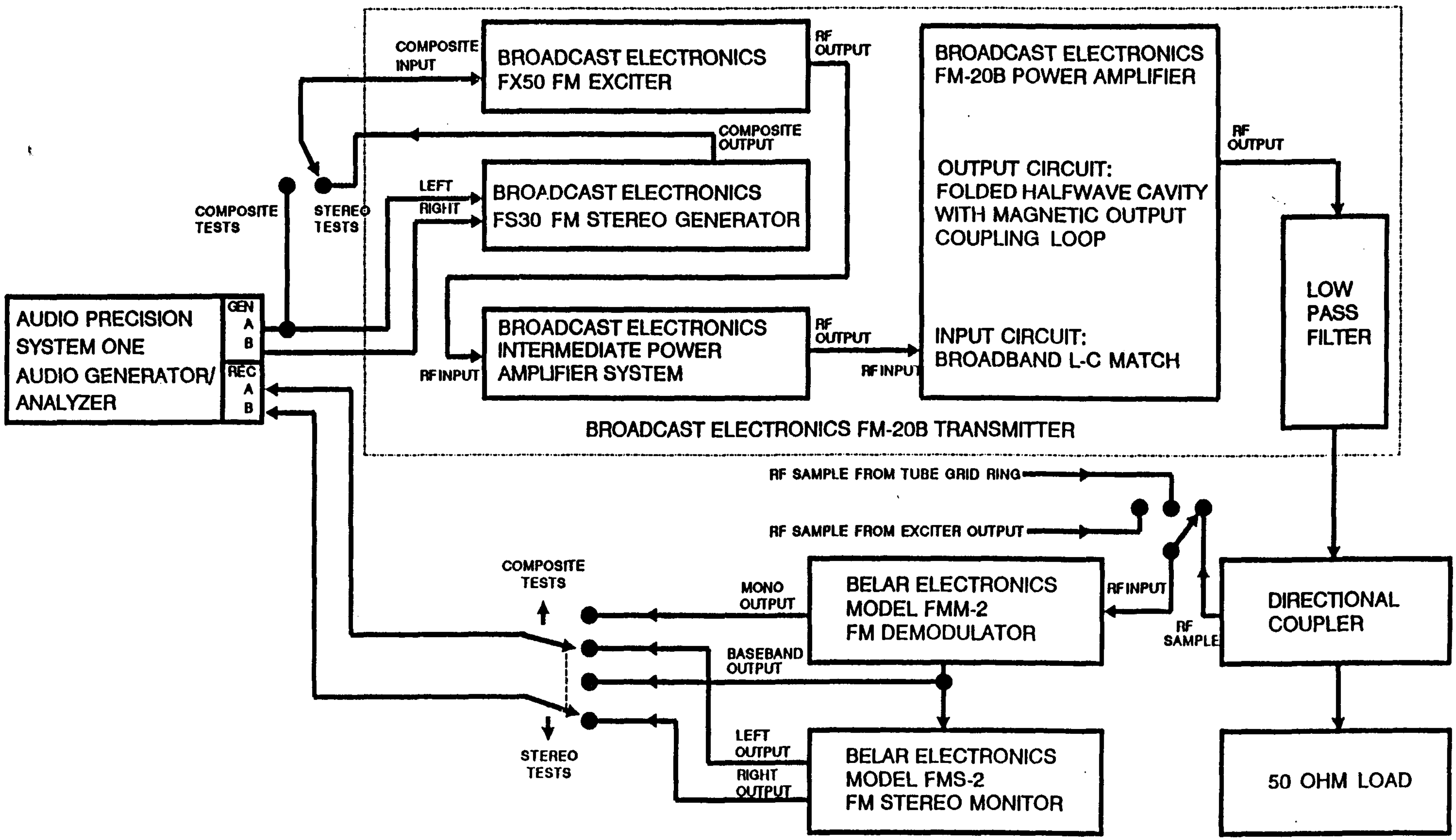


FIGURE 25. FM-20B MODULATION PERFORMANCE TEST SETUP

```
     FM-20B XMTR TEST 95.5 MHZ AT 20KW                      TUE 09/05/89    09:45:37AM
     BROADCAST ELECTRONICS INCORPORATED    MODEL FM-20B       S/N 0000    VR2.2
------------------------------------------------------------------------------------
CONDITION: NORMAL OPERATION
------------------------------------------------------------------------------------
POWER AMPLIFIER (PA)    EFFICIENCY= 80%    | TRANSMITTER POWER OUTPUT
                                           |
                  PLATE    SCREEN    GRID  | AUTHORIZED 20.00KW=100%       0.0KW ERP
VOLTAGE          8.94KV     710V    -298V  | ACTUAL     20.00KW=100%       0.0KW ERP
CURRENT           2.81A    149MA     31MA  |
POWER OUTPUT    20.00KW                    | REFLECTED   0.00KW=  0%
DISSIPATION      5.12KW     106W           | VSWR        1.0:1
------------------------------------------------------------------------------------
INTERMEDIATE POWER AMPLIFIER (IPA)         |
                     -1-       -2-         |
VOLTAGE             27.9V     28.0V        |
CURRENT             11.4A     11.3A        |
FORWARD POWER        204W      203W        |
REFLECTED POWER        2W        1W        |
DISSIPATION          114W      113W        |
TOTAL POWER        FWD= 367W  RFL=    3W   |
-------------------------------------------| EXHAUST AIR TEMP=  46 C
EXCITER FORWARD POWER        36W           | APC ON
EXCITER REFLECTED POWER       1W           |
```


FIGURE 26. MVDS SCREEN PRINTOUT SHOWING METER READINGS OF FM-20B TRANSMITTER

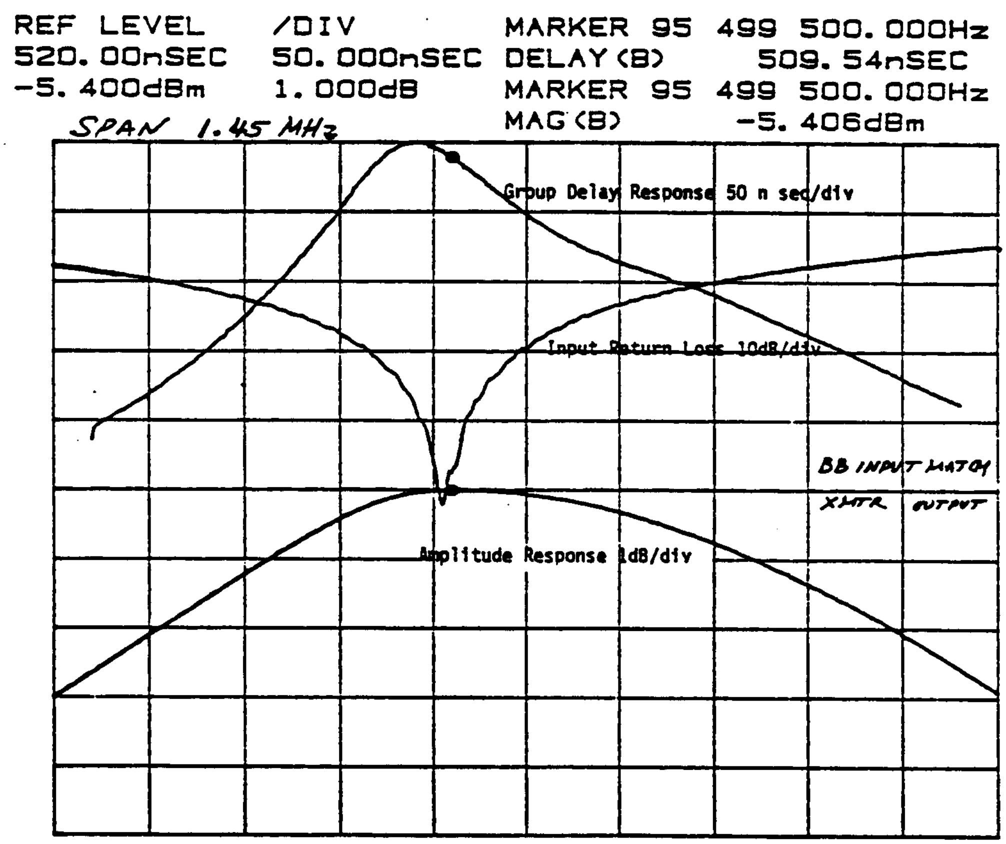


FIGURE 27. MEASURED AMPLITUDE, GROUP DELAY, AND INPUT RETURN LOSS RESPONSES OF FM-20B TRANSMITTER

FM-20B Synchronous AM

Figure 28 shows the measured synchronous AM (amplitude modulation) performance of the transmitter. The performance degradation is 11 dB from the exciter to the tube grid ring and 2 dB from the grid ring to the transmitter output.

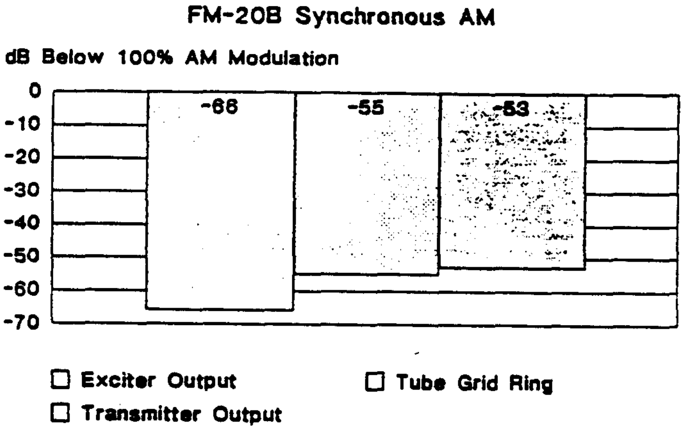


FIGURE 28. FM-20B SYNCHRONOUS AM

FM-20B Composite Amplitude Response At 100 kHz

Figure 29 shows the measured composite amplitude response performance degradation of 0.08 dB from the exciter to the tube grid ring and a further 0.08 dB from the grid ring to the transmitter output.

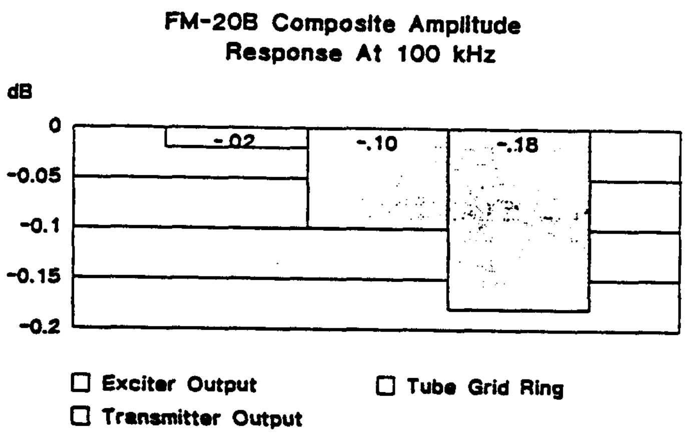


FIGURE 29. FM-20B COMPOSITE AMPLITUDE RESPONSE AT 100 kHz

FM-20B Composite THD+N At 15 kHz

Figure 30 shows an increase in THD+N at 15 kHz of 0.006% from the exciter to the tube grid ring and 0.052% from the grid to the transmitter output.

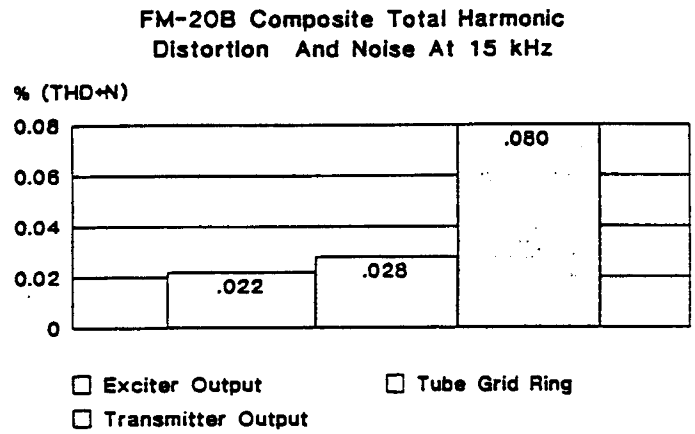


FIGURE 30. FM-20B COMPOSITE TOTAL HARMONIC DISTORTION AND NOISE AT 15 kHz

FM-20B Composite THD+N At 400 Hz

Figure 31 shows 0.001% increase in THD+N at 400 Hz from the exciter to the tube grid ring and no performance degradation from the grid ring to the transmitter output.

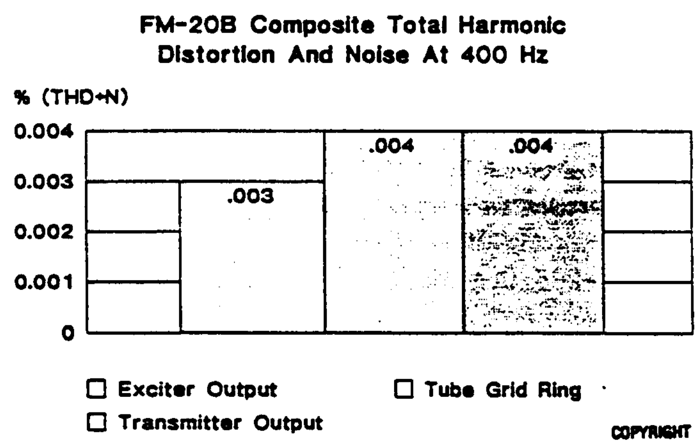


FIGURE 31. FM-20B COMPOSITE TOTAL HARMONIC DISTORTION AND NOISE AT 400 Hz

FM-20B Composite SMPTE IMD (60Hz/7kHz 1:1 Ratio)

Figure 32 shows the performance degradation of 0.002% IMD from the exciter to the tube grid ring but no change from the grid ring to the transmitter output.

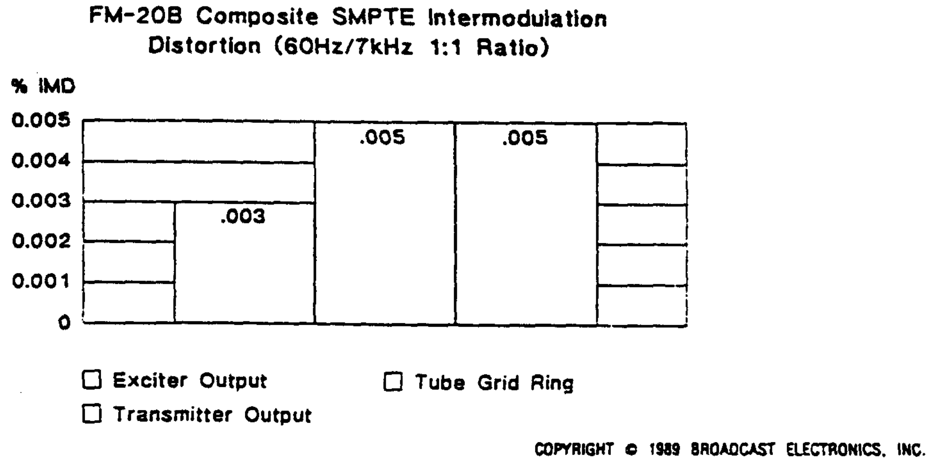


FIGURE 32. FM-20B COMPOSITE SMPTE INTERMODULATION DISTORTION (60 Hz/7 kHz 1:1 RATIO)

FM-20B Composite FM Signal-To-Noise Ratio

Figure 33 shows 1 dB drop in FM Signal-to-Noise ratio from the exciter to the tube grid ring and no degradation from the grid ring to the transmitter output. The small degradation may be caused by the changes in the input circuit and output circuit reactance due to mechanical vibration.

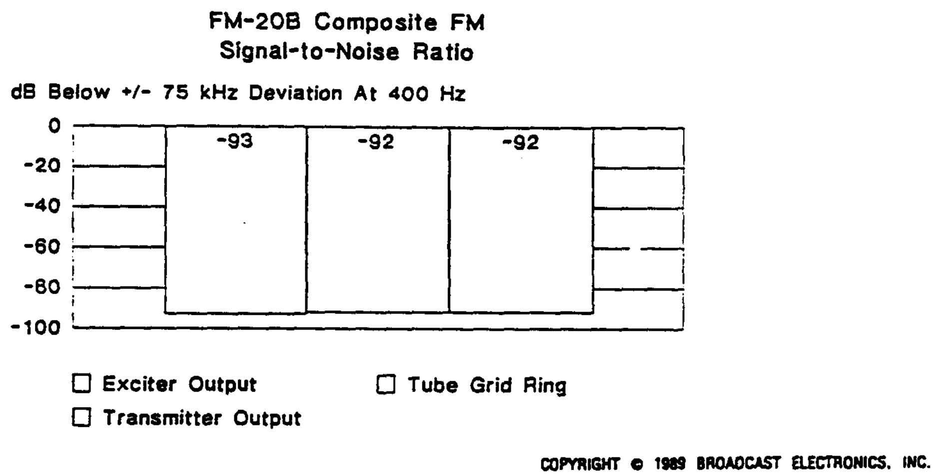


FIGURE 33. FM-20B COMPOSITE FM SIGNAL-TO-NOISE RATIO

FM-20B Stereo THD+N At 15 kHz

Figure 34 shows an increase in THD+N at 15 kHz of 0.011% from the exciter to the tube grid ring and 0.030% from the grid to the transmitter output.

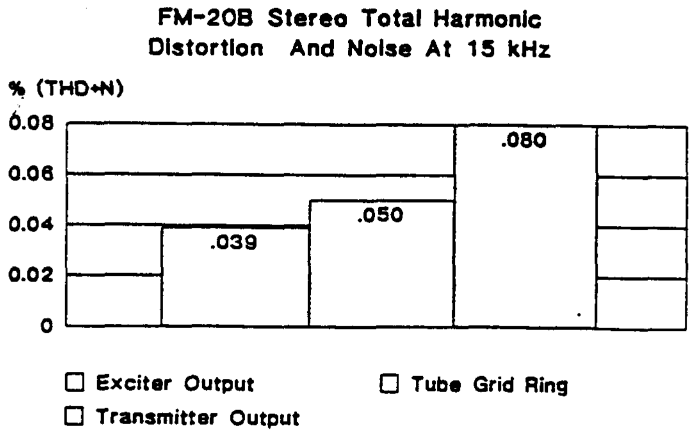


FIGURE 34. FM-20B STEREO TOTAL HARMONIC DISTORTION AND NOISE AT 15 kHz

FM-20B Stereo THD+N At 400 Hz

Figure 35 shows no performance degradation of THD+N at 400 Hz from the exciter to the tube grid ring as well as from the grid ring to the transmitter output.

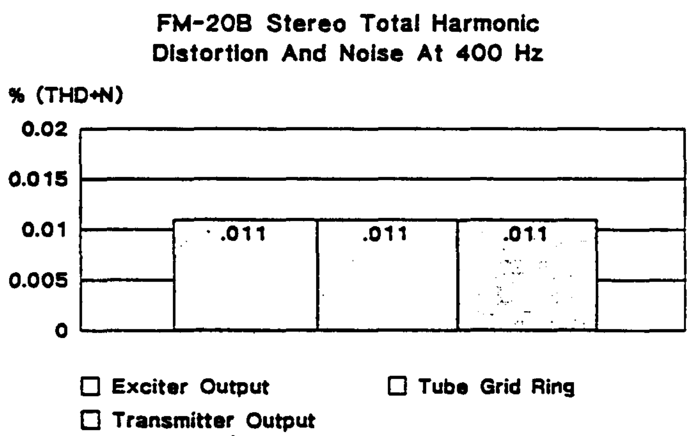


FIGURE 35. FM-20B STEREO TOTAL HARMONIC DISTORTION AND NOISE AT 400 Hz

FM-20B Stereo SMPTE IMD (60 Hz/7 kHz 4:1 Ratio)

Figure 36 shows the performance degradation of 0.012% stereo IMD from the exciter to the tube grid ring. There is actually an improvement of 0.009% stereo IMD performance from the grid ring to the transmitter output which may be caused by partial cancellation of distortion products in the left and right channels.

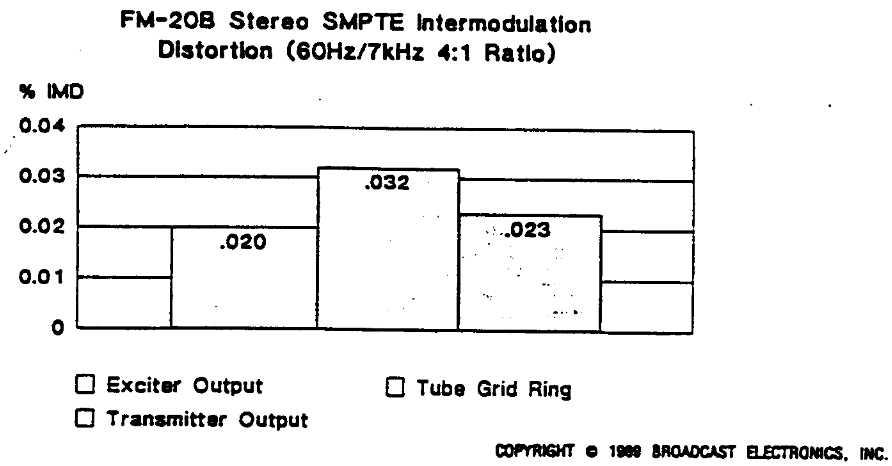


FIGURE 36. FM-20B STEREO SMPTE INTERMODULATION DISTORTION (60 Hz/7 kHz 4:1 RATIO)

FM-20B Stereo Separation At 15 kHz

Figure 37 shows a 6 dB drop in stereo separation from the exciter to the tube grid ring and a further 3 dB drop from the grid ring to the transmitter output.

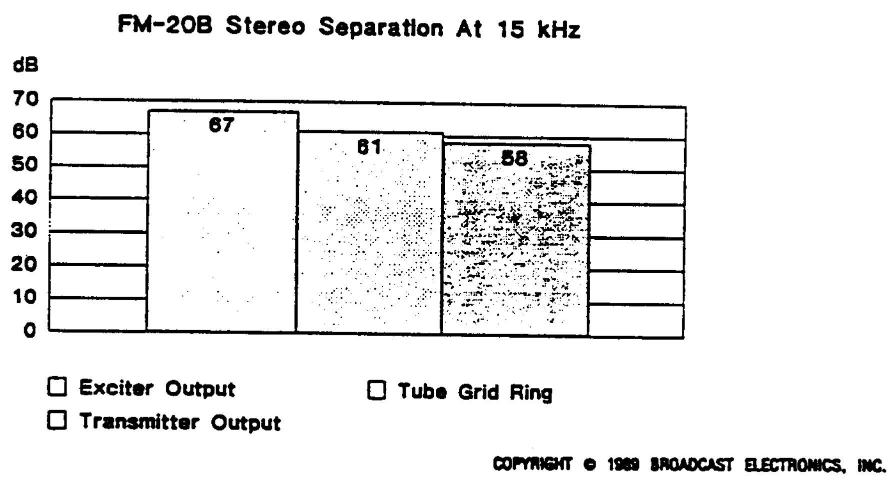


FIGURE 37. FM-20B STEREO SEPARATION AT 15 kHz

FM-20B Stereo Separation At 400 Hz

Figure 38 shows a 6 dB drop in stereo separation from the exciter to the tube grid ring and a further 3 dB drop from the grid ring to the transmitter output.

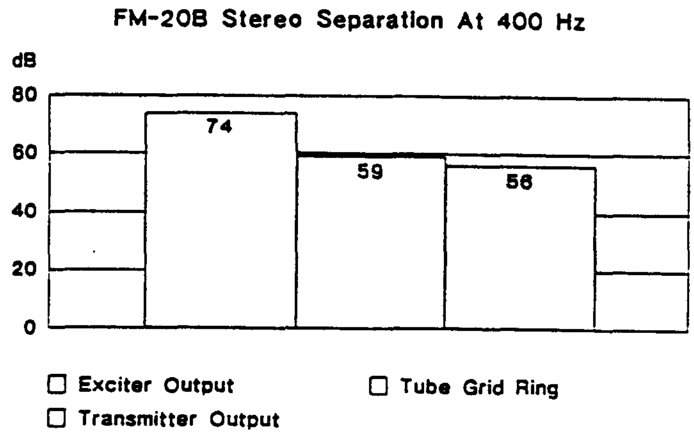


FIGURE 38. FM-20B STEREO SEPARATION AT 400 Hz

CONCLUSIONS

The design of tube power amplifiers for optimum FM transmitter performance requires careful considerations in the choice of input and output circuits due to their effects on the transmitter bandwidth and group delay. The conclusions reached are as follows:

1. RF bandwidth affects audio performance. It is, therefore, necessary to minimize bandwidth limiting components in the RF path to reduce performance degradation.

2. Good engineering judgement is called for to balance the tradeoffs between bandwidth and immunity to RF intermodulation. A bandwidth of 1.0 to 1.5 MHz seems adequate for excellent modulation performance while providing a reasonable degree of immunity to RF intermodulation.

3. Topology of input and output circuits has a combined effect on the power amplifier bandwidth. Certain topologies such as broadband input matching circuit and magnetic output coupling loop result in better overall transmitter performance.

4. Equipment manufacturers may provide information on the amplitude and the group delay responses of transmitters. This would allow the broadcast system engineering design to be tailored to particular needs. It would also create opportunities for designing delay equalization networks in the system to compensate for transmitter caused group delays as well as delays caused by filterplexer and combining system.

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Mr. Shrestha's practical experience involved engineering, operations, and management work as director of engineering for the National Radio Broadcasting Network of Nepal. His earlier experience includes several years of engineering and management work in broadcasting, as well as aeronautical communications and navigational aid equipment.

The author holds a U.S. Patent for electronic design utilized in broadcast equipment and is a member of the Institute of Electrical and Electronics Engineers. He is also a member of Tau Beta Pi and Phi Kappa Phi honor societies.

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