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Broadcast Transmission Systems

at

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PART 7 - AUDIO; BROADCAST; RECEIVERS

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VIDEO MODULATION LIMITER

L. S. Sadler
Television Station WMTV
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ABSTRACT

A simple video modulation limiter is described which can be built at low cost by making use of equipment already in use at most television stations. Modulation can easily be held at 95 per cent to 100 per cent with inputs varying over a range greater than 20 db. Transient effects are negligible, and the unit is suitable for use with network color. The work of the video operator is greatly simplified while a brighter, more uniform picture is put on the air.

COLOR TV RECORDING ON MAGNETIC TAPE

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ABSTRACT

It has been established that a practical method of recording monochrome tv signals on magnetic tape is the technique of high speed transverse scanning of a 2-inch-wide tape moving at 15 inches per second. Using this same recording technique for color tv signals presents some interesting and challenging problems. This paper describes some of the precise signal handling techniques and special equipment that had to be developed in order to build a broadcast quality video tape recorder good for both color and monochrome.

AN AUTOMATIC LEVEL CONTROL USING VERTICAL INTERVAL TEST SIGNALS

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Summary

The introduction of a peak white reference bar in the vertical interval of television signals by three major networks has pointed up the importance of this heretofore unused time. The transmission of such information provides all concerned with an easily visible amplitude reference, which may be viewed on either horizontal or vertical oscilloscope presentation. Such information in the signal permits manual adjustments to be made at various control points to assure proper levels throughout the system.

A new device is described, which automatically responds to any amplitude variation in this reference white signal. A servo system is used to automatically compensate for these variations. Such a device provides ALC (Automatic Level Control) which eliminates the necessity for manual adjustment thereby reducing brightness variations and corresponding distortions.

Introduction

During the past few years, numerous suggestions have been made to transmit some kind of data simultaneously with a television program. This data in its present form may be used for checking, observation and control of the characteristics of a television network or facility. Such observations have been made in the past during non-program times, generally by use of several standardized test signals such as the multiburst, stairstep and window.

The simultaneous transmission of test and program signals without conflict or interference is accomplished by using the end of the vertical blanking period to send the additional information. This simultaneous transmission has many advantages. Some of these are:

1. A peak white reference is always present. This permits the most important level of a video signal to be established.

2. Deterioration or potential deterioration of video facilities is instantly indicated and corrective measures may be undertaken during program time.

3. The behavior of video facilities under dynamic signal changes such as varying average picture level is instantly shown.

4. Color or monochrome signals from different studios, cameras or encoders can be adjusted for the same operating conditions.

History

The first attempt to use the vertical interval for transmission of additional information was reported in 1943 by Dr. P. Goldmark and associates in connection with the CBS color system. Information was sent during vertical time to synchronize the color wheel of the home receiver.

In 1953, one of the authors of the present paper proposed to the FCC an extensive color test signal which would be sent both at the beginning and end of the vertical interval. This proposal was reported in the technical press and also demonstrated in that year. Mr. E. W. Chapin and W.K. Roberts of the FCC Field Engineering Laboratory investigated vertical interval signals in 1955 in conjunction with Telechrome.

Vertical Interval test and control signal transmission was reported in Germany in 1955. This signal included a stairstep, a sine-squared pulse and bar and routing, cueing and point of origin information.

NBC began experimenting with various test signals in 1956. At this time, the FCC authorized temporary transmission of vertical interval test signals by all broadcasters. ABC and CBS began transmitting their own test signals in 1957. Today all three networks are making extensive use of these vertical interval test signals and the EIA has set up a sub-committee BTS-5 to establish standards for a proposal to the FCC for permanent authorization.

Figure 1 shows the vertical blanking interval of one field of a received NBC program. Note the vertical interval signal occupies three lines at the end of vertical blanking.

Figure 2 shows an off air photograph of one line of an NBC vertical interval signal. The normally present subcarrier has been filtered out. A peak white reference is at the right of the line.

Figure 3 shows an off air photograph of one line of an ABC vertical interval signal. Note that a peak white reference is present at both extremes of the line.

Figure 4 shows an off air photograph of one line of CBS vertical interval signal. CBS alternates a stairstep and multiburst signal with the one shown. Also a peak white reference is transmitted at the right of the line.

Equipment Requirements

The main use of all the presently transmitted test signals is for visual observation and manual adjustment when correction is required. No automatic controls have been used to adjust levels, frequency response, envelope delay, differential gain or phase or any other characteristic.

The purpose of this paper is to describe equipment for automatically controlling the amplitude of a television program by sampling the amplitude of a peak white reference bar transmitted during vertical interval time. It is also our purpose to indicate other areas of automation to which the principles of this equipment may be readily extended.

In order for automatic level control equipment to be generally useful, it is necessary for some degree of standardization to be agreed upon. The BTS-5 sub-committee has agreed "that when reference signals are employed, the right hand edge of at least two lines in the region from 2 to 12 microseconds preceding the leading edge of horizontal sync, shall be reserved for amplitude reference signals, which signals shall be concluded by reference white signals of at least five microseconds duration."

Figure 5 shows this interval graphically. The design of the present equipment has been based on reference white signals as shown here.

Some of the questions considered in the design are for example: What range of level variation shall be corrected for? Whal shall be the speed of correction? What shall be the method of correction?

Level variations may result from three main causes. One cause is amplifier gain variation. This variation is generally insignificant except over long time periods. Another cause of level variation is fading during microwave transmission. Fades may range from negligible to a total outage and may last as long as several seconds. A third and possibly the most important cause of level variation is program and studio switching. Consideration of these causes made us decide to recommend a maximum correction range of $\pm$ 50% from a nominal operating level. Thus for a 1 volt video signal a variation down to 0.5 volts or up to 1.5 volts will be corrected for. Any variation in excess of this range will operate an alarm to call attention to the abnormal condition.

The rate of correction was chosen to be a relatively slow one. The response of the unit to a step function has a time constant of about 1 second. Rapid level variations are ignored. The narrow effective bandwidth given by this long time constant means that the device will operate with extremely noisy signals.

The correction method chosen was an electromechanical one, namely a servo motor controlling a low impedance potentiometer which varies the video level. Methods of video level correction have been described in the literature which derive their level information from the signal itself and use bias variation of vacuum tubes to effect video level variation. We believe an electromechanical system, as used in the present unit, is simpler. However, the principle of operation of the unit is equally applicable to an electronic gain control. In fact a suitable control voltage is provided as an output. An electromechanical system has the advantage that it is adaptable with only minor changes to control of other video characteristics.

Basicly then, the present unit samples the amplitude of the peak white bars of a vertical interval test signal. The amplitude of the white bar is transformed to a proportional dc voltage. This voltage is compared to a reference to develop a dc error signal. This error signal is converted to a 60 cps signal with proportional amplitude and sense. The amplified 60 cps error signal is applied to a servo motor which controls a video level potentiometer.

Block Diagram

Figure 6 shows a simplified Block Diagram of the unit. The diagram is divided into four functional blocks. Block 1 is the Video Section. The program signal is terminated by a 75 ohm potentiometer. This potentiometer is the means used to control the video level. The arm is normally set at the half voltage point. This permits correction at ± 50% of the nominal input to be easily obtained. The arm feeds a X2 amplifier which makes up for the initial attenuation of the potentiometer. The system thus has nominally unity gain. The X2 amplifier is a high quality device, suitable for color programs.

The video output is applied to Block 2, the Sampler Section. After passing thru an isolation amplifier, the signal is clamped at blanking level by a conventional keyed clamp circuit. Clamping is essential to maintain the position of the peak white reference bars on a tube grid base independent of the average picture level of the signal.

The Keyer Output of the Model 1008-A Vertical Interval Signal Keyer is applied to this block to drive a sampling gate generator. This generator produces three microsecond wide pulses which are times to occur during the white bars of the vertical interval signal. The minimum width of the white bars was designated by BTS-5 as five microseconds. The three microsecond wide gate permits some leeway in accuracy of adjustment and drift of either gate or white bar before lack of coincidence occurs. Smaller gate widths reduce the sensitivity and the signal to noise ratio. It should be mentioned that the equipment is designed to sample two white bars occurring in successive lines. This is the minimum number of lines of vertical interval signal which would be added to the program. The amplitude sampler is a gate circuit which produces a pulse whose width corresponds to the sampling gate and whose amplitude is proportional to the white bar of the program signal.

The output of the sampler is applied to Block 3, the DC Section. The function of this section is to convert the sampler pulses into a proportional dc voltage. This could be done by an integrating network. However the duration of the pulse is only 6 microseconds with a repetition interval of 17,000 microseconds and only a minute dc voltage would be obtained. It was found desirable to use a peak detector type of circuit to obtain a reasonably high output voltage with the inherently small duty cycle of the input. The dc voltage is applied to a cathode follower and brought out as an external signal. This can be used to operate a recording voltmeter to obtain a permanent record of video level. This record may be obtained even though the unit is not used for automatic correction. This dc error signal could be used if desired to operate an electronic gain control. However as discussed previously, an electromechanical approach was preferred. Therefore the dc voltage from the peak detector is applied to Block 4, the 60 cps Section.

This section contains a DC to 60 cps converter which compares the dc signal with a dc reference. Deviation of the dc signal from the reference produces a 60 cps error signal whose amplitude is proportional to the deviation and whose phase is 0° or 180° depending on the sense of the deviation. The 60 cps error signal is phase adjusted, amplified and applied to the control field of a 2 phase servo motor. This motor through a gear assembly controls the video level potentiometer discussed in Block 1. The gear assembly reduces the maximum speed of the potentiometer to about 5 RPM. This gives a correction rate for large errors of approximately one DB per second.

The 60 cps amplifier also drives a signal level alarm circuit. This circuit has several functions. The most important function is to shut off the correcting motor if the program signal disappears completely. If the motor were not shut off, it would attempt to make maximum correction. When the program reappears a large spurious correction would exist. The alarm circuit prevents this from happening by shutting off the motor. The circuit is self-resetting, so that when the program reappears, the alarm goes off and automatic correction begins again.

Another important function of this circuit is that it can be used to detect the presence of vertical interval signals. If a vertical interval signal is present, the unit will correct. If it is not present no correction will take place and the program will automatically bypass the unit.

This circuit can also act as an automatic local-remote vertical interval signal switcher. For example if no remote vertical interval signal is present on an incoming program, normally a local vertical interval signal will be added.

If studio control switches to a remote program containing a vertical interval signal, the circuit under discussion will sense this immediately and switch the local one off. Whenever the remote vertical interval signal goes off the local one will automatically be switched back on.

Circuit Features

The following are some of the circuit features of this equipment. The 6 db gain video amplifier of Block 1 is shown in Figure 7. This amplifier is a high quality type which provides essentially distortionless amplification of color signals. It is a modification of a circuit originally described by Schroeder. The amplifier consists of a conventional amplifier direct coupled to a cathode follower. This in turn drives a "totem pole" type of output stage. No peaking coils are used. The amplifier is flat to 10 MC and has low frequency tilt of less than 1%. Differential gain and phase are less than 1% and 1° respectively.

Returning to the block diagram, Figure 6, the amplitude sampler is based on a 6AS6 gate circuit. The position of the sampling gate is controlled by a precision delay circuit. Figure 8 shows the sampling gate superposed on an off air vertical interval signal by means of a dual trace scope.

The output of the amplitude sampler goes to a peak detector circuit which converts the sample pulses to a proportional dc level. Several points of this circuit are worth consideration. The circuit is essentially a half wave rectifier. The top of Figure 9 shows a simplified circuit. The input voltage is shown as E_{in} with the pulse duration equal to T_1 and the interval between pulses equal to T_2. R_1 represents the total effective resistance in series with C when it is charging. This includes both the source and diode resistances. R_2 is the total effective resistance in series with C when it is discharging. It is required to know how much dc voltage will be obtained for a given peak input signal. The result shown on the slide states that the ratio of E_{dc} to E_{in} equals:

$$\frac{1}{1 + R1\,T2 \,/\, R2\,T1}$$

Note that C does not appear in the expression. It serves only to establish the ripple in the output and the response time to input voltage changes. For vertical interval application the ratio of T_2 to T_1 is very large, approximately 3000 in the present equipment. At the right of the slide, a curve is plotted showing the ratio of E_{dc} to E_{in} as a function of R_2/R_1 with T_2/T_1 equal to 3000. Note that to obtain a voltage ratio of 0.5, the ratio of R_2/R_1 must be 3000. If R_2 is 3 Megohms, then R_1 must be 1000 ohms. Without special circuitry, it is impractical to operate under conditions which give a true peak. However, operation on the straight line portion of the curve is satisfactory, provided gate width and circuit constants are stable.

The dc output is applied to a cathode follower whose cathode resistor is returned to a regulated negative voltage. A potentiometer forming part of this cathode resistor may be adjusted to provide a positive, negative or zero quiescent voltage output about which a dc error signal will vary as a function of program signal level.

Returning to Figure 6, the next block of interest is Block 4, the 60 cps Section. This section received a dc error voltage, converts is to a 60 cps error voltage and drives the servo motor. A circuit of interest is the DC to 60 cps converter shown in Figure 10. The basic circuit is tried and true but has been neglected to some extent in these modern days of choppers.

Here we have the dc error voltage applied to one grid of a twin triode tube. A dc reference is applied to the other grid. The cathode resistor is returned to ground through a 60 cps voltage, in our case the 6.3 volt heater supply. The plate circuit uses a push pull interstage transformer as an output. With no dc error the tube is balanced and equal in-phase 60 cps signals are developed at the two plates. These signals cancel and no output is developed. A dc error voltage will vary the transconductance of the left hand triode, causing more or less 60 cps plate signal to be developed. Cancellation will no longer take place and a 60 cps error voltage of one phase or the other will be developed across the transformer primary. The secondary is connected in a phase shifting circuit, permitting the phase of the control field of the servo motor to be properly adjusted.

Conclusions

The characteristics of the unit are such that at maximum sensitivity it will correct for ± 1% deviation in video level. Under this condition of maximum sensitivity a sudden large error will be corrected with an initial overshoot of about 15% and negligible secondary overshoot. A feedback tachometer can be added to the motor to eliminate overshoot in cases where it is found objectionable.

Figure 11 is a block diagram of the equipment used to demonstrate automatic level control. This diagram is typical of the normal connection in a broadcast studio. Here the program signal is an off air program coming from a TV receiver which has been modified to have a detector output and a video input. This detector output is bridged through the output of the 1005-A1 Test Signal Generator to permit a vertical interval test signal to be added. It is then bridged through the 1008-A Vertical Interval Signal Keyer to develop proper timing signals. The program is then terminated in the video level potentiometer of the 1009-AR Automatic Level Control. For purposes of demonstration another potentiometer has been introduced into the line which permits manual variation of program level into the 1009-AR unit. The level corrected output of the 1009-AR is bridged across the scope and terminated in the video portion of the receiver.

This equipment may be readily modified to provide automation of other transmission characteristics. Two changes are all that are required. One is a different sensing mechanism responding to the characteristic of interest, such as frequency response, for example. The other is a passive or active equalizer operated by a rotating shaft.

The Model 1009-AR Automatic Level Control Equipment extends automation into video networks. With this equipment, TV can be more efficient and provide better performance and greater economies in every day operation. This automatic level control device is a first step, made possible by vertical interval test signals, toward complete automation of other variables which affect video transmission quality.

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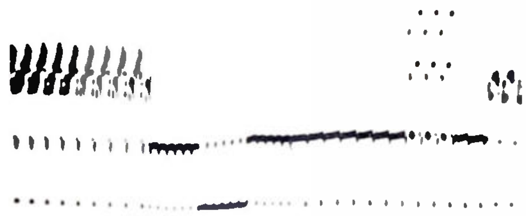

Fig. 1 Vertical blanking interval showing test signal.

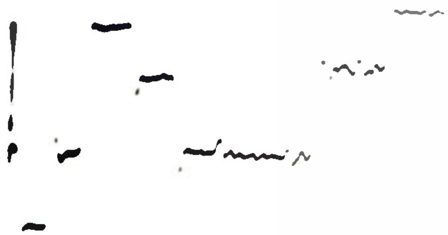

Fig. 2 NBC vertical-interval signal.

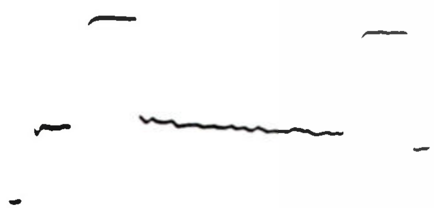

Fig. 3 ABC vertical-interval signal.

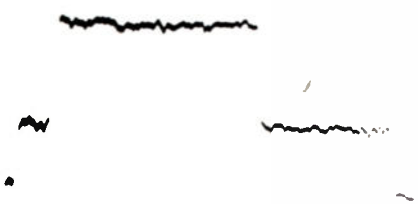

Fig. 4 CBS vertical-interval signal.

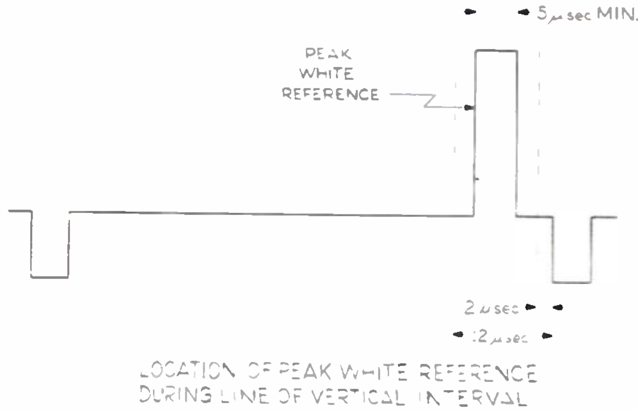

Fig. 5 Location of peak white reference during line.

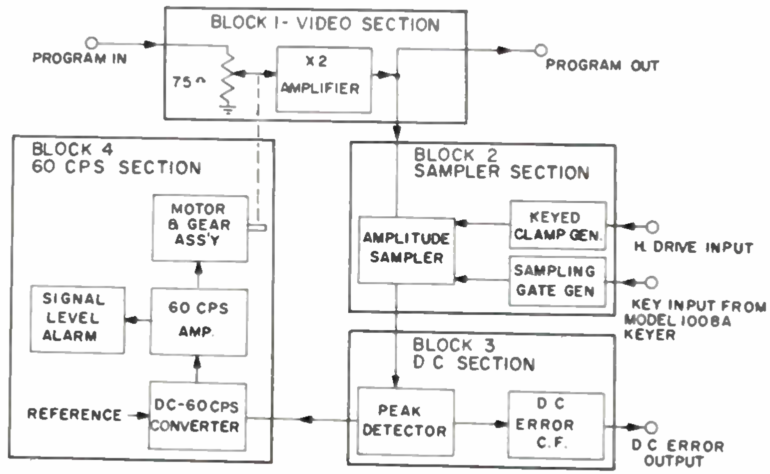

Fig. 6 Block Diagram, 1009-AR automatic level control.

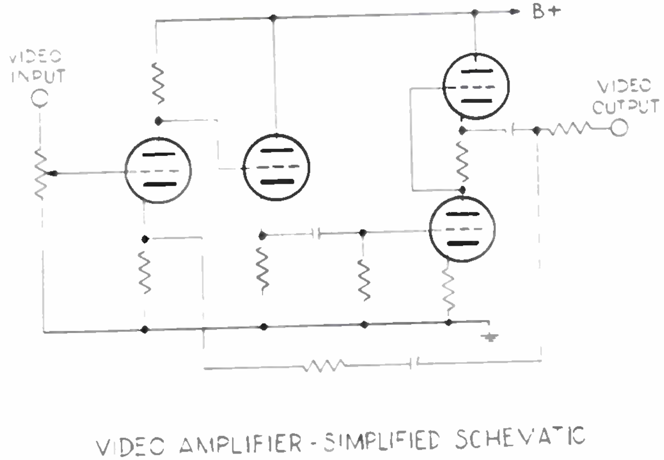

Fig. 7 **Simplified schematic, video amplifier.**

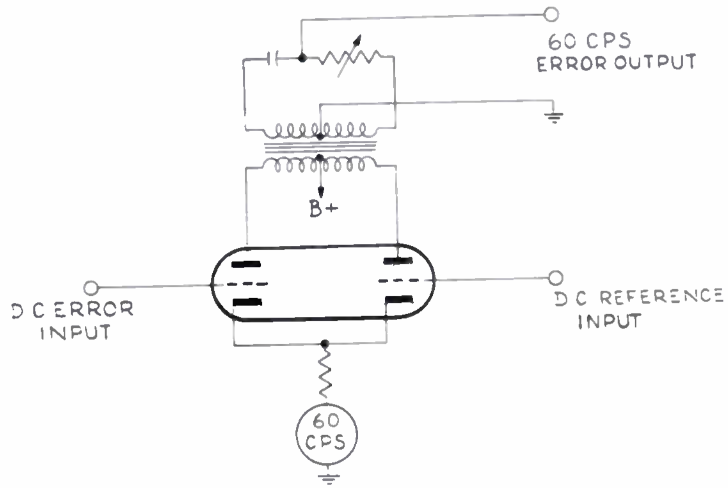

Fig. 10 **Simplified schematic, dc to 60 cps converter.**

Fig. 8 **Superposed display of test signal and sampler gate.**

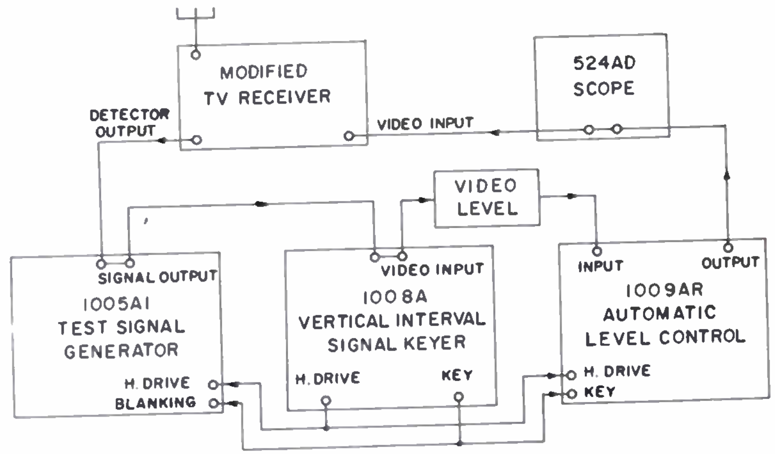

Fig. 11 **Block diagram, automatic level control demonstration.**

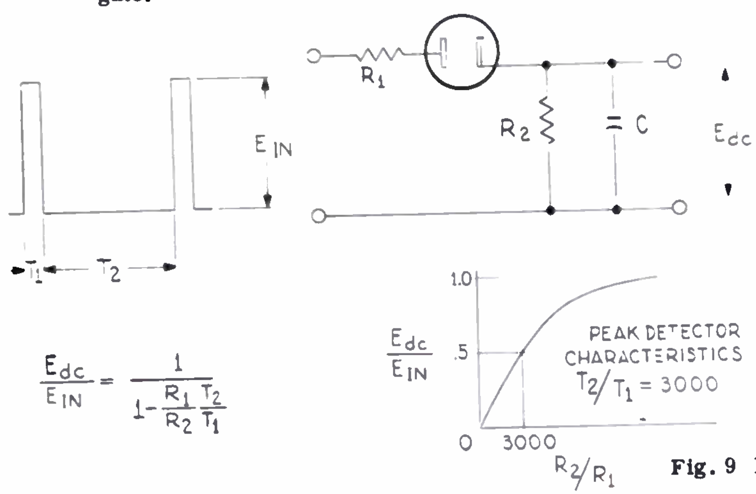

Fig. 9 **Peak detector circuit.**

REMOTE CONTROL

OF A

BROADCAST TRANSMITTER AND

DIRECTIONAL ANTENNA SYSTEM

By

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Triangle Publications, Inc.

During March of 1955, WFIL, The Triangle Publication Station in Philadelphia started to install an experimental remote control system for its 5,000 Watt R.C.A. BTA5F Transmitter and the associated three tower directional antenna system. WFIL transmitts directional on both its day and night patterns. The transmitter operation and monitoring was handled at the remote location by a first-class operator. During this experimental period, another first-class operator was on duty at the radio transmitter, and he kept the official watch and logs. The remote control point is at Roxborough, Pa., the site of WFIL's TV Transmitter: the radio transmitter is in Lafayette Hill, Pa., approximately a distance of 3.7 miles. The remote control equipment was manufactured by Rust Manufacturing Co., of Manchester, New Hampshire. In addition to the Rust equipment, there was installed at the remote point a General Radio Modulation Monitor; both of the units were fed from a Rust Receiver. Connections between the two sites were handled by leasing two 3,000 ohm Class "B" telephone lines.

At the remote point, we were able to monitor and adjust the following:

- Plate Voltage - On-off and measured
- Filament Voltage - On-off and measured
- Final Tank Current and Tuning
- Antenna Change Over Night to Day Pattern
- Antenna Common Point Current
- Antenna Phase Angles
- Antenna Current Ratios
- Tower Lights Current
- Tower Lights On and Off
- Overload Reset

At the remote point, it was possible to monitor the current or voltage while adjustments were being made to the circuit. On the control unit there was a meter, a meter calibrate switch, a meter read switch, a raise-lower switch, a telephone dial, and a series of tally lights to indicate the circuit being controled.. The meter reads the voltage or current under test, the raise-lower switch either operated latching relays to turn on or off various circuits or operated reversible motors which tuned various transmitter circuits. The telephone dial actuated a stepping switch in the control unit to light the proper tally light and produces the pulses used to drive the stepping switch in the transmitter unit. The transmitter contains a series of tally lights, a raise relay, a fail safe relay, a lower relay, a group of switching relays which connects the metering buss and the raise and lower relays to the proper circuits and a stepping switch which takes the pulses from the control unit and supplies potential to the proper switching relay. When no function is being dialed, there is a 13 volt potential on the control line: if this line fails and the voltage is removed, for more than 2 seconds, the fail safe relay will open and turn off the transmitter.

The circuits to be monitored have their normal current or voltage changed to the proper D.C. level by a divider network for the voltage and a shunt for the current. R.F. is monitored by a pickup loop and a rectifier which provides D.C. to the transmitter unit. A remote control phase monitor was employed. This monitor consists of two balanced amplifiers whose inputs are added vectorally and whose output is fed into a vacuum tube voltmeter. All selecting, balancing, and calabrating are handled by built-in relays. Phase relation indications are independent of modulation.

During part of the experimental period, a Minneapolis Honeywell "Strip Chart Recorder" was employed to automatically record final plate voltage and current, common point current, and current ratios between A-B Tower and A-C Tower. This unit recorded consecutively each of the readings at 30 second intervals. The readings were printed on a conventional strip chart. The sequential switching of the inputs was handled by the built-in timing device.

The operation of this unit is straightforward and relatively foolproof. Accuracy of the amplifier system is 1/5 of 1% of full scale deviction. The heart of the recording system is a "continuous balance amplifier". This amplifier is a servo amp. whose D.C. input is chopped and amplified. The amplified signal generates a 2 phase voltage which drives a motor. This motor is connected to a dial indicator or print wheel, being attached mechanically to the slide wire pot, will indicate the voltage which caused the original movement. When indicating ratios, the system essentially operates as a bridge with the servo amplifier across the null point. The motor in this case, drives one of the slide wire potentiometers to a balance point. When the balance is achieved, the indicator records the ratio diference between the two signals.

After the equipment was installed, WFIL was faced with a number of growing pains. These proved, for the most part, simple modifications of the original equipment to meed our particular conditions. Various relays had to be replaced as did meters to insure maximum reliability. Over a period of several months, work progressed to increase reliability of the equipment. The manufacturer made adjustments and substitution till finally everyone felt the equipment was operating satisfactorily.

During November and December of 1955, WFIL conducted and submitted to NARTB a series of tests and readings to substantiate the argument for allowing a station with a directional array to be operated by remote control. It is now a matter of record that the F. C. C. believes a station operating a directional antenna system may go remote, and they are accepting applications for this type of operation.

In WFIL's transmitting plane, there is a 5KW transmitter, a 1KW auxiliary transmitter and a 35KW standby engine-generator for power. Because of these facilities, we feel that the minimum requirements for a remote control system for a plant of this size is as follows:

1. Filaments, On and Off
2. Filaments line voltage measured
3. Plate voltage on and off
4. Plate voltage measured
5. Plate current measured
6. Final tank tuning
7. Antenna change over night-day pattern
8. Antenna common point measured
9. Low power switch
10. Overload reset
11. Measure antenna phase angles A-B, A-C
12. Measure antenna current ratios A-B, A-C
13. Tower lights on-off
14. Measure tower lighting current
15. Change over from main to auxiliary transmitter
16. Change over from regular to standby power source
17. Change over from regular to emergency audio channel

Based upon our experience and the experience of other stations, WFIL believes a transmitting plant with a directional antenna system can successfully be operated by remote control.

USE OF A SECTIONALIZED TV TOWER IN AM BROADCAST SERVICE

A. C. Goodnow
Westinghouse Broadcasting Company, Inc.
New York, New York

Summary

The 905 foot antenna structure of KYW-TV has been adapted to use as a two-element Franklin antenna at 1100 kc/s, the standard broadcast frequency of station KYW. The TV lines, FM line, and power conduit are insulated within the tower over portions of its length in such a manner as to function as coupling elements at 1100 kc/s. Insulated power transformers or phase monitor isolation coils are not required across the sectionalizing and base insulators, and the tower has a single driving point at the base. Construction details and tuning procedures are discussed; the results of performance measurements show that a horizontal field intensity of 301 mV/m for one kilowatt is realized.

Introduction

Over the years, considerable attention has been given to the problem of developing antenna designs yielding optimum values of horizontal radiation in the standard broadcast band. As a result of early investigations[1] pointing to the desirability of use of vertical radiators in the vicinity of one-half wavelength or slightly greater in height, and further work[2, 3, 4] emphasizing the allied problem of suppression of high-angle lobes of radiation, the 190 degree radiator (or a top-loaded equivalent) had during the thirties come into wide use in high power broadcasting stations as representing an economically feasible approach to an ideal antenna. With but few exceptions[5, 6], the use of towers of appreciably greater height was not undertaken because of the rapid increase in cost with increasing height.

Within the past decade, however, construction of towers to decidedly greater heights has become commonplace in the television broadcasting service. The possibility of adapting these taller structures to use as standard broadcast radiators also, where practicable, has aroused considerable interest, and methods of accomplishing this have been proposed[7, 8]. This paper describes the manner in which the KYW-TV tower in Cleveland, Ohio has recently been put into service as a radiator for KYW.

The FCC authorization for the KYW operation stipulates reduction of the field in northerly directions by approximately 4 db, in order not to increase the previously existing field in the service area of adjacent channel Canadian stations. The required pattern was obtained through use of a parasitically excited reflector in the form of a cable suspended from one of the guys of the tower. Reference to directional operation, however, will be made herein only where it has significance in considering performance of the main tower.

The KYW Antenna; Choice of Operating Mode

The KYW-TV tower is a uniform-section, triangular guyed tower, 805.5 feet in height and 6 feet in width between corner members. An FM antenna is side-mounted on the tower 753 feet above ground. Atop the tower is mounted a 6-bay superturnstile television antenna, increasing the over-all height of the structure to 905.5 feet. This height corresponds to 364 electrical degrees, or slightly in excess of a full wavelength at the KYW frequency (1100 kc/s).

With the radiator height thus established at approximately one wavelength, operation with the two halves in phase in the Franklin arrangement immediately suggests itself, although consideration of other possible modes of operation was not neglected.

In the quest for maximum values of horizontal radiation, many types of current distribution along vertical radiators have been proposed. Some of these are difficult to realize physically, and none seems to have any great advantage (in this particular application) over an approximate sine function distribution. Brown[4] points out that the horizontal field of a two-element full-wave Franklin antenna is the same as that of an antenna of the same height having a hypothetical uniform current distribution. The "optimum" distribution of La Paz and Miller[9] for this height (developed to determine the maximum possible horizontal radiation, without regard to whether the indicated distribution is realizable) yields as a theoretical limit a field only slightly higher than that of the Franklin.

These considerations, together with the results of experience with a similar arrangement in the three-fourths wave antenna at KDKA, made firm the decision to operate the KYW radiator as a two-element Franklin antenna, with approximately equal and in-phase currents in the two halves.

Details of Coupling and Isolation Sections

The tower is insulated at the center and the base, and insulated cross-bars at thirteen-foot intervals are provided up to the three-fourths wave elevation for supporting the TV and FM lines and a power conduit within the tower. Permanently installed current sampling loops are provided at the one-fourth and three-fourth wave levels, and the cables coupling these to the remote phase monitor are enclosed within the power conduit. The latter, since it may carry an appreciable amount of radio frequency current, is a copper conduit made up of the outer conductor only of the same type of coaxial line (3-1/8 inch) employed for the TV and FM circuits.

The unavoidable presence of TV and FM transmission lines within the tower at first appears to be an annoying complication in the consideration of means of driving the tower as an 1100 kc/s radiator. It soon becomes apparent, however, that in so disposing these lines as to permit the desired voltage and current relations to be developed on the tower, they can themselves be utilized to function as coupling elements between the sections, so that the tower need be driven only at the base. Moreover, by suitably fitting the power conduit into this arrangement, the need for insulated power transformers at the center and the base (along with sampling line isolation coils) is eliminated, and a feed system of gratifying simplicity results.

As shown schematically in Fig. 1 and in greater detail in Fig. 2, the TV lines, FM line, and power conduit are grounded at the base and insulated within the tower to a point one-fourth wavelength above ground. This is a well-known means of isolating such lines so that the tower may be series fed at the base. Both below and above the insulated half-wave level, the lines are similarly disposed as quarter-wave sections in a mechanical arrangement permitting take-off of power, sampling, TV and FM circuits at appropriate points along the tower. Inasmuch as the quarter-wave line sections so formed are connected with thier open-circuited ends across the center insulators, they act as a resonant circuit and thus are the means of reversing voltage phase and exciting the upper half of the tower.

The method of excitation bears a superficial resemblance to that of Wheeler's coaxial cable antenna, but the need for a metallic connection (for power and sampling circuits) between the lines and the lower tower element prevents it from functioning in the manner described by Wheeler[10]. Rather it is equivalent to the method shown by Smith and the Huttons[8] in their example of a 258 degree sectionalized top-loaded tower, although in the latter case TV and FM circuits were not considered and a steel cable within the tower functioned as a coupling section.

Tuning Procedure

It was anticipated that the mechanical length of the line sections would require reduction by means of shorting straps between the central lines and the tower structure at points in the vicinity of the one-fourth and three-fourths wavelength levels, principally because of the loading effect of insulator capacitance at the base and the center of the tower. In the process of adjustment, as the shorting straps were moved over a range of positions, measurements were made of base impedance, field intensity, and loop current ratio and phase (as indicated by the phase monitor). During the preliminary adjustments, equal active line lengths were maintained in the three sections involved. (The length of short-circuited section below the three-quarter wave point was maintained approximately 10 ft. greater than the other short-circuited portions with the intention of compensating for the lengthening effect of expansion loops formed in the central lines above the sectionalizing insulators.) Final 'trimming" adjustments were made by varying the length of active line section below the half-wave level only.

The measured values of loop current ratio and phase versus active length of coupling line sections are shown in Fig. 3. Inspection of the curves of this figure shows that as the active length of the coupling lines is varied between approximately 168 and 158 feet, the ratio of the loop currents in the upper and lower elements varies from 0.2 to 1.6, while the phase difference varies relatively slowly between approximately 40 and 115 degrees. Apparently over this range the reactance of the coupling sections passes through the point of resonance with the capacity of the tower gap, and it is in this region that the desired mode of tower excitation is approached. On either side of this range of adjustments, the tower gap is bridged by rapidly decreasing values of reactance, and a distinct reversal of phase in the currents of the two elements occurs; this condition is an approach to what should be expected were the tower gap non-existent. With the active length of coupling line sections at a little greater than 162 feet and approximately unity current ratio, the indicated phase lead of the upper element loop current is approximately 83 degrees. Although some rotation of phase is to be expected along the length of the tower, the measured amount is greater than anticipated; as discussed later on, this measurement is believed to be not representative of the phase relation over any considerable portion of the tower length.

A more detailed investigation of the distribution of current amplitude along the tower was made by use of an exploring loop and meter assembly. The measurements were made at a power level of approximately 10 KW. Two-way UHF communication sets were used to facilitate recording the measurements and de-energizing the transmitter while the tower team changed positions.

A set of measurements was made on each of the three tower legs at corresponding positions on the structural sections along the tower; the vertical interval between measuring points was 26 feet in most instances. The results of these measurements, with readings for the three tower legs averaged, are shown in Fig. 4. The discontinuities in the measured leg currents at elevations corresponding to the shorting strap levels are indicative of the functioning of the line sections; at these levels the tower legs begin to carry not only the "antenna" current but also "transmission line" current in their role as the outer conductor of the line sections. Although the transmission line current in the legs is balanced out in the far-field region by the out-of-phase current in the central conductors, it shows up in the loop measurements because of the closer proximity of the loop to the tower legs. The probable distribution of the antenna current component over pertinent tower sections is indicated by dashed lines in Fig. 4.

The measured distribution is a reasonable approximation of the sine function distribution shown by the dotted line in Fig. 4, although exhibiting somewhat more triangular maxima and broad minima; some reduction of velocity is apparent, particularly in the lower element.

Measured Performance

In the process of completing the adjustments and measurements requisite for putting the system into service, a logical work sequence could not always be followed because of prevailing conditions of high wind or icing which made tower work difficult and hazardous. A period of severe weather interrupted work on the tower after it has been tuned approximately to the condition shown by Fig. 4, but before the current ratio indicated by the phase monitor could be verified by measurements with the exploring loop. At this time, however, in the interest of completing the work as rapidly as possible, an extended set of non-directional field intensity measurements (in accordance with the procedure set forth in the FCC Rules and Regulations) was made forthwith. The measured RMS value of the essentially circular pattern resulting was 2070 mV/m, as compared to a value of 2150 mV/m calculated with a 5% allowance for expected system losses.

Subsequently, as a result of the distribution measurements along the tower, it was found that the discrepancy between the readings of the permanently installed sampling loops and the exploring loop measurements was approximately 27%. In the belief that the exploring loop measurements (involving a number of readings averaged over the three tower legs) are less subject to error than the remote readings from the sampling loops, a readjustment of the coupling section lengths was made, to yield the distribution shown in Fig. 4. The field intensity as monitored at a single check point showed no sensible change during this readjustment; the field intensity remains nearly constant on a broad maximum as the current ratio approaches unity. It is probable, however, that a slight improvement in the horizontal radiation was obtained, since a similar field survey made after the change but with the system operating directionally resulted in an RMS field value only 1% less than a similarly computed value. Accordingly, this is believed to reflect an improvement in the horizontal field of the tower itself that would result in an RMS value of 2130 mV/m in non-directional operation, corresponding to 301 mV/m for one kilowatt.

The apparent error in sampling loop ratio readings may be attributable to a difference in active areas of the loops or their relation to the tower members, although every effort was made to preserve symmetry in their mounting. As mentioned in Section IV above, the approximate quadrature phase relation in loop currents indicated by the phase monitor may also be questionable. Sources of error such as differences in lengths of sampling lines, instrumental inaccuracy etc., were eliminated by appropriate checks, and the fact that the sampling loops are located on portions of the tower which do not carry "transmission line" currents should eliminate any complication associated with effects of the coupling sections. On the other hand, a calculation of the RMS value of the horizontal field assuming quadrature phase difference throughout both elements yields a figure of 1780 mV/m, or 16% less than the value realized. It may therefore be concluded, giving due weight to the measured amplitude distribution of Fig. 4, that the currents are substantially in phase throughout the greater portion of the tower length.

An estimate of the base impedance to be expected was made by assuming that the calculated loop resistance of the upper section (100 ohms) could in this case of half-wave elements be considered to be transferred unmodified to the loop of the lower element and there added to the calculated loop resistance of the latter (127 ohms) the tower then being treated as a fictitious 182 degree radiator with the resultant loop resistance of 227 ohms. These values were used in Schelkunoff's modification[11] of the Siegel and Labus formulae, and yielded a calculated base impedance of 319-j264 ohms. The measured base impedance was 606-j50 ohms. The actual base impedance, however, may assume any value on a circular locus through the origin, depending on the selected value of the shunting reactance of the lower isolation section. The calculated value, if shunted by a reactance of $+j650$ ohms, for example, becomes $537+j0$ ohms, a reasonably close approximation of the measured value. It is evident, therefore, that by some device similar to that described, the base impedance in application such as this can be predicted closely enough for system design purposes.

Operational Advantages

The operational advantages realized through adaptation of the KYW-TV tower to AM service may be summarized as follows:

1. An increase in horizontal field intensity was obtained equivalent to that corresponding approximately to a 50% increase in power with a conventional antenna.

2. For non-directional operation in Cleveland, an increase of approximately 26% in area and 27% in population served within the primary service area would result.

3. For the actual directional operation authorized in Cleveland, an increase of approximately 34% in area and 53% in population served within the primary service area was realized.

Acknowledgements

Throughout the planning and completion of the project, the cooperation of the KYW engineering staff, in particular Messrs. R. J. Plaisted and A. H. Butler, was invaluable.

Portions of this material have been included in Applications and Proof of Performance submitted to the Federal Communications Commission.

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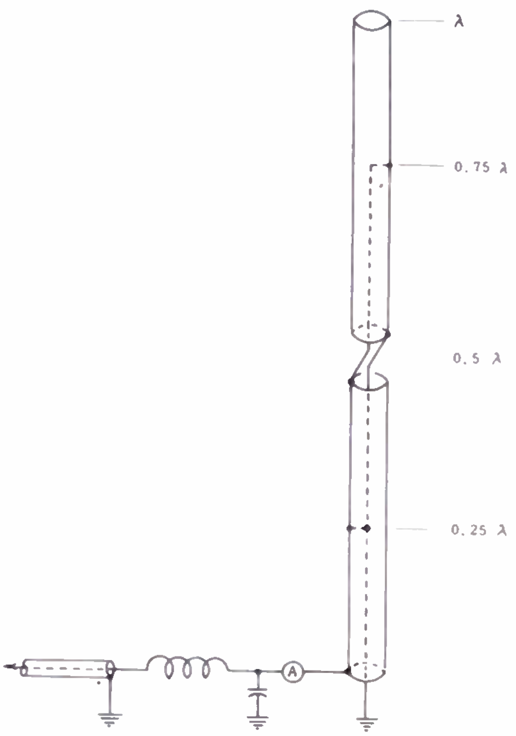

Fig. 1 **Schematic diagram of antenna and coupling and isolation sections.**

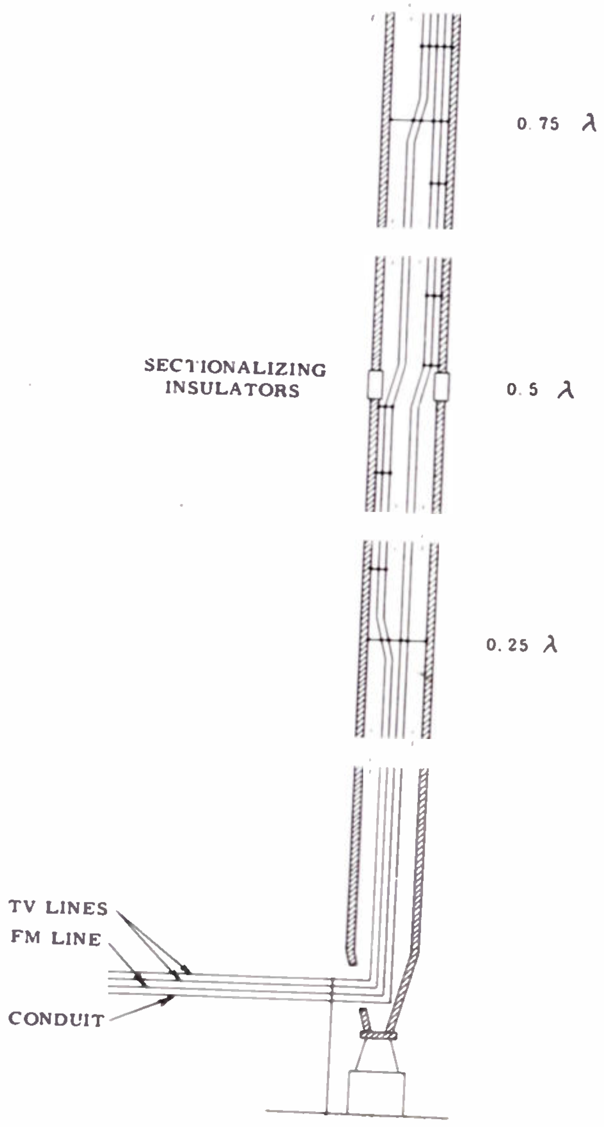

Fig. 2 **Detailed schematic of coupling and isolation line sections.**

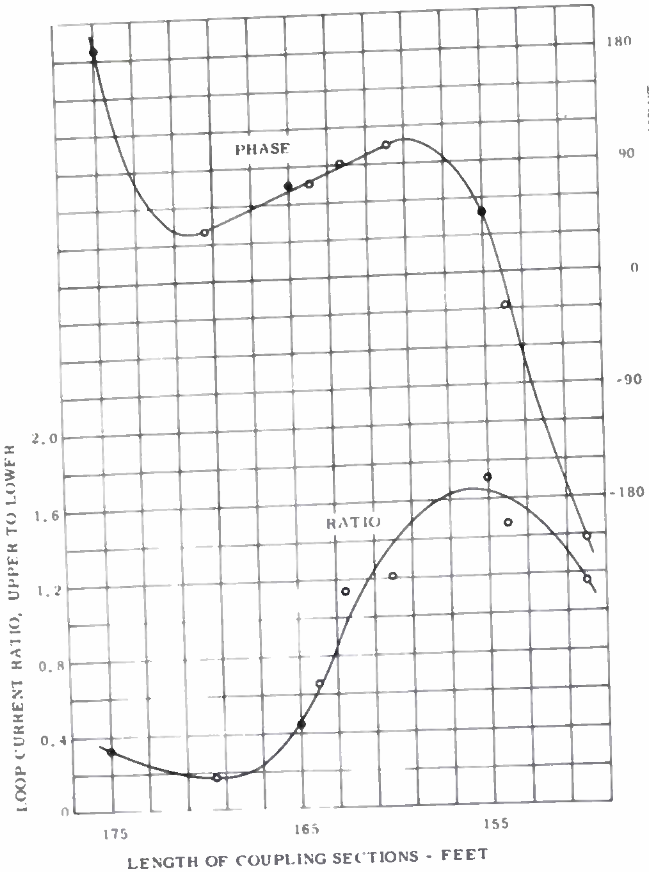

Fig. 3 Measured loop-current amplitude and phase relations vs. coupling-line length.

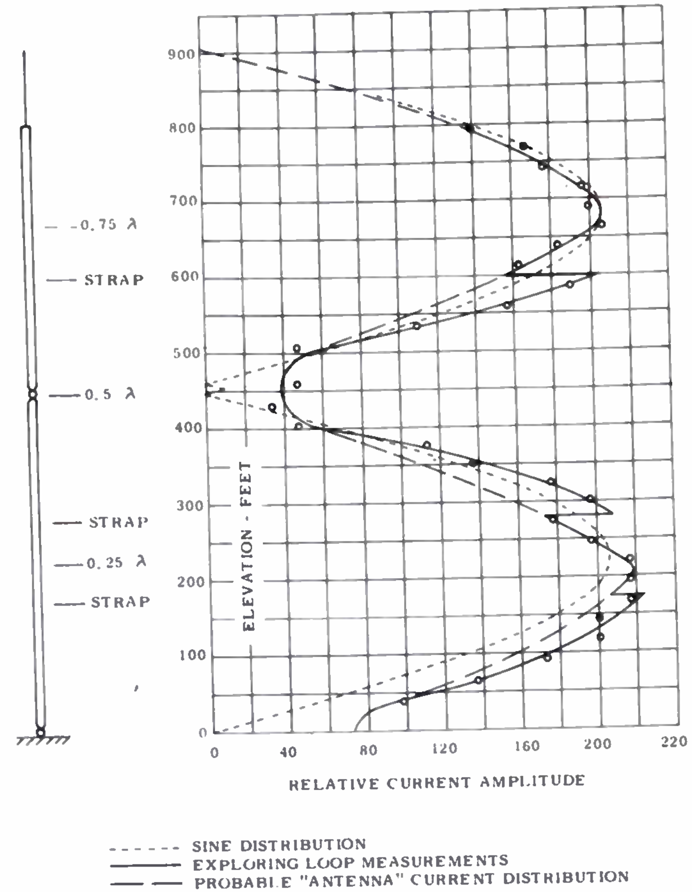

Fig. 4 Measured current amplitude distribution.

REMOTE CONTROLLED
50 KW BROADCAST TRANSMITTER

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Summary

This paper outlines adaptation of a Westinghouse Type HG-1 transmitter built in 1940 to remote control.

In order to assure maximum reliability, freedom from fire and intruder hazard, the entire transmitting facility will be moved to a new small, fireproof and intruderproof remote control type building located immediately adjacent to the transmitting tower, thus removing the necessity of transmission lines and tuning and terminating equipment external to the transmitter itself.

The transmitter building is a small, single-story structure having an area of less than 800 square feet which are ample for housing all necessary equipment including the transmitter and remote equipment.

The paper also outlines a separately located 5 KW CONELRAD and emergency plant facilities in the center of Pittsburgh.

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The remote control of a 50 KW broadcast transmitting plant can be divided into three basic problems:

1. The construction of the most reliable possible 50 KW unattended transmitting plant;

2. The remote control of this plant;

3. Provision for transmitting from alternate facilities during emergencies.

This paper describes some of the more important features of plans to convert KDKA-AM 50 KW transmitting plant to an unattended remote controlled operation.

The remote control equipment and its circuitry had previously been tested on an experimental basis as part of the NAB field tests. As a result of these tests and the FCC's later approval to permit remote operation of 50 KW broadcast transmitters, money was budgeted in 1957 for conversion of KDKA and WBZ to remote operation in 1958.

Since both the telephone company and electric power utilities have for years operated much of their plant on an unattended basis, it was natural to investigate what their experiences had taught them; particularly with respect to the buildings the unattended equipment was housed in and what protective features were necessary to guard against fire and unwanted intruders. Finally, what could be done with the equipment to improve its reliability?

The practices of the telephone company and the power company, perhaps as might be expected, were not always similar. However, there were several basic areas of agreement which can be summarized as follows:

With respect to the building:

Make it fireproof and weatherproof with the minimum possible number of doors and windows (if possible, one fireproof door and no windows);

Avoid basements, build on ground that is dry and stable;

Reduce all outside features to a minimum, use a strong, flat roof with slope only for water runoff;

Preferred building material cement blocks or reinforced concrete, with or without brick facing, depending on location -- residential areas usually require brick facing for appearance.

With respect to the unattended equipment:

Design and build it with unattended

operation in mind, and requiring minimum maintenance with long intervals between maintenance periods;

Reduce fire hazard inside the building to a minimum and make it easy to keep clean and protect;

Avoid gadgets which may not work. In other words, avoid all expendable items possible and design it rugged enough so that it needs only a minimum of protection equipment.

With the above criteria in mind, an examination of the existing KDKA transmitter building and transmitter quickly led to conclusions that major changes would be required in both building and transmitter.

Fig. 1 shows the exterior of the present KDKA transmitter building. It is a large, two story building with basement built in the late '30s to house three high power transmitters. It is only partially fireproof and, as can be seen, has many windows and doors. It is obvious that it would be difficult to make this building completely fireproof or intruder-proof.

Fig. 2 shows also that the transmitter building is a long distance (almost a thousand feet) from the transmitting tower which is connected to the transmitter by a four wire open line which is subject to weather problems, particularly ice and snow.

Fig. 3 shows the interior of the transmitter building looking toward the glass-enclosed control room with its input and monitoring and transmitter control equipment.

Fig. 4 shows the transmitter which is on the first floor and is air cooled. The input airduct is underneath the transmitter and mounted on the ceiling of the basement floor. The output air exhausts into an air duct running over the top of the transmitter.

Fig. 5 shows some of the equipment in the basement which houses all of the auxiliaries, mostly oil-filled transformers, metal rectifiers and the main blower equipment for cooling the transmitter.

The transmitter building, in addition to having space for three large transmitters, has a double-car garage, an office for the Chief Engineer, a shop, a kitchen, a tube and small parts room and a large storage area on the second floor as well as the basement. The total floor area of the basement and first and second floors is almost 15,000 square feet.

A more idealistic approach seems to be to:

1. Install the transmitter in a small, fireproof and intruder-proof building as close to the transmitting tower as possible.

2. Modify the present transmitter, removing all oil-filled transformers and replacing with air-cooled units.

3. Remove or replace all possible expendable components and simplify the control and supervisory equipment to the barest necessities.

Fig. 6 shows the interior of the proposed transmitter building. It is to be of cement block construction with reinforced concrete floor with strength as required by the transmitter and transformer loadings. Its interior floor area is approximately 650 square feet, slightly over 20 by 30 feet, and with eight foot inside clearance between the floor and ceiling. The roof will be prestressed concrete, flat and sloped only for rair runoff. There are no windows and only one double set of steel doors. The building is obviously fireproof and has nothing in it except the transmitter and its auxiliary equipment and a small toilet and wash basin.

Since the transmitter is air cooled it is possible to cool all equipment inside the building by dividing the inside building area into two sections, or plenum areas. The larger area, containing the blower, transformer equipment and "rack" equipment, is the input air area with the input air normally coming into this area via the storm hood, louvers through the filters and thence across the transformer equipment to the high voltage metal rectifier and through the blower. This air movement is due to the several inches of negative pressure produced by the blower on its input side. The output blower air under several inches of positive pressure is forced through the under floor concrete air duct running to and underneath the transmitter cubicles. It then travels upward through the several cubicles and is exhausted out the top into the exhaust area plenum space formed by a short drop wall from the ceiling of the building to the top front of the transmitter cubicles. A short door on the end of the transmitter, normally closed, completes the exhaust area space. Again, under normal conditions, air is exhausted out the rear wall of the building through protective louvers.

In winter weather automatic recirculation is arranged, using a Minneapolis-Honeywell control system which operates louvers on both the input and exhaust ducts as well as additional louvers not shown in the drop wall at the back of the transmitter. This system is to be controlled by a remote bulb temperature controller placed inside the building and can be arranged to have all outside air excluded until the interior building temperature exceeds a preset value. It can also be put in a manual position when maintenance is required on the transmitter. Also, if necessary for maintenance engineers' comfort during cold weather when the transmitter is not operating, additional building heat is to be supplied from electric strip heaters.

The building, externally, is completely intruder-proof except possibly against attempts to force liquids into the input or output air louvers, which will be screened and bared strongly, but still permit passage of air. The use of an ordinary electrical interlock on the double doors into the building is considered desirable, although a suitable mechanical lock should prevent any undesired entrance.

The public utilities' experience indicates almost complete freedom from any attempts to break into their unattended properties. Probably partly because it is clear they would be difficult to break into, and perhaps there is little of value to be found inside, anyway.

Protection of the building's contents from fire is first made as easy as possible by reducing to a minimum any combustible material inside it. All transformers and all other material in the transmitter, except in the two 2300 volt circuit breakers, is either completely high temperature dry insulated or non-combustible. The two 2300 volt input circuit breakers are oil insulated and, unfortunately, no suitable air operated replacement is available. Since these two circuit breakers are housed inside a completely enclosed steel cubicle, it appears practical to flood this cubicle with CO_2 if danger of combustion is indicated.

The fire alarm and control system under consideration is manufactured by Pyr-a-Larm, a division of Baker Industries, Inc. The detector system is a nuclear device which operates on low temperature combustion products and can therefore detect a fire before it actually becomes a flame. It provides the earliest possible warning and operates long before the usual temperature rise type detector will actuate. The Pyr-a-Larm is used by Con-Edison in their unattended substations.

One Pyr-a-Larm detector head is to be installed in the power switching cubicle and another in the center of the exhaust area. The detector heads feed a fire indicating cabinet, which will contain relays to operate remote alarms and operate fire extinguishers to flood the power cubicle with CO_2 gas.

The transmitter which is Serial #1 of the HG 50 KW plate modulated air cooled line is almost 20 years old, but it has proved extremely reliable and simple to operate. It contains a total of 20 tubes, 10 radio frequency and 10 audio. The radio side has five stages of amplification, including the plate modulated final. The audio side also has five stages including the modulator. All rectifiers are metal and none have been replaced. Tube life has been exceptionally long. The two modulator tubes now in service have over 50, 000 hours service, and the two power amplifier tubes have almost 90, 000 hours each. Other components have been designed most conservatively and are almost completely trouble free.

It is not contemplated to make any fundamental changes in the transmitter circuitry or components except to replace the oil-filled modulation transformer and Heising choke. All other components are already air cooled. We will also remove, where possible, unnecessary supervisory circuit and electrical gate interlocks which will be replaced with Cory type key interlocks.

The remote control equipment will be the same as used in the previous field test and is familiar to most broadcast engineering. Operation is based on measuring voltage across various circuits in the transmitter through a circuit of known and controllable resistance.

Assuming the maximum reliability possible has been built into the unattended remote control transmitter plant. its reliability still is never going to quite reach 100 percent.

For this reason, and also for convenience in CONELRAD operation, a second unattended low power (5 KW transmitter) is planned for installation in mid-town Pittsburgh (at former FM site of KDKA). This transmitter will have two RF sections and one modulator rectifier. It will also be remotely controlled and operate unattended. It will be used for all CONELRAD operations and for emergency operation in event of failure of the 50 KW transmitting plant.

Fig. 1

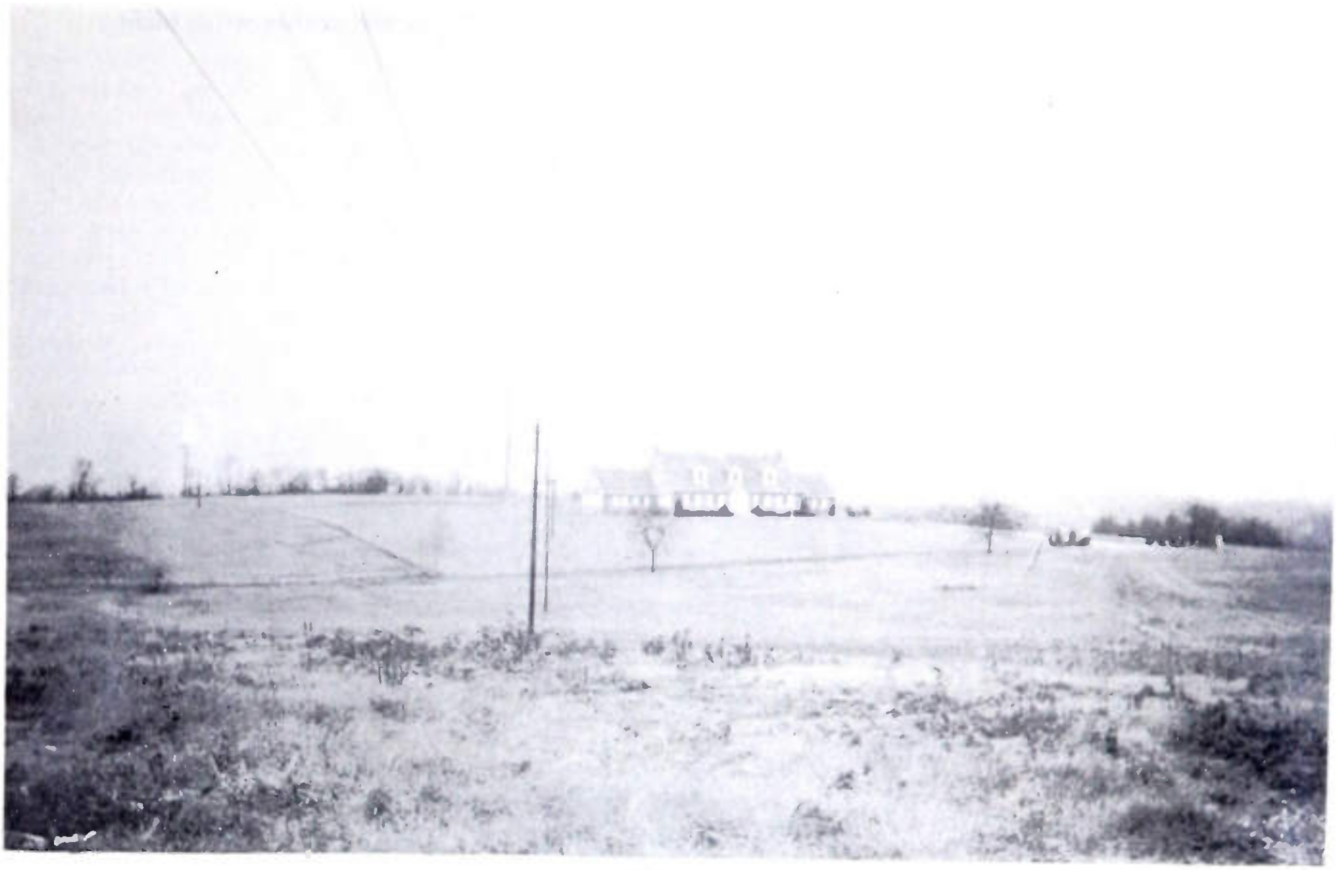

Fig. 2

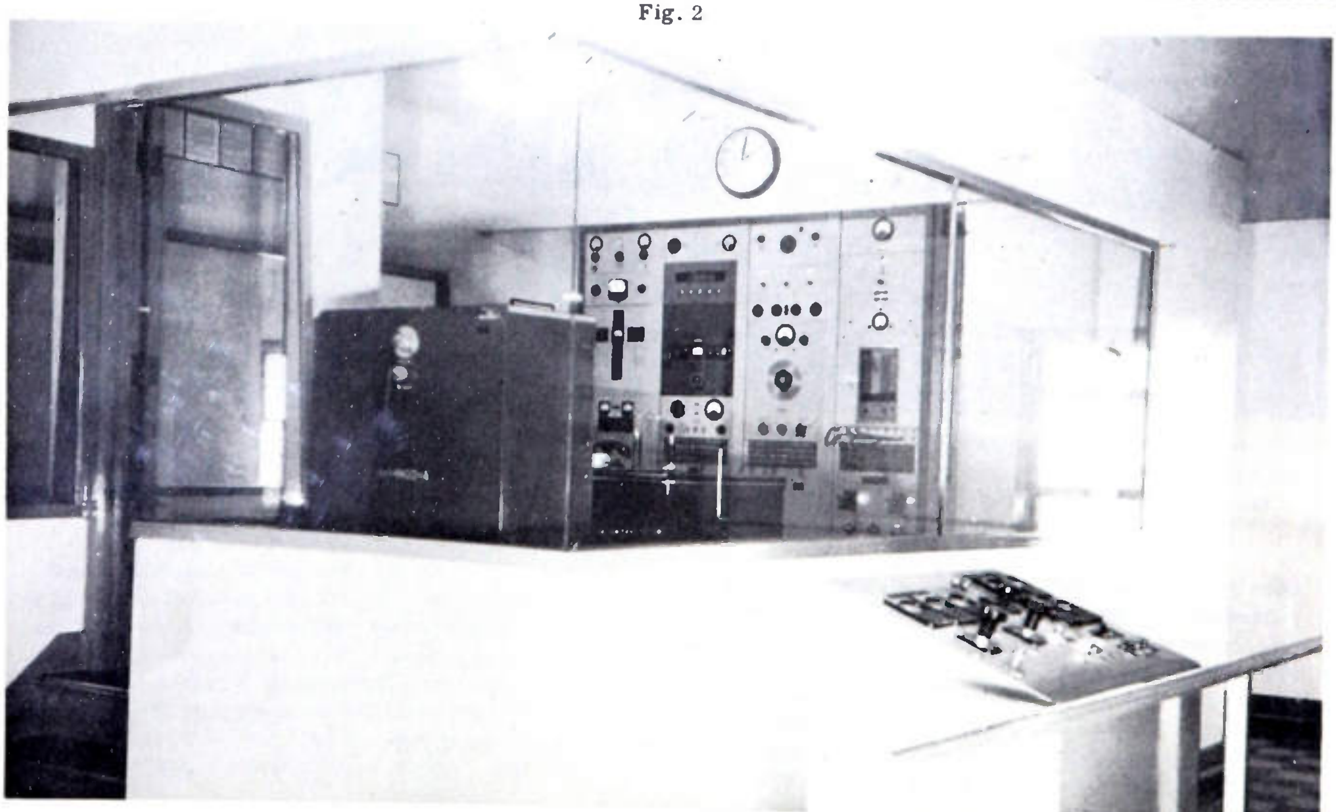

Fig. 3

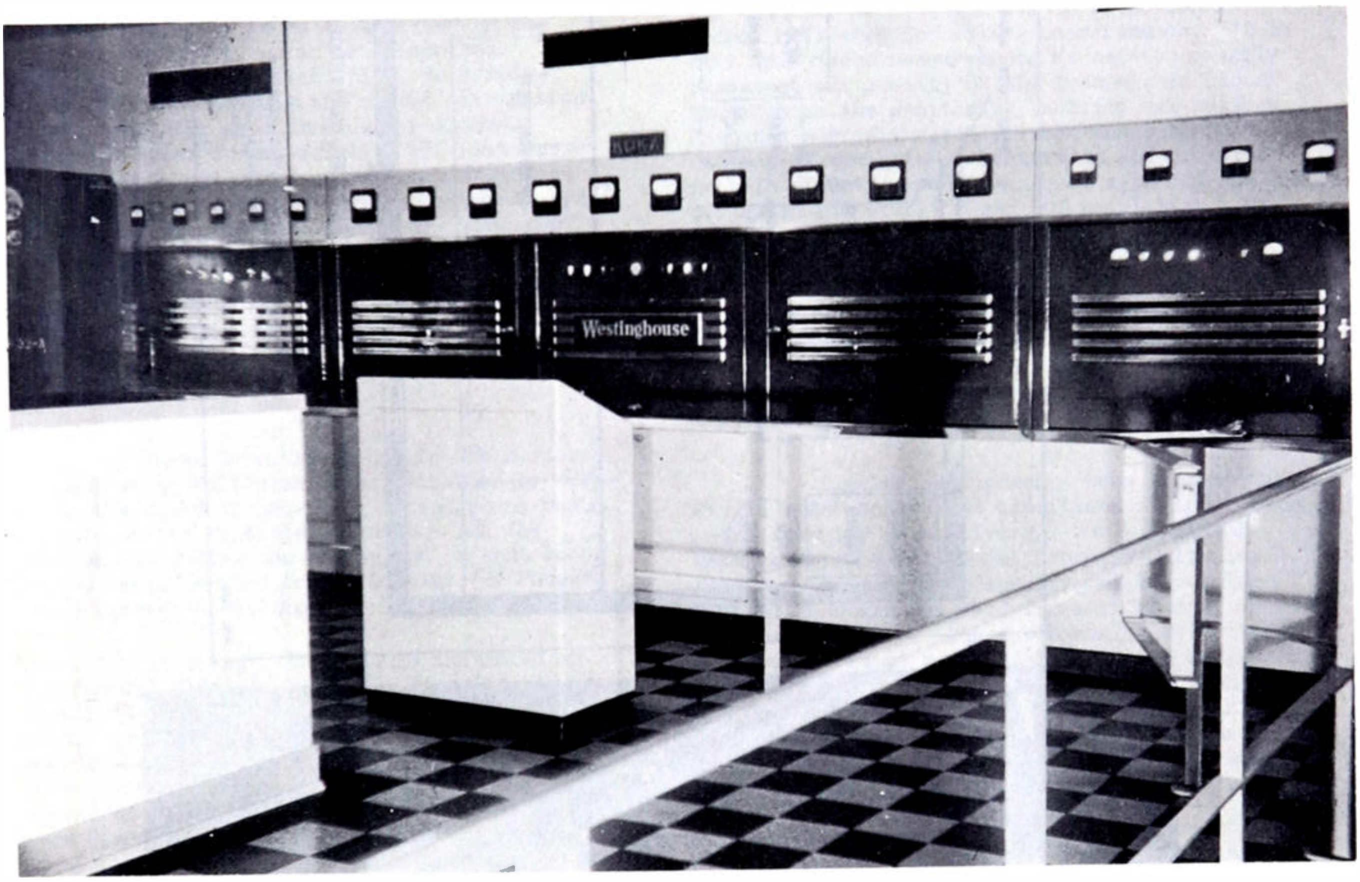

Fig. 4

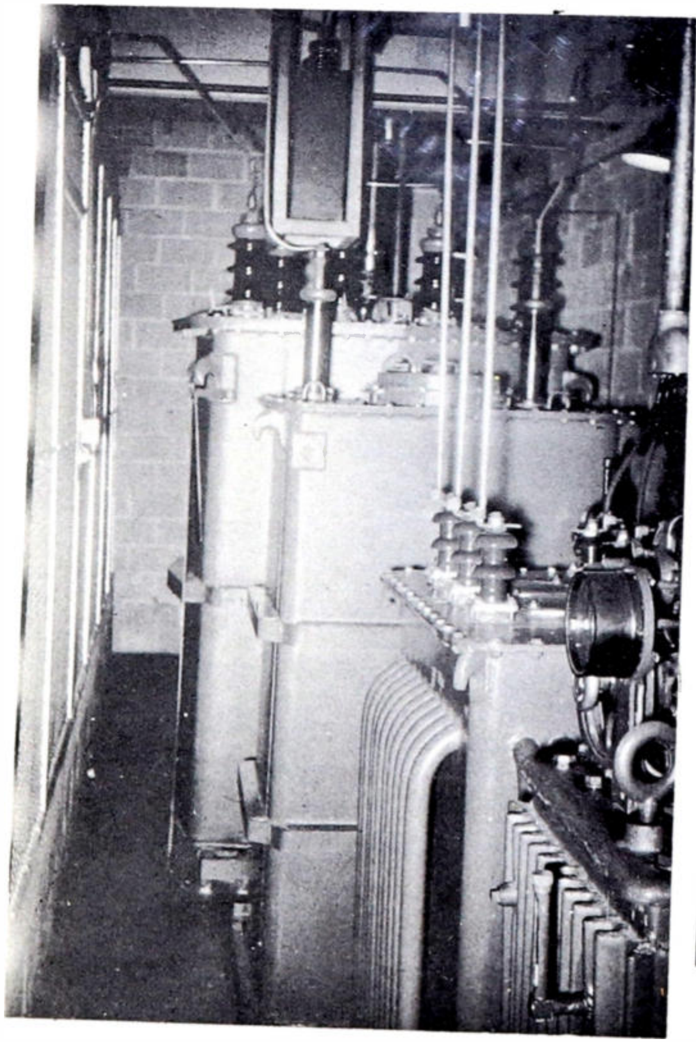

Fig. 5

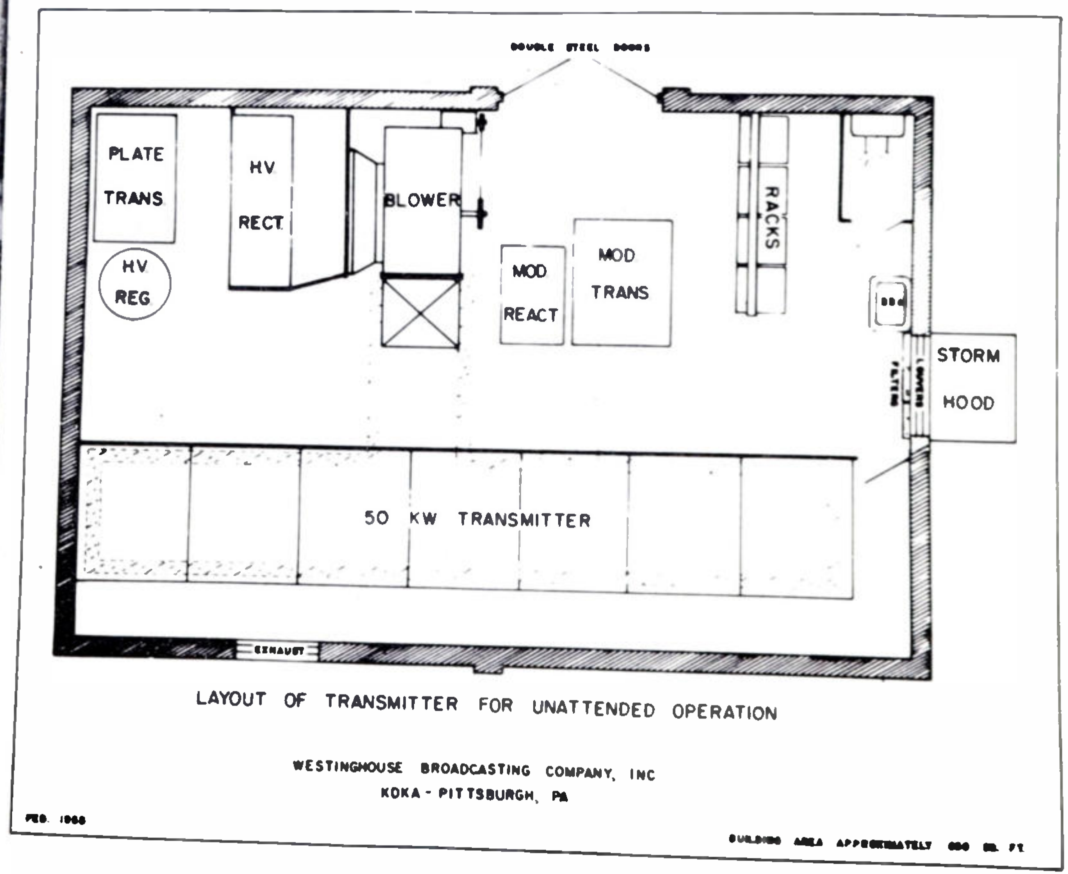
PLATE TRANS
HV REG
HV RECT
BLOWER
MOD REACT
MOD TRANS
RACKS
STORM HOOD
50 KW TRANSMITTER
LAYOUT OF TRANSMITTER FOR UNATTENDED OPERATION
WESTINGHOUSE BROADCASTING COMPANY, INC
KDKA - PITTSBURGH, PA

Fig. 6

REPORT ON MULTIPLEX EXPERIMENTAL WORK AT WCAU-FM

- By -

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Summary

The results of continuing Multiplex experimentation by an FM broadcaster in the field of background music are described. Problems relating to distortion, intermodulation between channels and signal attenuation on the Multiplex subchannel are reviewed. Practical steps toward solution of these problems are enumerated and operational standards developed as a result of these experiments are presented. The approach toward the addition of selective muting to Multiplex service is discussed. Direct comparison between main channel and subchannel service is afforded. An evaluation of the present status and future promise of Multiplex subchannel transmission of background music, based on this work, is offered.

There is outlined herein the results of continuing Multiplex experimentation by WCAU-FM in the field of providing a background music service on its Multiplex channel - and the problems attending the conversion of that background music service from a Simplex (or "beep" type) service to Multiplex.

In June, 1957, WCAU-FM installed on its General Electric 10 KW FM Transmitter a prototype model of the RCA multiplexing equipment. During the first several months of operation, tests were conducted by RCA and by WCAU engineering personnel to determine the degree of quality of the Multiplex transmission over WCAU-FM. Initial standards were established, using the subcarrier frequency of 67 KC and various percentages of modulation and deviation.

Problems immediately arose relating to distortion, intermodulation between channels (more commonly known as crosstalk) and signal attenuation on the Multiplex subchannel. Based upon the limitations which were immediately apparent in the early development of the Multiplex facility at WCAU-FM, steps toward solution of the problems described above were taken with a view to developing operational standards for the Multiplex service.

Crosstalk, noise and distortion measurements were taken on the 67 KC subchannel of the FM transmitter. Measurements were made at the transmitter location, at the Studio location approximately 2 miles from the transmitter, and at a number of fringe-area locations. Noise and distortion on the subchannel, crosstalk into the subchannel from the main channel and crosstalk into the main channel from the subchannel were measured under various operating conditions. Measurements were made with 15% modulation and 30% modulation of the subchannel on the main channel, using $\pm$ 8KC and $\pm$ 12KC deviation on the subchannel.

The purpose of measurements made at the transmitter location was to establish a common reference for future measurements. It is felt that these measurements do not necessarily represent the quality of the transmitted signal inasmuch as the monitoring facility was neither designed nor calibrated for use under such conditions of operation. Distortion and crosstalk in both directions were measured with 8KC and 12KC deviation and with 15% and 30% modulation. Results are tabulated in TABLE 1.

Crosstalk results are plotted in FIGURE 2. The manufacturer of the Demodulator states that inherent distortion is high and is not considered suitable as a measuring device. For this reason, distortion measurements are not included.

Similar measurements were made at the Studio under controlled conditions of input level, using Browning FM Receivers. For subchannel measurements, a late-model Browning Multiplex Receiver was used. An earlier-model Browning Simplex Receiver was used for main channel crosstalk measurements. The subchannel receiver was first measured as received from the factory - after recommended set-up adjustments - then measured again after adjusting the I.F. strip and discriminator transformer for minimum crosstalk at 3,000 cycles, which seemed to be the frequency at which the highest crosstalk was induced.

The improvement resulting from this re-alignment varied from 4db to 17db, depending upon the degree of subcarrier modulation. While not tabulated or plotted, the effect of main channel receiver re-alignment varied as much as 40db. R.F. input level to the subchannel receiver was varied from 100 to 15,000 microvolts, keeping at all times correct voltage at the 1st limiter grid. It was noted that no significant change occurred from 100 to 5,000 microvolts. As the signal level was increased above 5,000 microvolts, the crosstalk - main into subchannel - increased by approximately 3db at 15,000 microvolts.

Measurements and listening tests were

made at Reading, Pa., Lancaster, Pa., Schuylkill Haven, Pa., and Atlantic City, N. J. Tone was used on the subchannel to set reference level and for measurement of distortion. Crosstalk into the subchannel from the main channel was measured on the basis of main channel program peaks. Measurements and listening tests, in all cases, were made with 15% and 30% subchannel modulation, using ≠ 8 KC and ≠ 12 KC deviation on the Multiplex subchannel.

Measurements made in Lancaster, Pa., in February of this year at a distance of 57 airline miles from the WCAU-FM transmitter indicated a high level of impulse-type background noise on the subchannel. For this reason, accurate measurements could not be made. The results of an attempt to get some data for evaluation are shown in TABLE 5. Results agreed, roughly, with those obtained at the Studio. Crosstalk was not considered an item of concern. and the noise level was considered too high to provide a satisfactory commercial service. No noise was observed on the main channel.

In Reading, Pa., a distance of 43 airline miles, listening tests were conducted on both main and subchannel. Background noise on the subchannel and crosstalk into the subchannel were considered too high to provide a satisfactory service.

R.F. input to the receiver at Reading was approximately 2½ times that obtained at Lancaster. Again, no noise was observed on the main channel.

In Atlantic City, N. J., at a distance of 64 airline miles from the WCAU-FM transmitter, listening tests and measurements were made on the subchannel, and listening tests were made on the main channel. The results are tabulated in TABLE 6. Because of the lower background noise in this location, the accuracy of the measurements is considerably improved over those made at Lancaster: Results agreed more closely with those obtained at the Studio. The Multiplex channel was considered satisfactory for commercial service.

In general, our experience indicated a 3 to 4 db improvement of crosstalk into the subchannel by increasing deviation from ≠ 8 KC to ≠ 12 KC. An additional 1 to 2 db improvement was realized by increasing the percentage of modulation on the subchannel from 15% to 30%. Distortion on the subchannel was at all times less than 3%.

It should be noted here that it is difficult to accurately assign distortion or crosstalk figures to individual components in the overall system. This is primarily due to lack of information on the individual components and to lack of suitable field measuring equipment which would permit individual equipment evaluation. The results obtained and the data presented in this report, therefore, are overall system measurements. Until a suitable transmitter monitor is available to permit an accurate evaluation of the transmitted signal, it is felt that a Multiplex performance thinking must, of necessity, be along the lines of the overall system.

As measurements were taken in fringe locations, it was generally apparent that crosstalk was not the main problem in providing a satisfactory commercial service in those areas beyond 40 miles from the transmitter. Impulse-type ignition was affecting the weak Multiplex FM signal much in the same manner that it can affect a weak amplitude modulated signal on standard radio.

In recent weeks, two additional steps have been taken to improve the marginal performance in these fringe localities. A special R.F. booster, made on a custom-built basis for WCAU-FM by the Tapetone Company of Webster, Massachusetts, was utilized in extreme fringe areas to determine its effect upon the background noise which was evident on both main and subchannel in extremely-low-signal areas. Preliminary reports on the effectiveness of this unit indicate that a reduction of between 15 and 20db in the ignition-type noise may be obtained by use of a high-gain booster with front-end noise, on the order of 2db.

Based upon additional tests and discussions with Browning Laboratories, a new approach to a subchannel clipper was developed whereby the apparent peak-type impulse noise was reduced far beyond its measured characteristic, as shown on the charts developed previously, to a level which makes the transmitted signal commercially acceptable in almost all cases up to approximately 65 miles.

For the past two weeks, WCAU-FM has been serving a Multiplex subscriber 71 miles distant from the transmitter location. It is felt, therefore, that with the new limiting technique utilized in the last Browning Multiplex Receiver model, with the high-gain - low-noise booster, and with proper choice of receiving antenna and correct orientation of the array, it is possible to provide a commercial Multiplex service to roughly the same area as has been previously serviced on the main channel.

At the present time, WCAU-FM has accepted 67 KC as a standard subchannel frequency and is utilizing ≠ 8 KC deviation at 15% modulation.

Selective muting is being accomplished

in the Multiplex Service with frequencies of 20 KC and 25 KC for selective restoration and a common frequency of 22.5 KC for muting.

The spectrum between 25 KC and 59 KC is being reserved for the anticipated addition of a second Multiplex subchannel which will probably be centered at approximately 42 KC.

It is significant that in the continued development of Multiplex in recent months the long-standing problem of crosstalk has been greatly minimized and that the principal problem facing the FM broadcaster in conversion from Simplex into Multiplexing now becomes one of extending, on the Multiplex channel, the range which was previously available to him on his main channel. It is felt that from a standpoint of frequency response and distortion, Multiplex is no longer a "second-class service" but is fully as acceptable for background music - from an audio standpoint - as main channel FM service.

The author wishes to recognize the valuable contribution made by Frank J. Haney of the WCAU Engineering Staff who personally measured and evaluated the statistics shown in this presentation; and to acknowledge the assistance of Messrs. J. G. Leitch, George Lewis and James J. Ouinn of the WCAU Engineering Staff - as well as that of Messrs. Elliot Baker and Harry Paul of Browning Laboratories and Messrs. Frank Talmadge and Hans Bott of Radio Corporation of America - without whose excellent contributions the successful conclusion of these experiments would not have been possible.

TABLE 1

MAIN AND SUB CHANNEL MEASUREMENTS - TRANSMITTER

G.E. BM-1A monitor serial WG-293 with
RCA 67 KC demodulator

Freq. C.P.S.	Crosstalk into Sub. 15%		Crosstalk into Sub. 30%		Crosstalk into Main 15%		Crosstalk into Main 30%	
	t_{8KC}	t_{12KC}	t_{8KC}	t_{12KC}	t_{8KC}	t_{12KC}	t_{8KC}	t_{12KC}
50	-48	-52	-53	-57	-67	-67	-64	-64
100	-43	-47	-49	-53	-67	-67	-64	-64
200	-39	-43	-45	-49	-67	-67	-64	-64
400	-38	-42	-45	-49	-67	-67	-64	-64
800	-41	-45	-47	-51	-66	-65	-64	-64
1500	-41	-45	-46	-50	-62	-62	-56	-55
3000	-42	-46	-45	-49	-57	-57	-51	-50
6000	-43	-47	-46	-50	-55	-55	-49	-49
10,000	-48	-52	-52	-56	-67	-67	-66	-66
15,000	-50	-54	-53	-57	-67	-67	-66	-66
S/N	-51	-55	-53	-57	-67	-67	-66	-66

(a) 15 KC filter used on main channel measurements.

WCAU-FM

TABLE 2

SUB CHANNEL MEASUREMENTS - STUDIO

Browning BR 100 Serial 0061 67 KC Multiplex
RCVR as received from factory

Freq. C.P.S.	Crosstalk into Sub. 15% ±8KC	Crosstalk into Sub. 15% ±12KC	Crosstalk into Sub. 30% ±8KC	Crosstalk into Sub. 30% ±12KC	Distortion 15% ±8KC	Distortion 15% ±12KC	Distortion 30% ±8KC	Distortion 30% ±12KC
50	-52	-55	-53	-56	1.4	2.8	1.0	2.3
100	-51	-54	-52	-55	1.5	2.7	1.2	2.2
200	-50	-53	-47	-50	1.7	2.6	1.4	2.0
400	-47	-50	-42	-45	1.3	2.3	1.2	1.8
800	-45	-48	-38	-41	1.4	2.0	1.1	1.6
1500	-42	-45	-35	-38	1.4	1.7	1.4	1.4
3000	-41	-44	-34	-37	1.3	1.4	1.4	1.2
6000	-47	-50	-38	-41	1.6	1.1	1.4	1.0
10,000	-52	-55	-52	-54	-	-	-	-
15,000	-52	-55	-52	-56	-	-	-	-
S/N	-52	-55	-53	-56				

(a) R.F. level into Rcvr 100 to 5,000 microvolts.

(b) Rcvr adjusted for 10 V at limiter Grid.

(c) All meas. on output of multiplex adapter unit.

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TABLE 3

SUB CHANNEL MEASUREMENTS - STUDIO

Browning BR 100 Serial 0061 67 KC Multiplex
Rcvr. adjusted for min. crosstalk; all
I.F. cans & discriminator adjusted

Freq. C.P.S.	Crosstalk into Sub. 15% ∠8KC	Crosstalk into Sub. 15% ∠12KC	Crosstalk into Sub. 30% ∠8KC	Crosstalk into Sub. 30% ∠12KC	Distortion 15% ∠8KC	Distortion 15% ∠12KC	Distortion 30% ∠8KC	Distortion 30% ∠12KC
50	-53	-56	-53	-56	1.4	2.8	1.2	2.3
100	-53	-56	-53	-56	1.5	2.7	1.3	2.2
200	-52	-56	-53	-56	1.7	2.6	1.4	2.0
400	-49	-53	-50	-53	1.3	2.3	1.2	1.8
800	-47	-50	-48	-51	1.4	2.0	1.1	1.6
1500	-45	-48	-47	-51	1.4	1.7	1.4	1.4
3000	-45	-49	-51	-54	1.3	1.4	1.4	1.2
6000	-47	-50	-52	-54	1.6	1.1	1.4	1.0
10,000	-53	-56	-53	-56	-	-	-	-
15,000	-53	-56	-53	-56	-	-	-	-
S/N	-53	-56	-53	-56				

(a) R.F. level into Rcvr 100 to 5,000 microvolts

(b) Rcvr adjusted for 10 V. at limiter grid.

(c) All meas. on output of adapter unit.

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TABLE 4

MAIN CHANNEL CROSSTALK - STUDIO

Browning CR-3 Simplex Rcvr Serial 2283 adjusted
for min. crosstalk; all I.F. cans & discriminator adjusted

Freq. C.P.S.	15%		30%	
	±8KC	±12KC	±8KC	±12KC
50	-66	-64	-63	-61
100	-66	-64	-64	-60
200	-66	-65	-65	-60
400	-66	-65	-65	-61
800	-66	-65	-65	-61
1500	-65	-64	-58	-56
3000	-60	-57	-49	-46
6000	-58	-54	-48	-44
10,000	-66	-66	-66	-66
15,000	-66	-66	-66	-66
S/N	-66	-66	-66	-66

(a) RF level into Rcvr 100 to 15,000 microvolts.

(b) Rcvr adjusted for 10 V at limiter grid.

(c) Meas. made on output of simplex adapter unit.

(d) 15 KC filter used.

(e) Adjustment of Disc. Transf. and I.F. cans will give variations of 40 db of above figures. As received from factory Rcvr was approx. 10 db poorer than above figures.

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TABLE 5

SUB CHANNEL MEASUREMENTS - LANCASTER, PA. (57 miles)

Browning BR-100 Serial 0061
67 KC Multiplex Receiver

	CROSSTALK INTO SUB (program material)				DISTORTION (400 cps)			
	15%		30%		15%		30%	
	±8KC	±12KC	±8KC	±12KC	±8KC	±12KC	±8KC	±12KC
	*	*	-46	-50	*	*	1.5	1.5
S/N	-45	-47	-50	-54				

(a) R.F. level into Rcvr 100 microvolts.

(b) Rcvr adjusted for 10 V. at limiter grid.

(c) All meas. on output of adapter unit.

(d) High background noise level, noise peaks to -20.

(e) *indicates background noise too high to permit measurement.

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TABLE 6

SUB CHANNEL MEASUREMENTS - ATLANTIC CITY, N.J. (63.5 miles)

Browning BR-100 Serial 0061
67 KC Multiplex Receiver

	CROSSTALK INTO SUB (program material)				DISTORTION (400 cps)			
	15%		30%		15%		30%	
	±8KC	±12KC	±8KC	±12KC	±8KC	±12KC	±8KC	±12KC
	-45	-48	-46	-49	1.5	2.3	1.2	1.4
S/N	-54	-56	-54	-56				

(a) R.F. level into Rcvr. 410 microvolts.

(b) Rcvr adjusted for 10 V. at limiter grid.

(c) All meas. on output of adapter unit.

(d) Low background noise level, occasional noise peaks to -40.

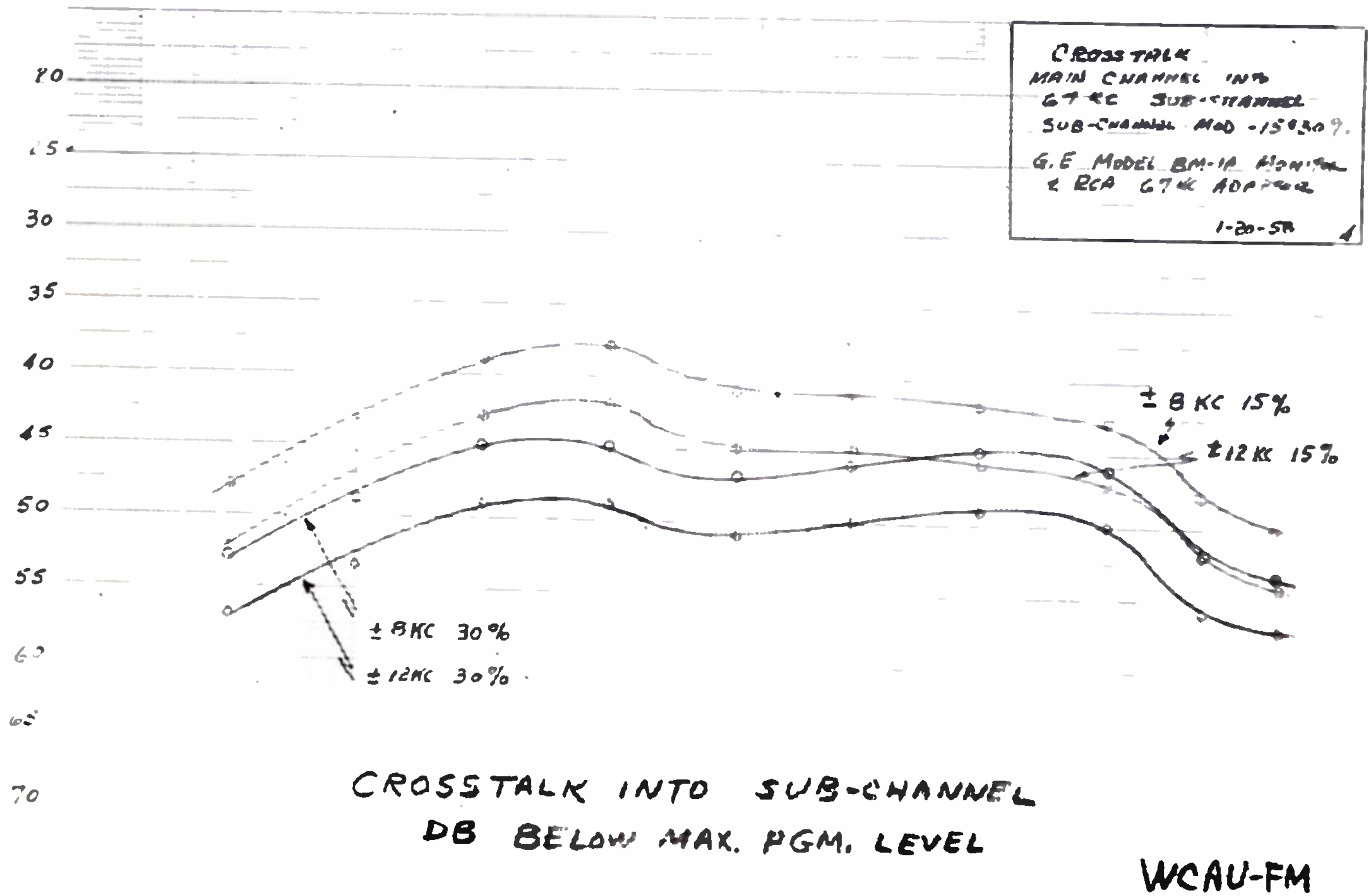

CROSSTALK
MAIN CHANNEL INTO
67 KC SUB-CHANNEL
SUB-CHANNEL MOD -15 & 30 %.
G.E MODEL BM-1A MONITOR
& RCA 67 KC ADAPTER
1-20-58
20
25
30
35
40
45
50
55
60
65
70
± 8 KC 15%
± 12 KC 15%
± 8 KC 30%
± 12 KC 30%
CROSSTALK INTO SUB-CHANNEL
DB BELOW MAX. PGM. LEVEL
WCAU-FM

Fig. 1

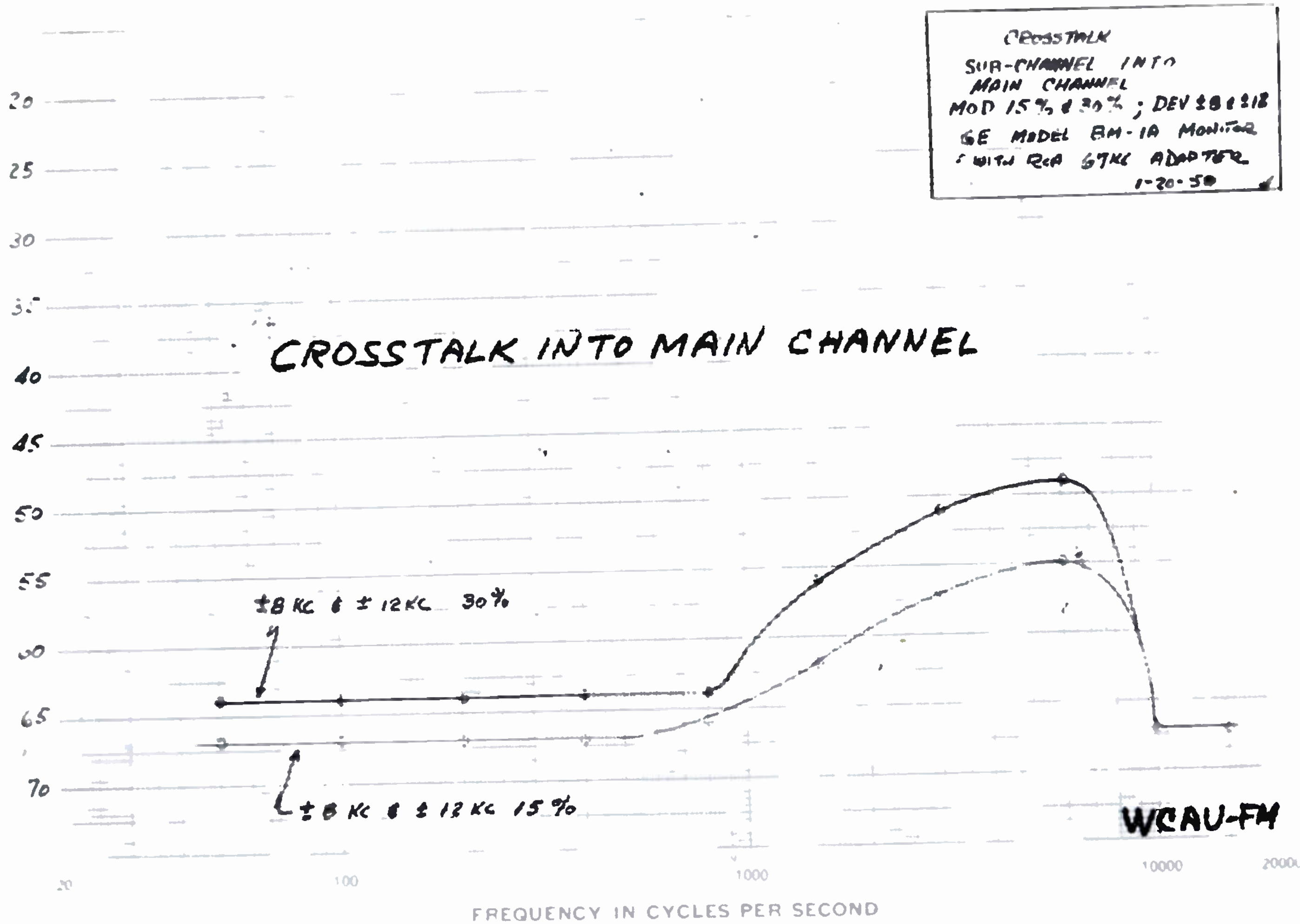

CROSSTALK
SUB-CHANNEL INTO
MAIN CHANNEL
MOD 15% & 30% ; DEV ±8 & ±12
GE MODEL BM-1A MONITOR
WITH RCA 67KC ADAPTER
1-20-50
CROSSTALK INTO MAIN CHANNEL
20
25
30
35
40
45
50
55
60
65
70
±8 KC & ±12 KC 30%
±8 KC & ±12 KC 15%
WCAU-FM
20
100
1000
10000
20000
FREQUENCY IN CYCLES PER SECOND

Fig. 2

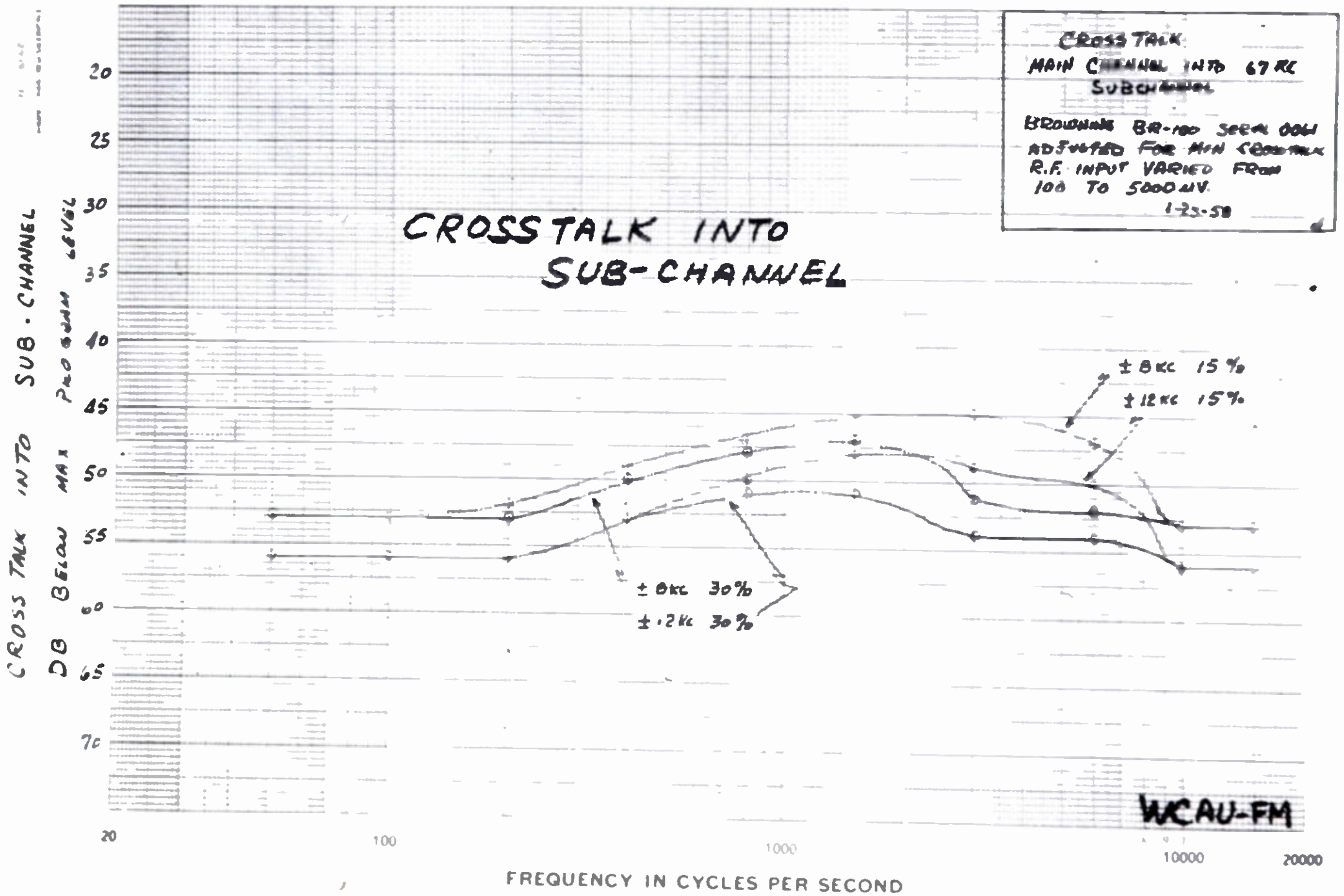
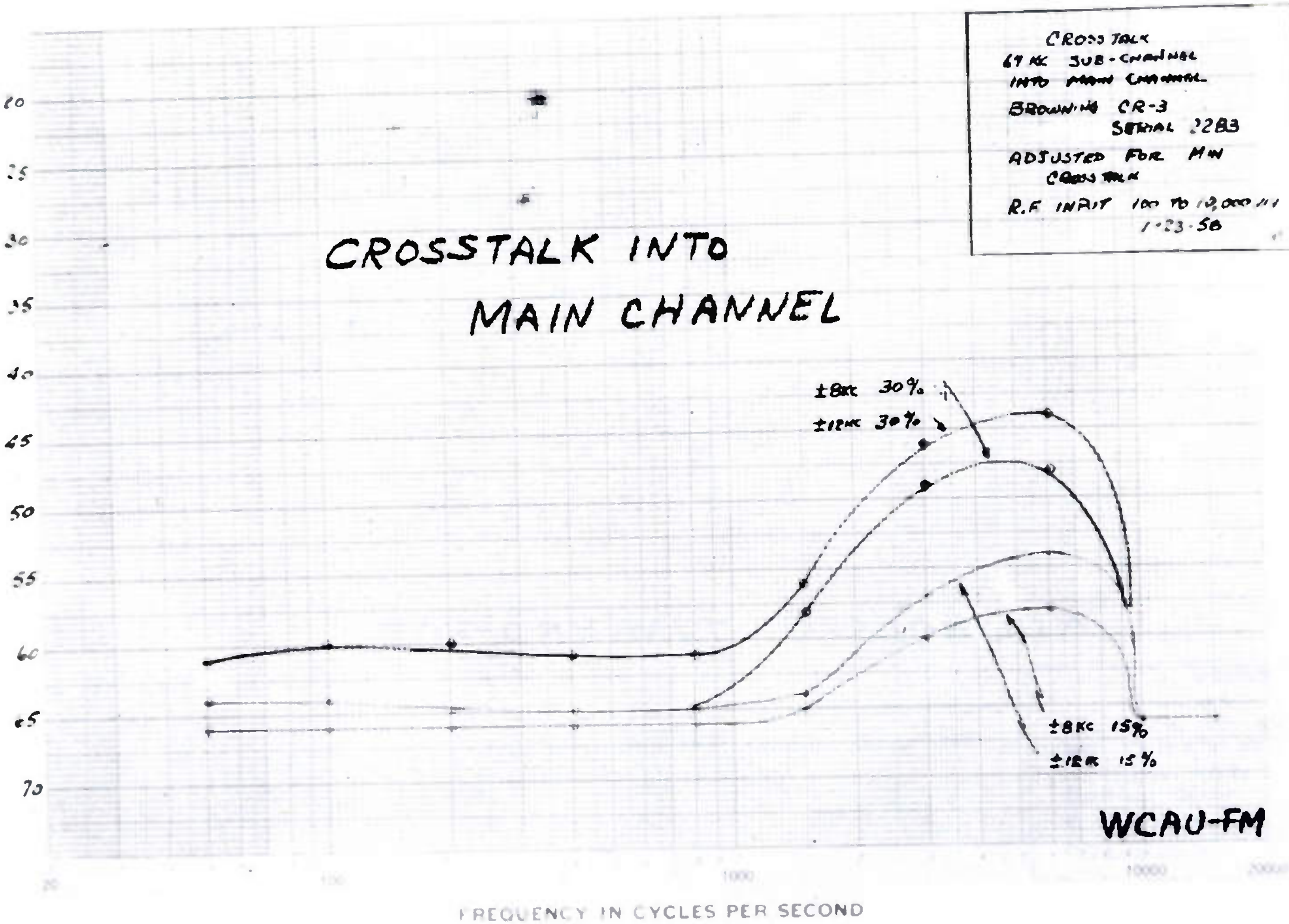

CROSS TALK
R.F. INPUT VARIED FROM
100 TO 5000 µV.
1-23-58
CROSSTALK INTO
SUB-CHANNEL
± 8KC 15%
± 12KC 15%
± 8KC 30%
± 12KC 30%
CROSS TALK INTO SUB-CHANNEL
DB BELOW MAX PROGRAM LEVEL
20
25
30
35
40
45
50
55
60
65
70
20
100
1000
10000
20000
FREQUENCY IN CYCLES PER SECOND
WCAU-FM

Fig. 3



CROSSTALK INTO
MAIN CHANNEL
CROSS TALK
67 KC SUB-CHANNEL
INTO MAIN CHANNEL
BROWNING CR-3
ADJUSTED FOR MIN
CROSSTALK
R.F INPUT 100 TO 10,000
1-23-58
±8KC 30%
±12KC 30%
±8KC 15%
±12KC 15%
WCAU-FM
20
25
30
35
40
45
50
55
60
65
70
1000
10000
20000
FREQUENCY IN CYCLES PER SECOND

Fig. 4

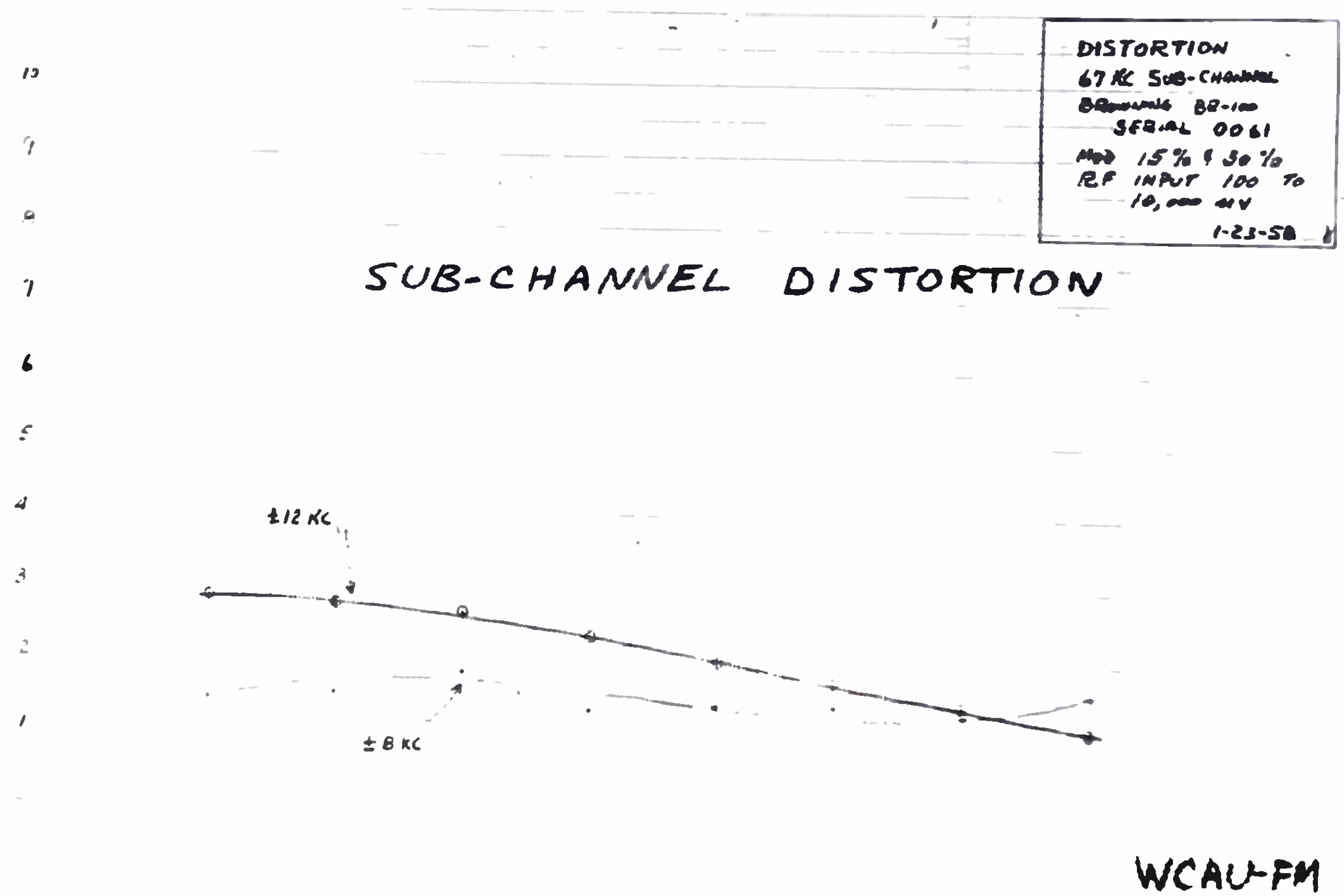

DISTORTION
67 KC SUB-CHANNEL
BR-100
SERIAL 0061
15% & 30%
RF INPUT 100 TO
1-23-58
SUB-CHANNEL DISTORTION
±12 KC
±8 KC
WCAU-FM

Fig. 5

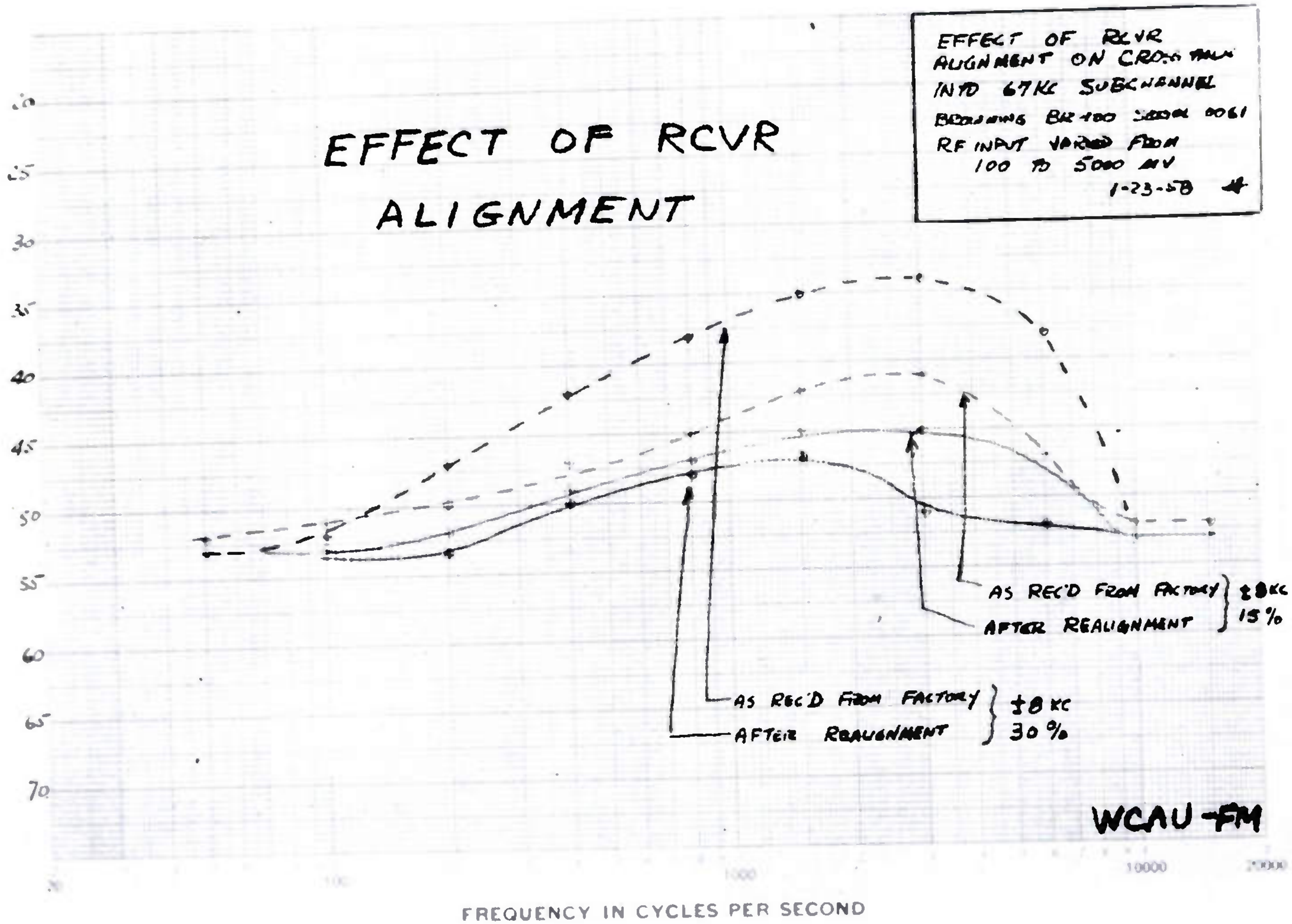

EFFECT OF RCVR
ALIGNMENT
EFFECT OF RCVR
ALIGNMENT ON
INTO 67KC SUBCHANNEL
BR-100 0061
RF INPUT VARIED FROM
100 TO 5000 MV
1-23-58
30
35
40
45
50
55
60
65
70
AS REC'D FROM FACTORY
AFTER REALIGNMENT
±8KC
15%
AS REC'D FROM FACTORY
AFTER REALIGNMENT
±8 KC
30%
WCAU-FM
1000
10000
20000
FREQUENCY IN CYCLES PER SECOND

Fig. 6

WABC Field Test of Compatible Single Side-Band Transmission

Frank Marx and Robert M. Morris

American Broadcasting Co.

During the early part of 1957, interest in Single Side-band transmission for Radio Broadcasting, increased sharply as a result of the work of Leonard R. Kahn in the development of what he called Compatible Single Side-band. The term, "Compatible", used in this connection, refers to the fact that this form of single side-band transmission is designed to be received by conventional broadcast receivers without the disadvantages of distortion arising from the use of conventional single side-band telephony with inserted carrier.

Tests, run principally during the early morning hours on Station WMGM in New York City, indicated the desirability of a more comprehensive field test to permit the evaluation of Compatible Single Side-band by the public with receivers currently in use for Radio Broadcasting. Accordingly, plans were made by the American Broadcasting Company in cooperation with the Kahn Research Laboratories to conduct an extensive field test of Compatible Single Side-band on Radio Station WABC.

WABC is a station of 50 kw, operating twenty-four hours a day on 770 kilocycles with an omni-directional antenna. An application was filed with the FCC in Auguest of 1957 requesting authorization for such a field test. It was indicated that anticipated advantages believe possible from the use of Single Side-band were:

1. An increase in effective signal equivalend to twice the power.

2. An improvement in fidelity with an attendant increase in loudness due to increased brilliance as heard on most home receivers.

3. A reduction in selective fading distortion in the fringes of the service area.

It was requested that authorization for test transmissions include the use of regular programs of Station WABC. Authorization of the FCC, as requested, was granted on August 13th and included the use of experimental call letters KE2XWJ. Having the authorization, steps were taken promptly to obtain the necessary equipment from Kahn Research Laboratories for CSSB operation and preparations were made for the interconnection of such equipment with the 50 KE RCA 50B transmitter in use at WABC. The 50B transmitter is a low level (500 watts) modulated transmitter with two linear stages at power levels of 5 and 50 kw. The RF output of the exciter, it was determined, would be at approximately 5 watts level and would require that it be inserted into the transmitter at the point normally fed by the crystal oscillator unit.

At this point, it might be well to discuss the characteristics of Compatible Single Side-band and the problems involved in modification of a conventional AM transmitter for this system. Figure 1, Curve A, indicates the envelope of a conventional two-tone Single Side-band signal involving a carrier and single frequency modulation at a level such that the Side-band is equal in amplitude to the carrier. A receiver having a conventional envelope detector receiving this signal will have in its output, distortion products of considerable amplitude as compared to the result obtained with single frequency modulation in AM shown in Curve I-B. The difference between these two curves approximates the maximum difference between CSSB and AM, which is compensated for by the processing of the signal in the Compatible Single Side-band exciter unit. This processing consists, basically, of the generation of a phase modulated carrier and a modulation envelope or audio signal with such envelope delay as to combine in the normal modulator of an AM transmitter with the phase modulated signal to produce the desired CSSB result. The equipment supplied by Kahn Research Laboratories for the WABC test is shown in Figure 2. It consists of six units mounted in an equipment rack separate from the transmitter. Two of the units are plate, filament and bias supplies. The upper unit, marked SSB-4, is a monitoring unit including a test oscillator, an oscilloscope and a Side-band analyzer for checking Side-band suppression. The other three units comprise the basic side-band generating equipment with all processing necessary for turning out the delayed audio signal to be applied to the transmitter audio input and the phase modulated RF excitation signal at the correct carrier frequency. Block diagrams of the generating equipment are shown in Figures 3,4 and 5.

SSB-1 (Figure 3) is the basic single side-band generating unit comprising a balanced modulator, into which the program audio and a 100 kc crystal oscillator are connected. The double side-band output of this balanced modulator is passed through a crystal filter and amplified as single side-band, suppressed carrier. The crystal oscillator is also connected to a multiplier with double tuned plate output from which 600 kc and 700 kc are derived. The 600 kc component together with the 100 kc single side-band signal is fed into a balanced mixer and the upper (700 kc) side-band is recovered. At this point, 700 kc separately derived from the multiplier is mixed with the 700 kc single side-band signal, thus creating, for the first time, single side-band with associated carrier in the desired amplitude relationship.

The second unit, SSB-2, (Figure 4), is a processing unit, the function of which is to create a phase modulated signal and the audio signal necessary for Compatible Single Side-band operation. In this unit, a product detector supplied with the single side-band signal and a 700 kc carrier creates an audio output with the correct phase delay for subsequent re-modulation in the transmitter. The single side-band signal is also fed through a limiter amplifier which eliminates amplitude variations in this signal and creates the phase modulated carrier necessary for the re-modulation process.

The third or converter unit, (Figure 5), is for the purpose of changing the 700 kc phase modulated output of SSB-2 to the carrier frequency required by the station, in this case 770 kc. It involves a double modulation process in which the output of a 770 kc crystal is added as a difference frequency, thus producing a 770 kc phase modulated output at a level of approximately five watts.

In adapting the 50B Transmitter to CSSB, it was necessary to install two relays between the output of the normal crystal oscillator unit and the grid of the 860 buffer amplifier as shown in Figure 6. These relays together with a separate tuned circuit connected to the Kahn unit made it convenient to switch between the two modes of operation. Much testing and adjustment was necessary, initially, to achieve satisfactory frequency response, side-band suppression and distortion when using CSSB. It was necessary to improve the by-passing of circuits in the 860 stage. It was also necessary to apply neutralization to the 860 stage, a tetrode, which, up to this time, had required no neutralization. The adjustment of neutralization on the 849 stage was also more critical. The bias on the 860 stage was reduced from 170 volts to 80 volts in order to increase output to the modulated stage. With this done, the transmitter operating in CSSB met the standards of performance normally required for AM transmitters.

Since the principle purpose of these tests was to determine what effect the use of compatible single side-band would have on radio program service as received by the public with existing receivers, first program transmissions were made with a minimum of publicity. Initial tests, starting on November 12th, were conducted without public announcement other than the experimental station identification which was made at 1:00 a.m. The schedule of these tests from November 12th to December 7th was as follows:

Even numbered days of the month - CSSB
Odd numbered days of the month - AM

It was believed that by conducting the tests in this manner, it would be possible to determine from spontaneous public response whether there was a significant difference between the two forms of transmission. If such a difference was disclosed, there would also be indicated which form of transmission was favored. This series of tests was distinctively negative. Only two comments were received from listeners. Both were from persons "skilled in the art" who were aware of the nature of the tests and who presented useful objective observations.

Following these initial tests, a second series of tests were run starting December 8th and terminating January 31, 1958. The schedule of these transmissions was as follows:

Weeks starting December 8th, 22nd, January 5th and 19th -
Amplitude Modulation from 8:00 a.m. to 6:00 p.m. (E.S.T.)
Compatible Single Side-band from 6:00 p.m. to 8:00 a.m. (E.S.T.)

Weeks starting December 15th, 29th, January 12th and 26th -
Compatible Single Side-band from 8:00 a.m. to 6:00 p.m.
Amplitude Modulation from 6:00 p.m. to 8:00 a.m.

The principle purpose of these tests was to provide a maximum of opportunity for the public to observe the basic difference in the two forms of transmission by changing from one form to the other at 8:00 a.m. and 6:00 p.m. This was believed to provide a maximum of opportunity for comparison especially with automobile receivers. As in the case of the first test, there was a very small indication of difference as shown by response from the public. Approximately 30 letters or postcards were received over this period of eight weeks. Several reports commented on the factors that were originally expected, namely, somewhat higher effective signal level with CSSB and in some cases, an improvement in quality. Against these were the reports of those whose receivers were not correctly tuned when the transfer to CSSB was made and who observed a reduction in level and a serious increase in distortion.

Objective tests and measurements were, of course, made to determine that the exciter and transmitter were properly adjusted before starting program tests. These measurements included frequency response, distortion, signal to noise ratio, and side-band suppression. Frequency response of the audio portion of the CSSB unit is shown in Figure 7. Two curves are shown; one made at the start of the program tests and one after their conclusion. The cut-off characteristic at both high and low frequencies is principally the characteristic of the 100 kc filter.

The distortion characteristic for the audio portion of the CSSB unit is shown in Figure 8. These curves are also for two dates covering the period of the program tests. Frequency response and distortion measured through the transmitter and its associated RF monitor are shown in Figures 9 and 10. These two curves were taken at a condition giving 85% modulation at 1000 cycles. For modulations in excess of 85%, the high frequency distortion increased rapidly under the conditions used for these tests. Modulation, therefore, was normally limited to 85-90%. It is understood that this limitation applies only to the prototype equipment used in these tests and that subsequent equipment has been designed to permit full 100% modulation.

Side-band suppression using tone modulation, with 50% modulation at 2000 cycles as reference, is shown in Figure 11. The shape of the lower side-band correlates well with the audio frequency characteristic shown previously. The upper side-band shows a suppression of approximately 30 db at the lower frequencies rising to a suppression of approximately 16 db at the higher frequencies. This lesser suppression at the higher frequencies is probably a necessary compromise to achieve the reduced distortion at the lower frequencies. The effect of this reduced suppression is made less apparent by the spectral distribution characteristic of normal program material which has significantly less energy in the higher frequencies than at 500 to 1000 cycles. Figure 12 shows data taken with tone modulation at 50% showing the distortion products on both upper and lower side-bands. As in the case of the unwanted side-band, the distortion products tend to increase somewhat with frequency.

It is beyond the scope of this paper to discuss the adjustments and different possible combination of compensation to achieve optimum side-band suppression and envelope distortion. It should be clear, however, that the possible variations are many and complex. It is hoped and believed that a practical optimum was achieved and used in the WABC tests.

One of the possible benefits of CSSB, as originally considered, was that of reduced selective fading. Observations were made of this factor at points within the so-called fading zone. This is defined as a zone within which the ratio of sky wave to ground wave or vice versa is 2-1 or less. This zone falls at a radius of approximately 100 miles. As is well known, to have really authoritative results in matters involving propagation, very extensive observations and data must be taken. This was considered neither possible nor necessary in this instance. In any single side-band test, the receiver used, its characteristics and its adjustment are all important factors in results obtained. This is probably the major reason for conflicting reports in comparison of CSSB with AM and is certainly a contributing factor along with variations in propagation to the matter of achieving a definite answer on fading. Such observations as were made of fading, however, indicated there to be a negligible difference between CSSB and AM on WABC's transmissions.

One factor, not originally contemplated, was brought to our attention during these tests. In the operation of television receivers in the home, serious radiation frequently is generated by the deflection circuits of the television receiver creating interference in other portions of the radio spectrum including the broadcast band. This interference can be heard on many channels in the broadcast spectrum at a spacing of 15,750 cycles. One such beat note falls at approximately 771 kc. It was observed by some that the greater side-band energy of CSSB on the lower side-band made it possible, with careful tuning, to achieve a reduction of this interference with CSSB as compared with standard AM.

This interference is probably most objectionable in communities where radio signal strength is low, however, it was observed and reported in the New York area under conditions of relatively high radio field strength. It is to be hoped that this interference will be eliminated or at least greatly reduced in the future by improved receiver design, involving better shielding and less spurious radiation.

The CSSB exciter used in these tests was a prototype and probably contains circuits and components which can be improved. In spite of this fact, the equipment performed well during the WABC field tests. There was only one significant failure involving a component during the tests and frequent re-adjustment for side-band suppression appeared to be unnecessary. Checks made both locally and at the Kahn Research Laboratories indicated side-band suppression of 24 to 31 db depending on adjustment, and type and amount of modulation. No significant extra-band radiation was observed. Modulation percentage as read on the modulation monitor appeared to correctly indicate the modulation envelope with CSSB as with AM.

From the observations on WABC, it is believed that CSSB does make possible a better signal to noise ratio, equivalent to approximately 3 db. It also is believed to be capable of providing better high frequency response without increase in band width. It is believed correct to conclude however that unless the more precise tuning of receivers necessary to these benefits can be achieved either by equipment design or by a campaign of public training, the benefits will be masked by distortion and will not on the average be forthcoming. It may develop that by co-operation between receiver manufacturers and broadcasters, this difficulty can be overcome and a different result with improved service can be achieved.

One possibility in this connection which may not be completely visionary in view of the current interest in stereo-sound is that of simultaneous transmission of separate upper and lower side-bands, one representing the left channel and one the right for 3-D sound broadcasting within the present spectrum.

It is believed also that an opportunity for improved service with the use of CSSB exists on the crowded regional and local channels. Here the use of upper and lower side-band transmission by two adjacent stations on the same frequency might sufficiently reduce co-channel interference and improve service of both stations. This principle might also tend, in some cases, to reduce the directional protection required. It is hoped that the industry may have the benefit of such a test in the near future.

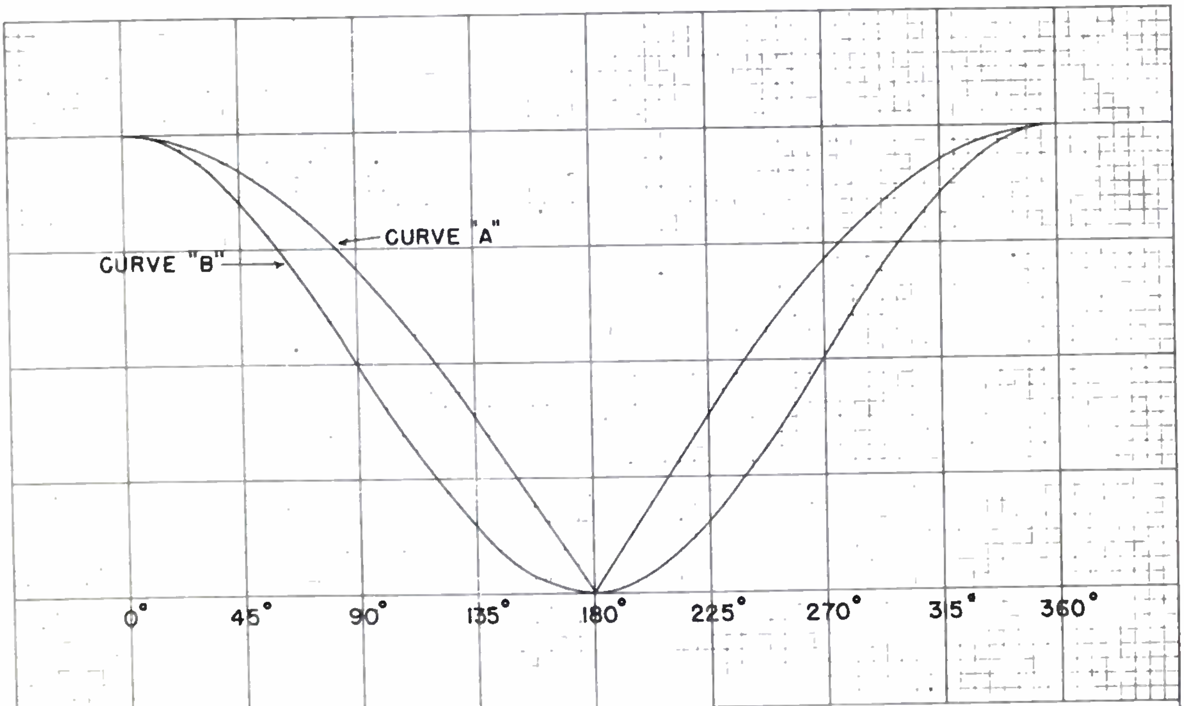


FIGURE-I

AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT

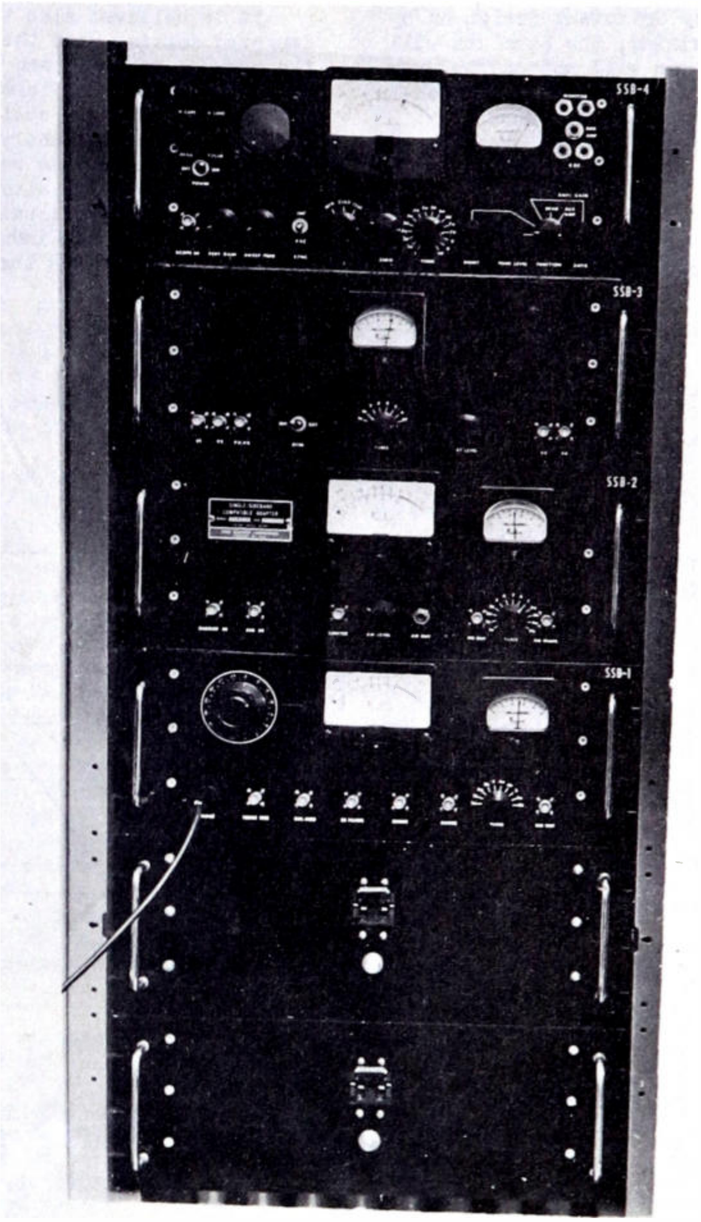


FIGURE-2

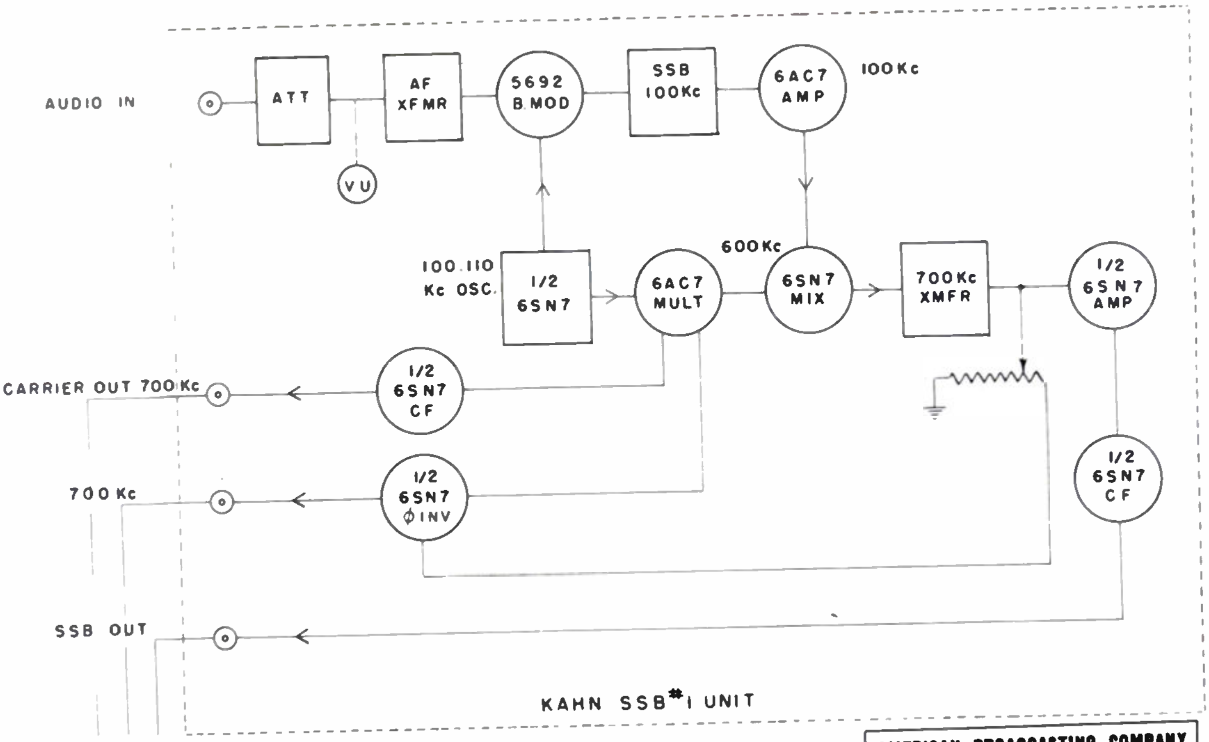


FIGURE-3

AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT

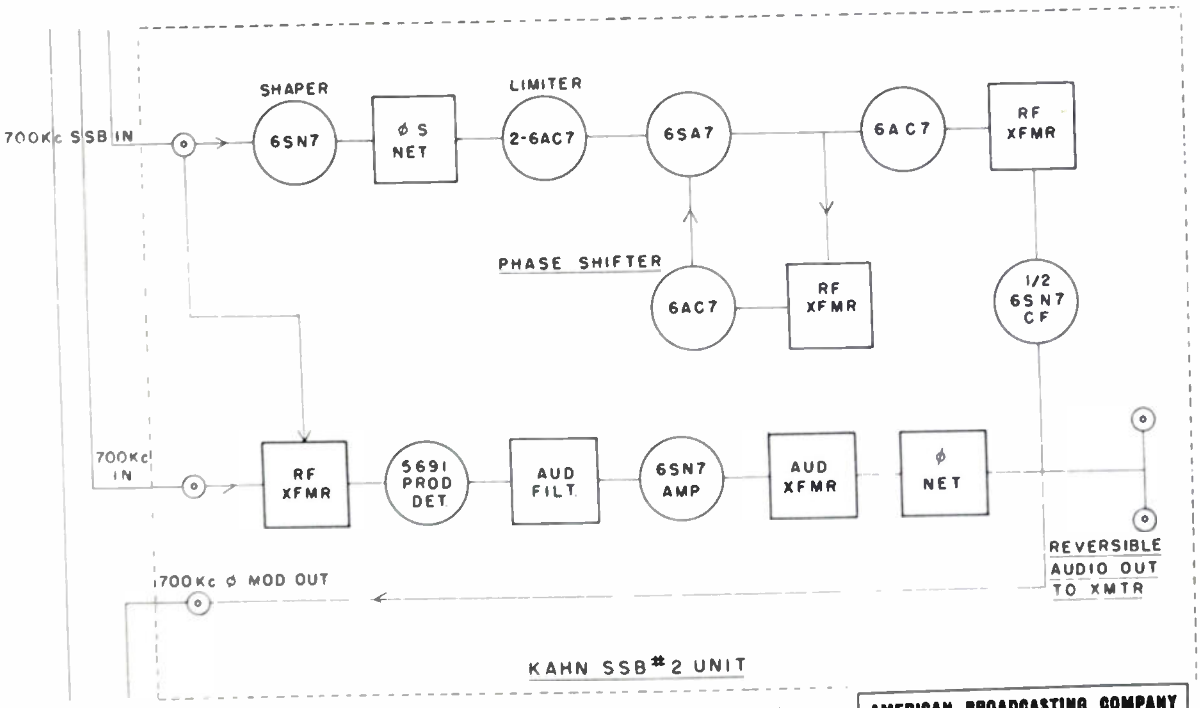


FIGURE 4

AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT

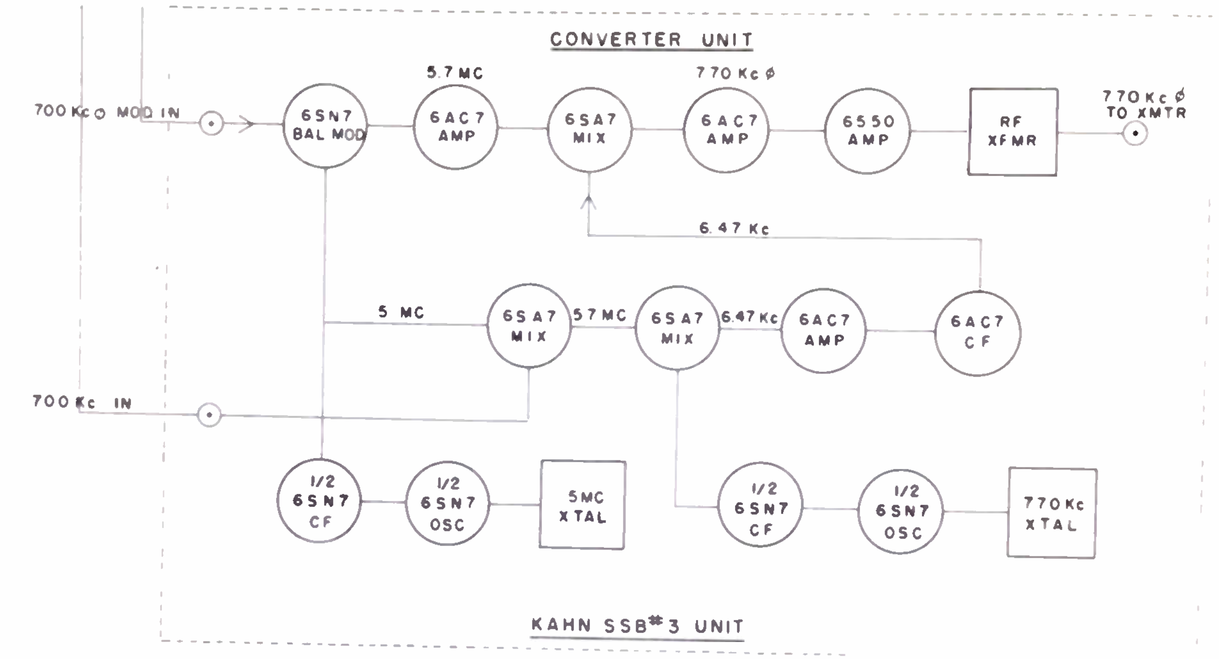


FIGURE 5

AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT

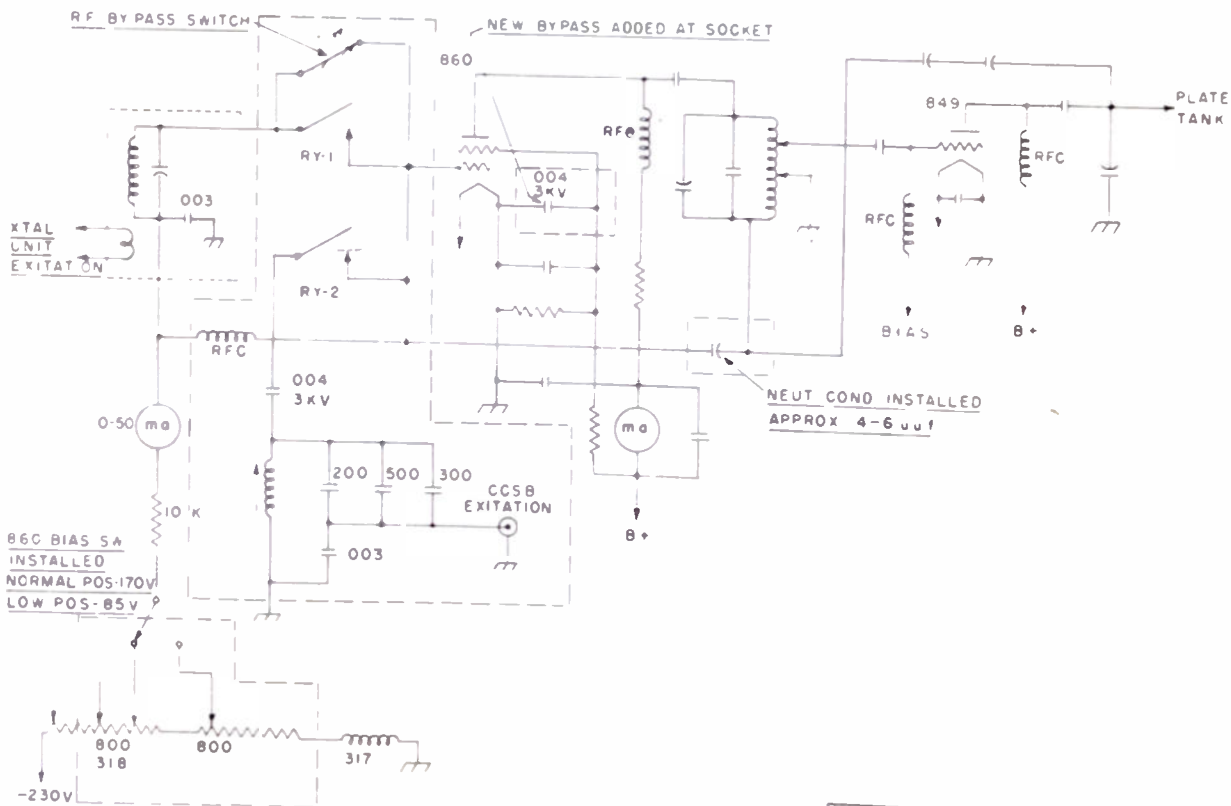
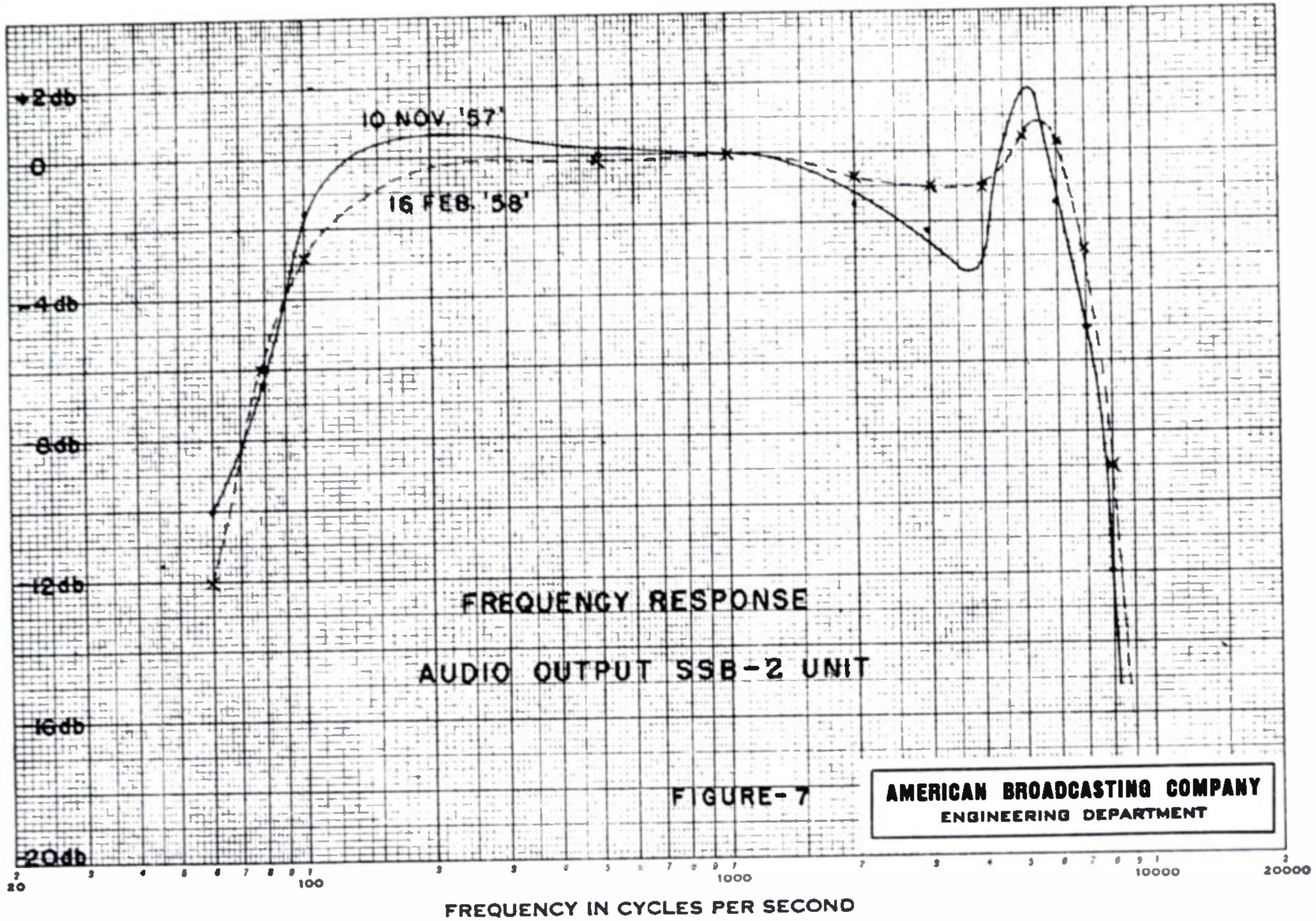
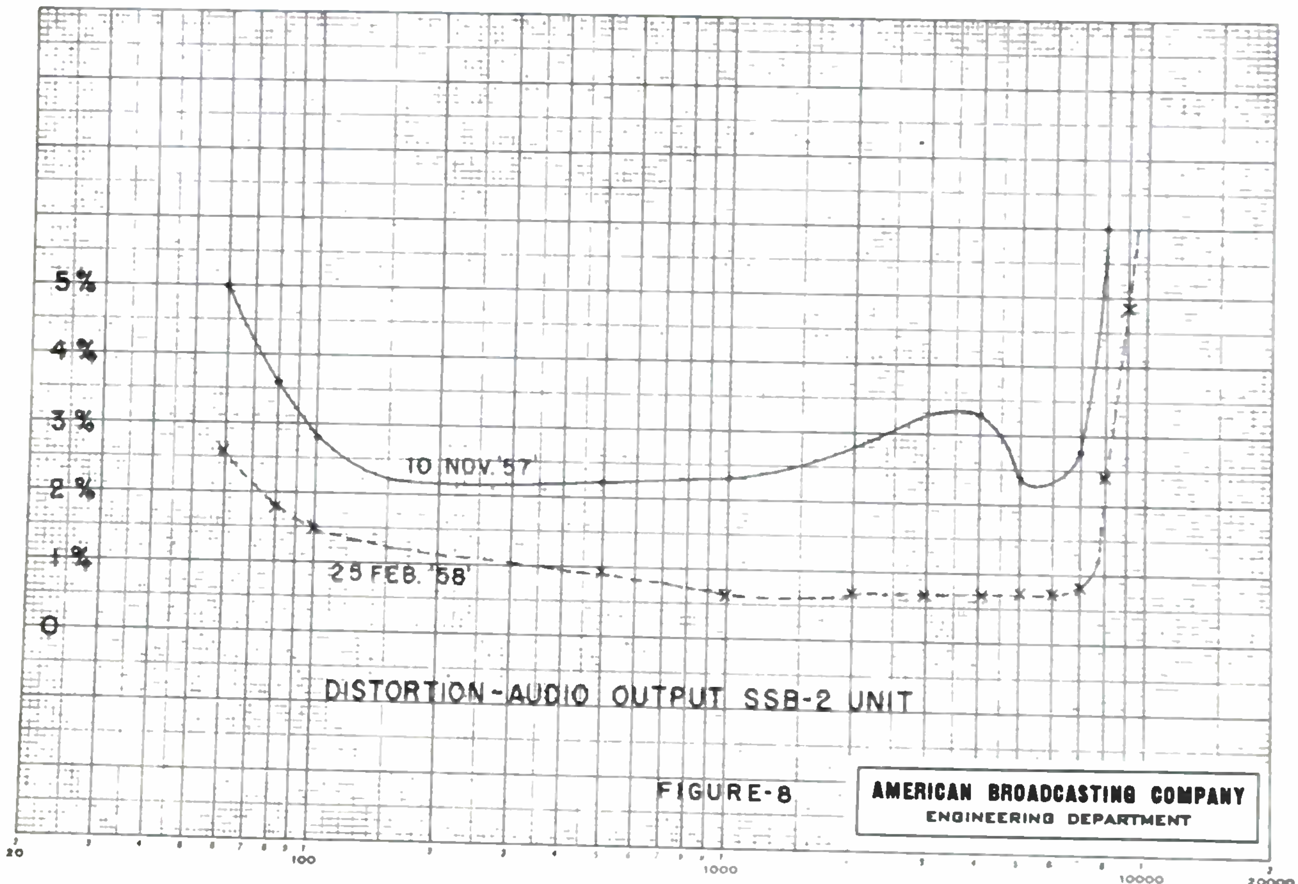


FIGURE-6

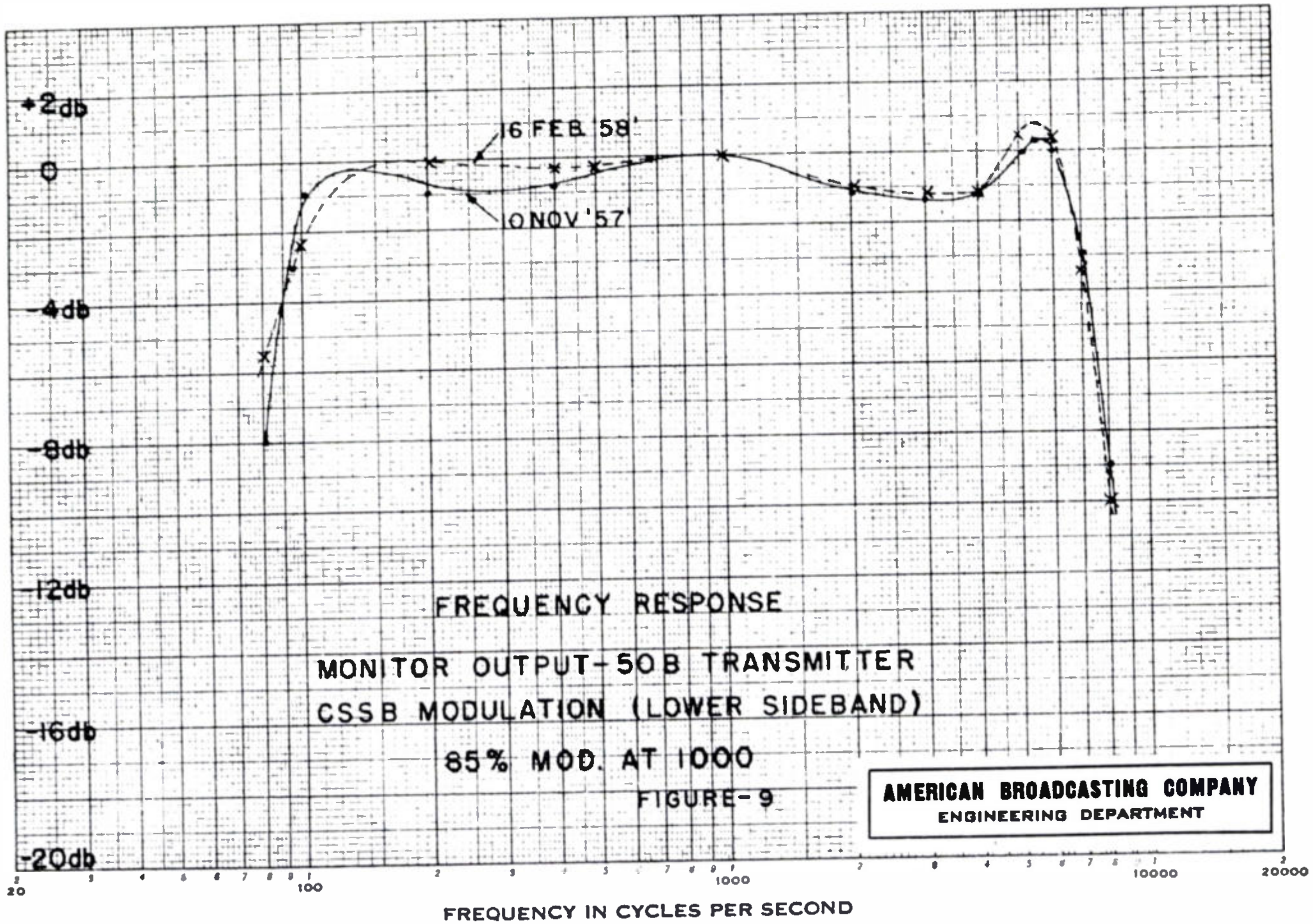
50-B R.F. MODIFICATION FOR CSSB

AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT


+2db
0
-4db
-8db
-12db
-16db
-20db
10 NOV. '57'
16 FEB. '58'
FREQUENCY RESPONSE
AUDIO OUTPUT SSB-2 UNIT
FIGURE-7
AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT
20
100
1000
10000
20000
FREQUENCY IN CYCLES PER SECOND



FREQUENCY IN CYCLES PER SECOND



+2db
0
-4db
-8db
-12db
-16db
-20db
16 FEB '58
ONOV '57
FREQUENCY RESPONSE
MONITOR OUTPUT-50B TRANSMITTER
CSSB MODULATION (LOWER SIDEBAND)
85% MOD. AT 1000
FIGURE-9
AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT
20
100
1000
10000
20000
FREQUENCY IN CYCLES PER SECOND

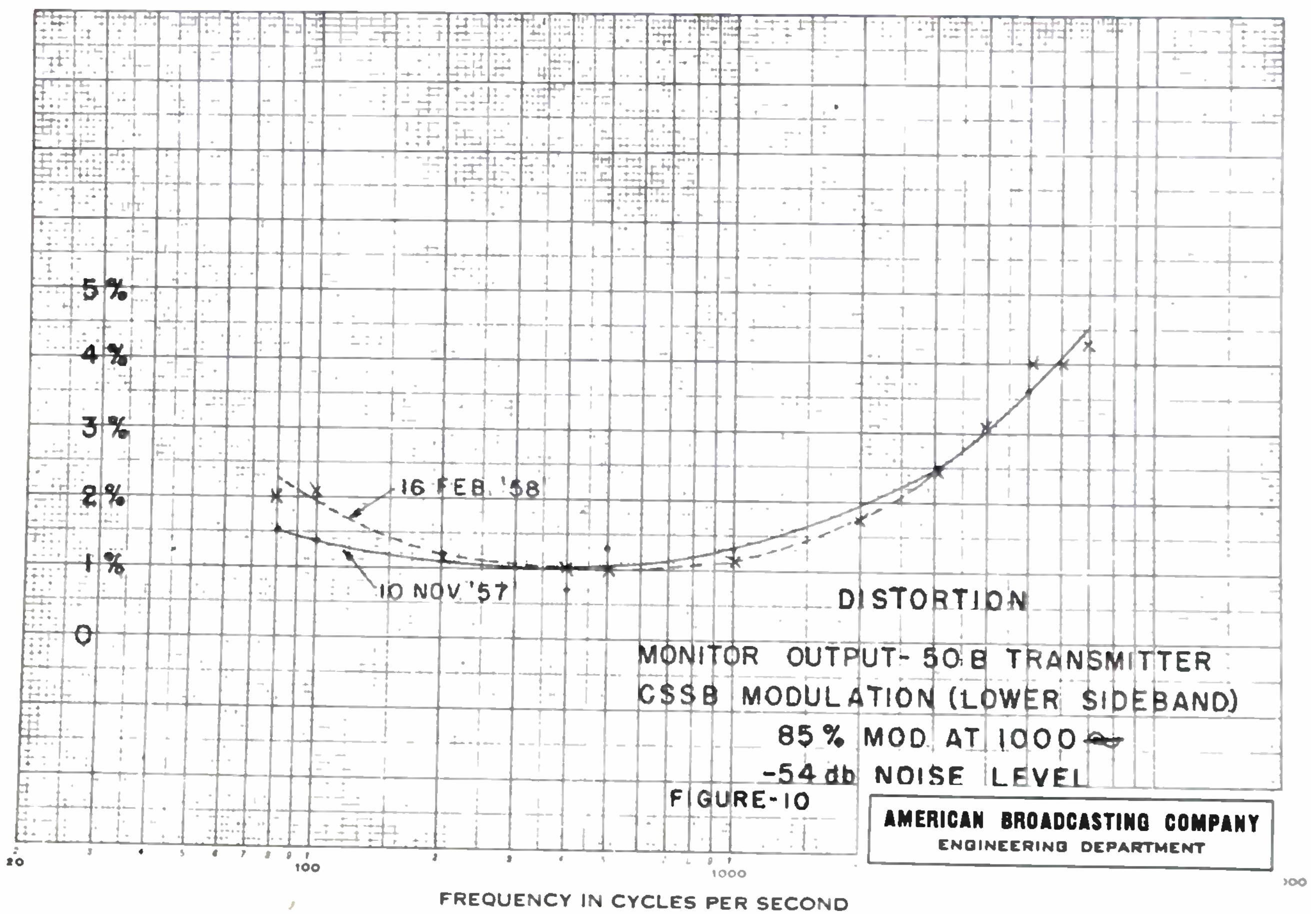
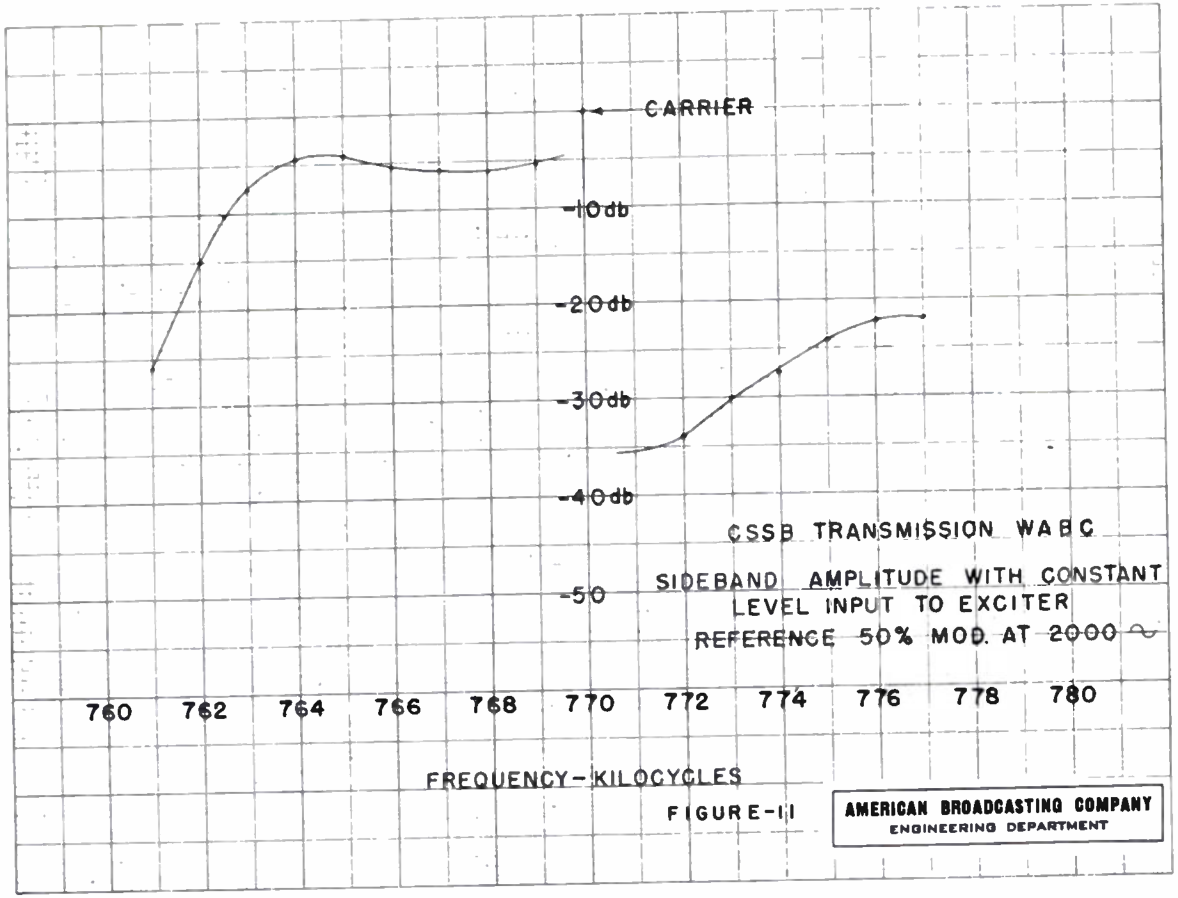

5 %
4 %
3 %
2 %
1 %
0
16 FEB '58
10 NOV '57
DISTORTION
MONITOR OUTPUT- 50 B TRANSMITTER
CSSB MODULATION (LOWER SIDEBAND)
85 % MOD AT 1000
-54 db NOISE LEVEL
AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT
20
100
1000
FREQUENCY IN CYCLES PER SECOND

FIGURE-10



CARRIER
-10db
-20db
-30db
-40db
-50
CSSB TRANSMISSION WABC
SIDEBAND AMPLITUDE WITH CONSTANT
LEVEL INPUT TO EXCITER
REFERENCE 50% MOD. AT 2000 ~
760
762
764
766
768
770
772
774
776
778
780
FREQUENCY-KILOCYCLES
AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT

FIGURE-11

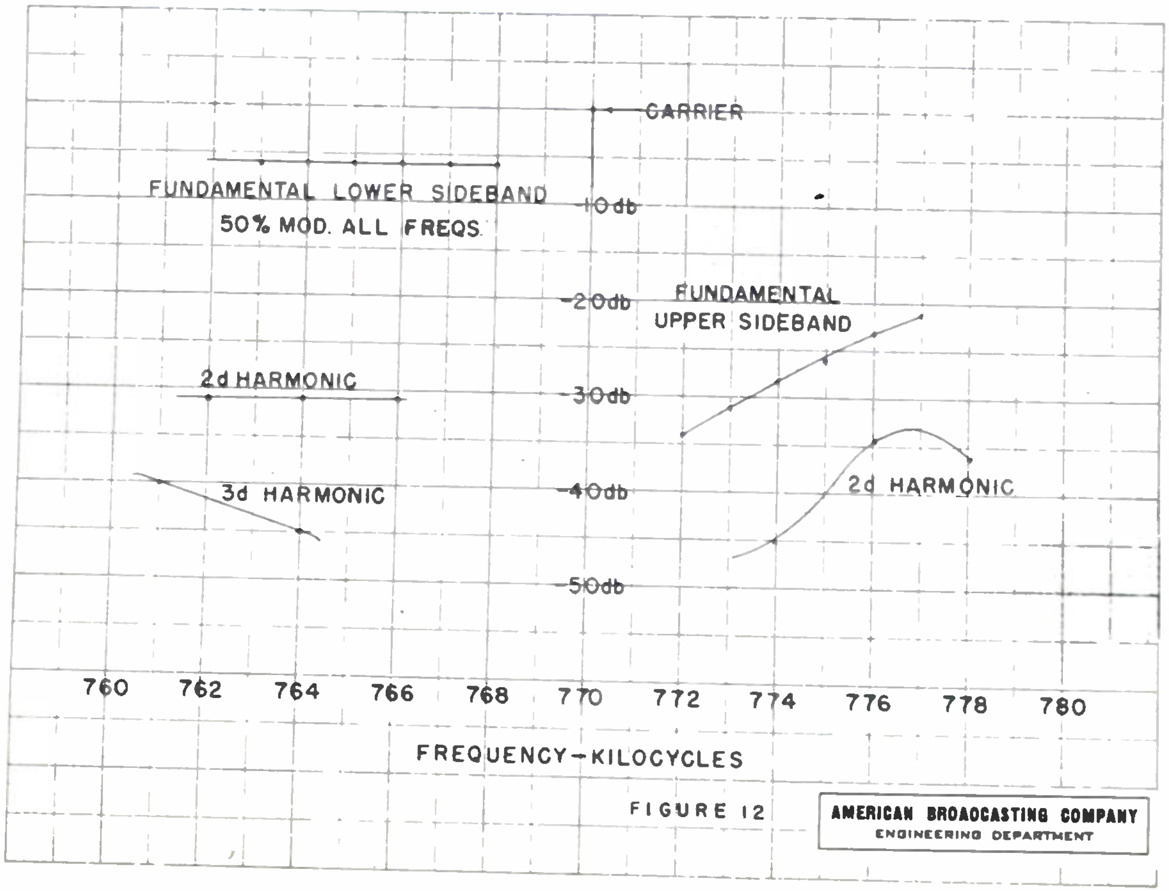

CARRIER
FUNDAMENTAL LOWER SIDEBAND
50% MOD. ALL FREQS.
-10db
-20db
-30db
-40db
-50db
FUNDAMENTAL
UPPER SIDEBAND
2d HARMONIC
3d HARMONIC
2d HARMONIC
760
762
764
766
768
770
772
774
776
778
780
FREQUENCY-KILOCYCLES
AMERICAN BROADCASTING COMPANY
ENGINEERING DEPARTMENT

FIGURE 12

IMPROVED COMPATIBLE SINGLE-SIDEBAND EQUIPMENT FOR STANDARD BROADCAST SERVICE

Leonard R. Kahn
Kahn Research Laboratories, Inc.
Freeport, L.I., New York

Introduction

Compatible Single-Sideband (CSSB) is a new technique for producing a single-sideband wave with an envelope characteristic that is inherently free from distortion. Therefore, its use is compatible with existing AM receivers. Adapters are available to convert any standard broadcast transmitter to CSSB operation, with little or no modifications. Equipment has been recently installed at radio stations WSM and KDKA, and tests are currently being conducted on a regular schedule during commercial hours.

Basic Advantages of Compatible Single-Sideband

1. Signal-to-Noise Gain. The signal-to-noise improvement afforded by CSSB is a function of the receiver used. If optimum receivers are employed for both CSSB and AM, a 3 db power gain will result. However, common, inexpensive table model receivers are far from optimum for double-sideband, and because of their restricted IF bandwidths they are appreciably better suited for CSSB reception. When such receivers are used, a 3 to 5 db power gain can be anticipated.

2. Reduction of Selective Fading Distortion. The second advantage is that CSSB helps reduce selective fading distortion. This is due to the fact that one of the sidebands is greatly attenuated, thereby removing annoying sideband beats. CSSB waves are also less sensitive to the correct phase relationship normally required between the carrier and the sideband for minimum selective fading distortion. Of course, it is quite difficult to measure fading characteristics, but exhaustive domestic tests are underway by different broadcast stations in an attempt to measure the actual amount of improvement offered in commercial broadcast service.

   It has also been noted that certain types of phase distortion produced in the nulls of directional antenna arrays is somewhat improved when Compatible Single-Sideband is used.

3. Reduction of Co-channel and Adjacent Channel Interference. CSSB greatly reduces adjacent channel interference by keeping sideband components on one side of the carrier and more than 10 kc removed from the desired signal. The system, therefore, appears to have considerable potential in easing interference problems now plaguing the broadcaster.

4. Television Receiver Interference. We have recently had an opportunity to analyze a form of interference that has been costing the AM broadcaster serious loss of listeners to other media, such as FM radio, home phonographs, and television. This problem is caused by radiation from TV receivers which in effect place some sixty-eight super power stations on the air. Recognizing the fact that the broadcast band is already extremely crowded, it is clear that there is no room for these stations, especially when each one effectively covers the entire nation.

   In order to determine the severity of this interference, we recently mailed a number of questionnaires to individuals selected at random from various New York metropolitan and suburban area telephone books. Out of the 1,500 questionnaires sent, 62 were returned because of change of address, 10 did not have radio receivers, 205 stated that television interference was not objectionable, and 211 said it was objectionable. Thus, slightly more than 50 per cent of the people who answered these questionnaires indicated that television receiver interference was a serious problem.

We have also checked with various radio repair shops and have been informed that they have received numerous calls to repair receivers that were simply picking up television receiver interference. They also complained that many of the sets repaired were returned by their customers as being improperly serviced because interferring whistles still persisted.

CSSB, however, permits this interference to be greatly attenuated, or virtually eliminated. Recognizing that television deflection circuits generate strong harmonics spaced 15.75 kc apart throughout the broadcast spectrum, signals will occur either above or below a given transmitter frequency and produce an audible interference. Since CSSB concentrates energy on only one side of the carrier the upper or lower sideband, whichever is free from interference, can be selected for transmission, permitting the listener to tune away from the interfering beatnote.

5. Improved Audio Fidelity. The full bandwidth transmitted by double-sideband AM cannot be fully utilized by conventional receivers because of the narrow bandwidth IF amplifiers employed. CSSB waves, however, occupy only one-half of the normal AM spectrum, thereby permitting these amplifiers to be used more efficiently. This fact causes a noticeable improvement in the audio fidelity of conventional home and automobile receivers.

It has been demonstrated by Mr. J. C. R. Licklider of Harvard's Psycho-Acoustic Laboratory, Bell Telephone Laboratories, and others, that the audio fidelity of a system is an extremely important factor in maintaining high intelligibility. CSSB, therefore, provides a significant improvement in marginal reception areas where static and interference would normally cause conventional AM reception to be completely unintelligible.

Description of Equipment

The first slide shows a simplified block diagram of the CSSB unit. It will be noted that a single-sideband wave is generated in a normal SSB generator using balanced modulators and a high quality crystal filter. The output of the generator is a full carrier SSB wave which is passed through a special amplitude limiting device which isolates the phase modulation component. This wave is then amplified in Class-C amplifiers and finally feeds a modulated stage.

The desired amplitude modulation component is obtained by electrically multiplying the carrier by the single-sideband wave to produce a distortion-free audio wave. It is then amplified and used to modulate the phase modulation component. The resultant is a SSB wave having an undesired sideband component of 30 db or more below the desired sideband, including all third order intermodulation distortion products.

Slide 2 shows the Standard Model CSSB-55-1A Broadcast Adapter. The bottom three units are electronically regulated power supplies. The unit with the sloping panel is a special test unit which greatly simplifies equipment operation. It is, in effect, a SSB receiver which allows switching from the desired to undesired sideband so that relative levels between the two can be instantly compared on a direct reading DB meter.

New Soviet SSB Development

It is interesting to note that the Soviets have recently developed a system aimed at obtaining the same advantages as CSSB. They call the system optimum modulation and we understand that it is to be tested soon in Kiev on a standard broadcast station. The optimum modulation method results in a signal with approximately 4 per cent distortion and an undesired sideband level of 18 db. An even more serious drawback is that the undesired sideband component is spaced from the carrier by two times the audio frequency being transmitted. CSSB, on the other hand, transmits an undesired sideband of approximately 30 db which is spaced from the carrier by only the audio frequency being transmitted.

Furthermore, distortion is less than one percent as measured on standard equipment. Overall transmitter and CSSB adapter distortion measurements show approximately the same distortion figures as measured on the transmitter alone.

CSSB Operational Tests

The first Compatible Single-Sideband test was made with the Voice of America's megawatt transmitter located in Munich, Germany, wherein a 4 megawatt peak envelope power CSSB was produced. This system has since been in continuous service for about two years.

Preliminary domestic tests with a prototype Compatible Single-Sideband system were also made at radio station WMGM and later at WABC, both of which are located in New York. But due to the limited time made available by station management for specific tests, the results obtained were more of an operational nature than an engineering evaluation of the system.

Although WABC tests confirmed earlier WMGM reports of the system's 3 to 5 db effective increase in signal strength and improved audio fidelity, information on fading characteristics, as well as a detailed analysis of interference improvement, was generally limited in scope. It is believed that proper use of A-B type comparison tests between AM and CSSB modes of operation would have added considerable significance to the results. However, they were not permitted, except during the scheduled switch-over times at 8 a.m. and 6 p.m. EST. Because of wave propagation at these times, measurements for improvement in selective fading distortion at the fading zone test site were of little value. In addition, selective fading on conventional AM was generally absent in normal fading zones due to prevailing sun spot cycle activity. Therefore, further domestic tests of the system are underway.

One conclusive result obtained from the WABC tests concerns the reduction in adjacent channel interference. It was noted that when WABC was transmitting in the CSSB mode, WBBM in Chicago, which is only 10 kc away, could be clearly heard in the New York metropolitan area. During this period, WABC was operating on 770 kc on the lower sideband, providing maximum protection to WBBM transmitting on 780 kc. In spite of the small frequency separation between the two stations and the fact that WABC was, of course, much stronger in the New York area, the latter did not interfere with WBBM. When WABC switched to conventional double-sideband transmission, however, WBBM was almost completely covered by interference. Thus, CSSB successfully demonstrated an ability to reduce adjacent channel interference, even in this rather extreme case.

During the WABC and WMGM tests, the undesired sideband was maintained at 25 to 32 db below the desired sideband, including the measurement of all radiated intermodulation products. Listening tests were conducted with many types of receivers, including automobile, AC-DC, transistor, high fidelity and communications models. In all instances reception was totally free from inherent distortion and complete ease of tuning was thoroughly demonstrated. When receivers were tuned to the carrier or desired sideband, no distortion was evidenced, however, when receivers were tuned to the undesired sideband, an appreciable reduction in signal strength was apparent. It should be noted that during the WABC test periods, when transmitters were switched between CSSB and AM modes of operation, receivers were not necessarily tuned in each case for optimum reception. Since no mention of this was made over the air, listeners did not retune their sets for each transmission mode. Of course, when CSSB is operated on a continuous basis, listeners would not have to retune their sets. They would tune to each equipped station only once for the loudest and clearest sound, which is optimum for both CSSB and conventional AM reception. Thus, no special type of receiver or precise tuning is needed to obtain the maximum benefits of CSSB.

It is expected that engineering tests currently under way on production model CSSB installations at KDKA and WSM will substantiate the above findings and will provide the type of detailed technical evaluation needed.

Conclusion

It has been shown in actual operation that Compatible Single-Sideband offers the user significant reductions in interference, double to triple the effective transmitter power with average receivers and improved audio fidelity. Because of CSSB's unique design and compatibility feature, the many advantages of single-sideband transmission may now be fully realized by the radio broadcast industry for the first time.

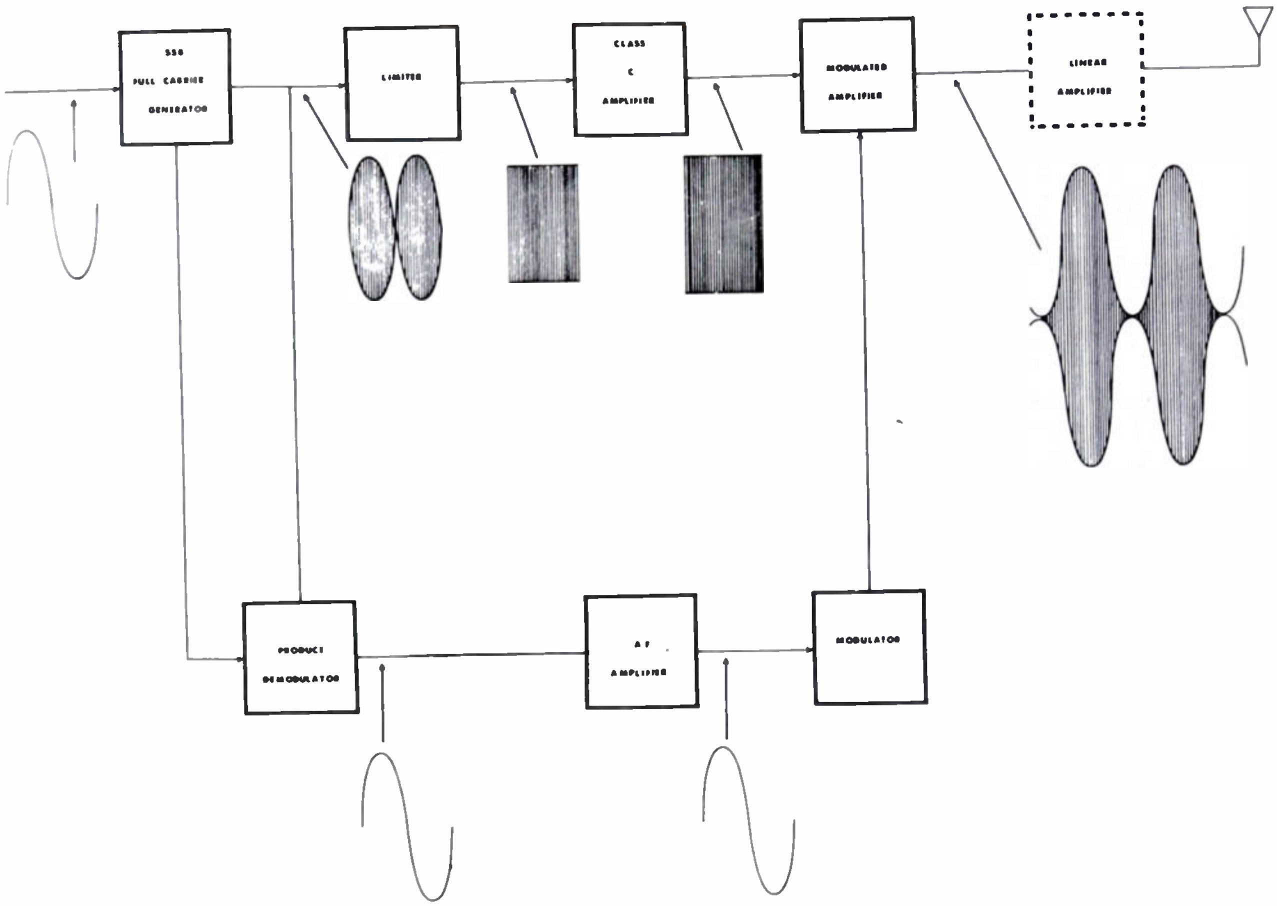
Acknowledgments

I would like to take the opportunity to acknowledge the efforts of others at Kahn Research Laboratories, Inc. in designing and developing the Compatible Single-Sideband broadcast adapter. Messrs. Lloyd A. Ottenberg, Robert C. Rogers and Dr. Leonard O. Goldstone all made important contributions in the design of the CSSB adapter. Also, Mr. Kenneth B. Boothe is to be thanked for his assistance in preparing this paper.

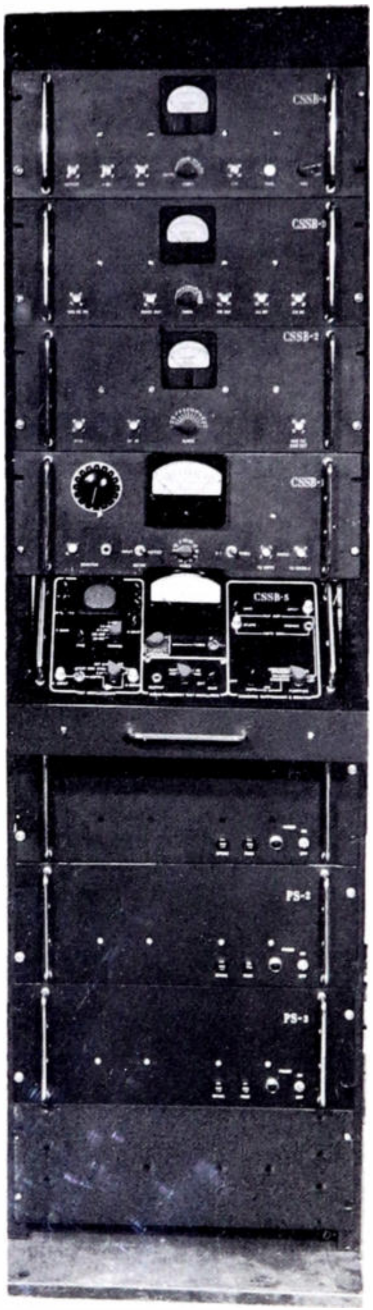
TABLE I

Chart Shows Typical Frequency Distribution Of Television Interference In The Radio Broadcast Spectrum

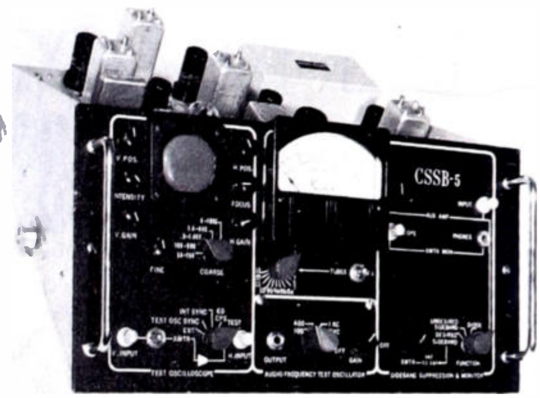
Interference Frequency	Interference Frequency
551.25	1086.75
567	1102.5
582.75	1118.25
598.5	1134
614.25	1149.75
630	1165.5
645.75	1181.25
661.5 (42nd harmonic)	1197
677.25	1212.75
693	1228.5
708.75	1244.25
724.5	1260
740.2	1275.75
756	1291.5
771.75	1307.25
787.5	1323
803.25	1338.75
819	1354.5
834.75	1370.25
850.5	1386
866.25	1401.75
882	1417.5
897.75	1433.25
913.5	1449
929.25	1464.75
945	1480.5
960.75	1496.25
976.5	1512
992.25	1527.75
1008	1543.5
1023.75	1559.25
1039.5	1575
1055.25	1590.75
1071	1606.5 (102nd harmonic)



Slide 1 - Block diagram of basic compatible single-sideband system.



Slide 2 - The model CSSB-55-1A adapter system for AM broadcast transmitters.



Slide 3 - The multi-purpose, self-contained test unit supplied as a component part of the Model CSSB-55-1A adapter. It is a typical illustration of attention given to system design from an operational point of view. The unit incorporates a new, simplified technique for accurately measuring in db the desired and undesired sidebands on a direct reading meter. These measurements are possible because the equipment includes a special single-sideband receiver with its own crystal filter. Thus all undesired sideband components are quickly measured without computations along with third order intermodulation distortion products. A small monitor oscilloscope in the unit not only features normal sweep, but permits measurements of Lissajous and trapezoidal patterns of the associated transmitter. The test unit also contains a utility audio amplifier and a built-in audio oscillator, providing fixed test frequencies at 100, 400, 1,000, and 5,000 cycles.

RIAA ENGINEERING COMMITTEE ACTIVITIES WITH RESPECT TO STEREOPHONIC DISC RECORDS.

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Columbia Records, Inc.
New York, New York

ABSTRACT

Many engineering problems arise with the introduction of stereophonic disc phonograph records. Considerations of the type and depth of modulation, groove shape, compatibility, etc., should be resolved to standards, so that an optimum system will be offered to the public. This paper outlines the efforts of the RIAA Engineering Committee in this direction.

THE WESTREX STEREODISK SYSTEM

By C. C. Davis and John G. Frayne
Engineering Division
Westrex Corporation
Hollywood, California

Abstract

The paper describes the Westrex StereoDisk Recorder which records two stereophonic channels in a single groove with a single stylus. The axes of the two recordings are at 90° to one another, each being at 45° with the horizontal plane of the record. By use of appropriate input circuits, the vertical-lateral type of stereo record may also be recorded. The recorder utilizes the electrodynamic feedback features of the Westrex lateral recorder. The design features of the recorder are described and the performance is discussed. Data on channel cross-talk, intermodulation distortion and frequency characteristics are given.

The design features and performance of the complementary StereoDisk Reproducer are described, including the desirability of maintaining the same vertical tracking angle in both recorder and reproducer. This reproducer employs dual d'Arsonval movements resulting in an exceptionally faithful reproduction.

Introduction

The current interest in stereophonic sound reproduction in the home has stimulated the development of the two-channel stereophonic disk system described in this paper. It is not generally recognized that considerable research was carried on in the 1930's on this type of stereo recording in which two channels are recorded in a single groove. For example, Blumlein of EMI in England was granted British patent No. 394,325 in 1933 on a two-channel stereophonic system. He disclosed the possible use of a vertical-lateral combination to produce a stereophonic effect. He also called attention to the 45-45 orthogonal system as an alternative method. At the Bell Telephone Laboratories, Keller and co-workers made stereophonic recordings as early as 1936 and Keller & Rafuse were granted U.S. Patent No. 2,114,471 in 1938. Blumlein and Keller built their experimental stereophonic recorders by linking existing lateral and vertical recording mechanisms. U.S. Patent No. 2,025,388 issued to Henning of BTL disclosed a reproducer capable of reproducing either vertical or lateral recordings. By using suitable output circuitry, Henning's reproducer has been successfully used to reproduce the Keller & Rafuse vertical-lateral recordings. The emphasis on defense work at the BTL in the late '30s, coupled with apathy on the part of commercial recording concerns, caused this development to be laid aside.

In developing the Westrex StereoDisk System, the authors were primarily interested in the development of a stereophonic system in which the two channels would give as nearly identical quality as was possible to achieve. In arriving at this conclusion, they were mindful of the work of Pierce & Hunt[1] in their comprehensive theoretical analysis of the lateral and vertical modes of disk recording in which they predicted the amount and type of distortion products obtained by tracing both lateral and vertical grooves. While this and other considerations enumerated below were important factors in support of the 45-45 orthogonal system, the symmetry of design it made possible for both recorder and reproducer was probably the determining factor. Further, the 45-45 system was capable of producing the vertical-lateral type of stereophonic disk if this method had been adopted by the industry.

General Description of 45-45 System

Having decided in favor of the 45-45 system, the next step was to evaluate design objectives in both recorder and reproducer in order to obtain a satisfactory stereophonic recording system for disks. A primary design objective in a dual channel stereophonic system is to obtain identical characteristics in the two channels. In the case of a disk recorder this means the stylus mounting mechanism must have equal compliance in all directions in a vertical plane normal to the recorded groove. If the recorder is to use moving coils, they should be identical and they should operate in identical amounts of magnetic flux. The coils should move the stylus through identical mechanisms. If this primary objective is not met, one channel must necessarily be inferior to the other in one or more respects requiring an unnecessary amount of effort to adjust and maintain the system for reliable commercial recording. It can be stated that consideration of this basic objective was an important factor in the adoption of the 45-45 method of recording. The final recorder design met the objective and also was capable of recording a vertical-lateral type of record while preserving the symmetry called for above.

It was decided that the new recorder should retain the feedback principle which has been an outstanding feature of both the Western Electric and Westrex recorders for the past 20 years and has been described previously[2]. This involved the mounting of a pickup coil on each of the drive assemblies in the recorder and including these coils in the feedback loops in the respective recording amplifiers. The design constants of the feedback circuit were of necessity a compromise of factors imposed by the design and performance requirements. The details of the feedback circuit are covered in the description of the recorder.

As a result of these considerations a feedback of approximately 27 db at the resonant frequency was employed. This held up the high-frequency response to at least 15 kc and the low-frequency response was brought up by the use of passive networks preceding the recording amplifiers. This arrangement of feedback maintained good stability over the significant range of audio frequencies, and, with the aid of the passive networks, essentially constant-velocity recording was provided over a frequency range from 30 to 15,000 cps, except for a very narrow dip of a few db in the vicinity of 11,000 cps. This characteristic is modified in commercial practice by inserting the RIAA or equivalent recording pre-equalization characteristic.

Two recording amplifiers, one for each channel, are required to operate the recorder. The circuits of the amplifier are shown in Figure 1 in block-schematic form since the details of design are not considered significant for this presentation. The recording circuit consists of a phase-inverter stage which drives a push-pull gain stage which in turn drives a push-pull, parallel power stage. The four-ohm output goes to one drive coil in the recorder. Adjustable low-frequency and high-frequency equalization is provided ahead of the phase-inverter stage. The output of the feedback coil in the recorder goes through a gain control to a gain stage. The output of this stage goes to the grid of the phase-inverter tube which also acts as a mixer. The output of the feedback coil also goes to a two-stage monitor amplifier which has a 600-ohm or 50-ohm output at a nominal monitor level of -15 dbm. The input network of the monitor amplifier provides the RIAA reproducing equalization.

Since the characteristic of the network used to provide constant-velocity recording is to some extent complementary to the RIAA pre-equalization recording characteristic, the two characteristics can be combined in a single equalizer with a considerable reduction of insertion loss.

The development of a reproducer was undertaken with the design objective of providing the highest quality of reproduction attainable. The three well-known basic principles of reproducer design were considered and the electrodynamic or rotating-coil type was selected as affording the best opportunity of meeting the design objective. Theoretical consideration indicated this type would provide appropriate compliance and low mass at the stylus compatible with basic requirements. It was indicated also that a satisfactory frequency response over the audio range with relatively low intermodulation and harmonic distortion might be expected. The details of design and performance of the reproducer are described elsewhere in this presentation.

Comparative Performance of V-L and 45-45

A comparison of the relative performance capability of the V-L and the 45-45 systems will indicate the reasons for the preference of the latter system. An analysis of the stereophonic groove cut with either the 45-45 or V-L system will show considerable similarities between the two types of recording. Thus, with the 45-45 system, both lateral and vertical modulated grooves will result depending on the phase relationships of the inputs. For random phase differences, a complex groove with both vertical and lateral components will be recorded. On the other hand, with the V-L system, vertical and lateral tracks will result only when one or the other microphone receives no signal. For the general case where sound is picked up by both microphones, the resultant groove will have both vertical and lateral components and will appear to be substantially the same as a groove cut with the 45-45 system.

In the V-L system, we might expect a preponderance of either vertical or lateral cut grooves while in the 45-45 we might expect a preponderance of 45° modulations. Thus, the limitations on groove depth imposed by recording vertical components should be more severe than in the 45-45 system. In addition, by deliberately phasing the channels in the 45-45 system so that in-phase signals are recorded laterally, the limitations imposed by the vertical components are further reduced. Because of this situation, we would expect greater quality differences between the two channels in the V-L system due to the inherent but different tracing distortion components in the vertical and lateral grooves[1].

The symmetry of the 45-45 system is not only of importance in balancing the quality of the reproduction from the two grooves, but has certain other very important by-products. In the record changer, the vertical component of rumble is generally more pronounced than the horizontal component which would mean more audible rumble in one channel in the V-L system which is highly undesirable from a listening standpoint. In the 45-45 system, rumble should be identical in both channels and lower by about 3 db than in a strictly vertical channel. Another by-product favorable to the 45-45 system is the ease of balancing the reproducing channels in the home. It is only necessary to play a standard lateral record and adjust the channel gain for equal loudness.

In the 45-45 system each channel will have lateral and vertical components of the impressed modulation. Thus, it should be possible to reproduce either the lateral or vertical components of both channels with a standard lateral or vertical reproducer. Since vertical-type pickups are rarely encountered in home reproducing systems the lateral type of pickup only needs consideration. If this type of reproducer has good vertical compliance it should be able to reproduce satisfactorily the lateral components of both channels. In this sense the 45-45 system of stereophonic recording is compatible with lateral records. This cannot be said of the V-L system since the lateral component represents only one of the recording channels.

Design Details of Recorder

Figure 2 illustrates the basic design of the recorder. The magnetic gaps of the drive and feedback coils are arranged in a series parallel fashion. The magnetic flux is provided to the system by a single magnet made from Alnico V D.G. This arrangement of magnetic paths insures equal flux densities in the corresponding gaps. Each of the magnesium coil forms contains a drive and feedback coil as shown. The shaded areas between the magnetic gaps indicate copper slugs or shields which reduce inductive cross-talk from the drive coils to the feedback coils. This cross-talk must be minimized to utilize the advantages of negative feedback control of the drive coil motion. Figure 3 is a bottom view of the recorder. The individual coil assemblies are mounted on removable subassemblies, and are shown together with the other principal components in Figure 4. This permits their alignment in an assembly jig as self-contained units before installation. The coil forms have small terminals which are in turn connected through pigtails to corresponding stationary terminals on the assembly. This arrangement eliminates breakage of these vital connections to the coils due to vibration and it simplifies examination or service of these parts. The coil-supporting springs are made from beryllium copper and are "V" shaped to maintain the coils in proper alignment. The coils drive the stylus supporting member through separate linkages. These linkages consist of wires braced along their midsections to prevent excessive lateral compliance. The result is a relatively stiff driving system in a forward direction but with relatively high compliance in a lateral direction. This important relationship avoids the necessity of one coil bearing the mass of the other coil which would result if the coils were rigidly mounted on the stylus supporting member. Thus the mass of each driving system includes only its components and a negligible portion of the other driving system.

A tubular cantilever spring was chosen for the stylus supporting member because of the many advantages it offers. Its compliance as a cantilever is inherently the same in all directions in the desired plane, which is essential in order that the complex motion of the stylus may present uniform impedance in any direction in the vertical plane. This is particularly important at those portions of the frequency spectrum where the negative feedback voltage has little control of the stylus motion. Another feature of a tubular cantilever is its relatively low rotational compliance which reduces the tendency toward cross-talk from one channel to the other.

The damper shown on the stylus supporting member has little or no effect below 10 kc and very little effect above 10 kc on the actual recording. It is used mainly to smooth out several peaks and valleys in the monitor output reading. Therefore, the effects of temperature and other factors upon the damper in no way affect the system damping which is, of course, supplied by the feedback coil and the associated negative feedback loop in the recording amplifier.

The stylus has a tapered shank and is of the same type as those used in the 2B Recorder. The included angle of the stylus has only second-order effects on the recording. The same types of hot wire terminals are used to record with a heated stylus. Inasmuch as it appears impractical to hinge the stylus mounting mechanism of a recorder or reproducer exactly at the surface of the record, a vertical tracking error will be introduced unless very nearly the same angle is used in the recorder and reproducer. What is normally considered as "vertical" motion of the cutting stylus tip in the recorder is actually at an appreciable angle from a vertical line. It does not seem practical from the standpoint of construction of both recorder and reproducer for the situation to be otherwise. The angle for the recorder is nominally 23° in a direction such that with upward stylus motion, there is a component in the direction of travel of the record surface. This angle was determined by measurements on an actual stylus supporting member and a mock-up stylus, and was confirmed by calculations.

Description of 45-45 Groove

The geometry of the groove cut by the 45-45 recorder with a stylus having a cutting angle of 90° merits consideration in some detail. Figure 5 shows schematically cross-sections of the groove for four limiting conditions. The upper left section shows the type of groove which will be cut when signal is fed only to the left channel. The right-hand groove wall is a slant line of varying depth. The right edge of the groove at the record will be a smooth line without modulation. The left edge of the groove will vary in width in accordance with the signal. The upper right section of the figure depicts the opposite condition in which signal is fed only to the right channel. It will be noted that the conditions are just reversed. The lower left section shows the type of groove which is cut when identical signals are fed to the two channels and are in phase at the drive coils. Under this condition a vertical recording is obtained. The lower right section shows the condition when like signals are fed to the two channels but out of phase at the drive coils. Under this condition a lateral recording is made. Any combination of these four conditions in varying amplitudes may occur in recording stereophonic program material.

Figure 6 is a photomicrograph of grooves recorded first with a single frequency on one channel and then on the other. One side of each groove is essentially straight while the other varies in width in accordance with the lateral component of the signal. The vertical component of the signal is shown by the varying shading in the photograph. Figure 7 is a photomicrograph of grooves recorded using program material and a stereophonic pickup. Here again the lateral components are shown by the varying width of the grooves while the vertical components are represented by the varying shading in the photograph. Some of the vertical components in this picture represent frequencies of 12 kc or higher.

Figure 8 shows the maximum excursion of a groove which will be cut with a given maximum amplitude of modulation of the drive coils and with a specified minimum depth of groove. The specified minimum groove depth is D and the maximum modulation amplitude of either coil is A. Under these conditions the maximum horizontal excursion of groove will be $2D + 4\sqrt{2}\,A$ and the maximum depth of groove will be $D + 2\sqrt{2}\,A$.

The relative output level from a groove of specified maximum width recorded with the 45-45 system is a matter of considerable interest. Figure 9 shows a comparison of this type of groove with the conventional lateral groove. In each case the specified maximum groove excursion is $2D + 2A'$. In the lateral groove the maximum amplitude of modulation is A which is equal to A'. In the 45-45 groove the amplitude of modulation in either channel is $\frac{A'}{\sqrt{2}}$.

Accordingly the output of each of the 45-45 channels will be down 3 db with respect to the output of the lateral channel. For these conditions the playing time will be the same with the

45-45 system as for the standard lateral recording.

Performance of the Recorder

Fundamentally the performance of the recorder is based upon the use of moving coils with negative feedback control. The use of feedback to control resonance effects in vibratory systems was suggested first by Maxfield and Harrison[3]. The advantages of this type of transducer have often been described[2,4]. The moving coil offers a maximum of motional linearity and when combined with negative feedback it comprises a stable well damped system without resistive losses[5]. However, high efficiency must be obtained to provide the recorded levels used currently without excessive power losses which might destroy the coils with excessive heat. In order to achieve a practical physical size, the component parts of the magnetic circuits of the dual channel recorder must occupy a limited amount of space. To offset this condition two design features were incorporated.

(a) Minimum mass of the drive and feedback coils was used.

(b) The natural resonant period of the vibrating system was made relatively high.

Since it is unnecessary to generate sufficient feedback voltage to control the system throughout the entire audio band if equalization is used at one or both extremes, the physical size of the feedback coil and its operating flux may be reduced. Thus the mass of the vibrating system may be reduced and the relative amount of flux at the drive coil increased.

The natural frequency of the vibrating system is placed about one octave above the geometric mean frequency. This results in power economy in the high-frequency range where considerable power is normally required with the RIAA type of pre-equalization. The amount of feedback at resonance is approximately 27 db. This is sufficient to control the stylus throughout the range of frequencies where the mechanical impedance of the vibrating system is low. This is the range where variable cutting resistance presented at the stylus may cause non-linearity or wave distortion. Below this range the system is controlled by stiffness and is unaffected by cutting conditions. Throughout this range pre-equalization is required due to the loss of feedback control. Pre-equalization is not needed at high frequencies for a constant velocity recording characteristic.

The copper shields between the drive and feedback coils are intended to prevent inductive cross-talk between the drive coil and feedback coil. The effectiveness of these shields is at a minimum at about 200 cycles. Here the inductive cross-talk is -50 db relative to perfect coupling (i.e., the coils used in a perfect transformer). Since the feedback voltage generated by the velocity of the coil is already low, the inductive cross-talk and the feedback voltages are of the same order of magnitude. Therefore, the apparent feedback voltage reads proportionately higher at low frequencies than that represented by the recording and the feedback voltage cannot be used to calibrate low-frequency response without calibration data. This condition in no way contributes to instability and the system is inherently stable.

Single-frequency power measurements give little useful information from feedback recorders. In order to arrive at a practical evaluation of the power required by program material with an RIAA pre-equalizer, records were made at high levels from orchestral and vocal numbers. These were made from a single-channel source connected to the recorder and phased to produce lateral-type recording. The material was of the type containing high-level sounds which produce extreme velocities with RIAA pre-equalization. The highest peak velocity recorded was 19 cm per sec. as measured with the light-band method. The highest RMS value of current through one drive coil was 0.6 ampere. Therefore, the maximum power consumed by each 5.6-ohm coil was approximately 2 watts. The maximum groove swing was $\pm$ 1.65 mils which indicates this was a high-level microgroove recording. It is doubtful if this level will be reached in stereophonic recording on disks with the normal pre-equalization characteristic.

The complex stylus driving mechanism of the 3A Recorder must necessarily contain additional masses and compliances not found in single-channel recorders. The recorded product covers the extremes from vertical through 45° recording to lateral recording. Differences in characteristics are to be expected under these conditions. Lateral recording is most subject to the effects of lateral stylus rotation at high frequencies. Rotation produces dips in level as well as inter-channel cross-talk. Vertical recording appears to contain a minimum of rotational effects. At this point in the development of the recorder it is difficult to attempt a true evaluation of the amount of cross-talk and the frequency response of the recorder above 8 kc. It may be stated the average amount of modulation on the record when viewed optically appears higher at high frequencies up to 20 kc than at mid-range frequencies. The curve in Figure 10 is indicative of the response of a single 45° channel with a calibrated reproducer. The recording was made with a production recorder on a lacquer disk at 33-1/3 rpm and at an approximate diameter of 10 inches.

Reproducer Design

In order to facilitate the description of the reproducer, it is desirable to review in greater detail the design objectives which should be met. The mass of the moving assembly, including that of the coils, was made as small as possible, consistent with appropriate strength and adequate output level. Stylus compliance was next chosen of such a value as to give approximately equal reactances of the stylus against the record groove wall at 400 and 8000 cps, the frequencies which represent the approximate range of program material of equal velocity on the record. An additional requirement of the compliance was that the static deflection of the stylus due to the tracking force was to be large compared to the maximum deflection due to record groove vertical modulation. This was necessary to insure that the stylus would not lose contact with the record. High torsional stiffness at the stylus was an obvious necessity to restrict the stylus movement to one of translation and not of rotation. The stylus radius of 0.70 mil was chosen as a good compromise between a small radius for low distortion and good high-frequency response versus a large radius for long record life. Stylus force was chosen as the minimum value which would provide tracking at the highest recorded levels normally encountered.

The electrodynamic or rotating-coil movement was selected as affording the best means of meeting these requirements. Figure 11 is an illustration showing a simplified front and side view of the principal operating components of the reproducer. The two coils are self-supporting and are mounted on Mylar hinges with the coil axes at right angles to each other and at 45° with the horizontal. The lower edge of each coil is connected by means of a link to a beam in which the stylus is mounted. The beam (which actually is not subjected to bending or twisting) consists of a hollow tube whose outside diameter is .031" and whose length is approximately 0.15". The rear of the beam is anchored in a flat spring which prevents rotation and at the same time provides essentially equal compliance at the stylus in any direction in the vertical plane. Some mechanical damping is applied to each link close to the coil by introducing a block of viscous semi-solid material between each link and the reproducer housing. This reduces the height of a peak in the frequency response characteristic at about 11 kc and also reduces cross-talk between channels. The drag wire prevents longitudinal motion of the stylus. It will be observed that the beam is tilted downward at an angle of approximately 3° with respect to the horizontal when the stylus is free. However, when the stylus rests on a record the beam is in a horizontal plane and parallel with the record surface. The vertical angle of the reproducer stylus motion has been made to conform closely with the angle of the stylus motion in the recorder to avoid the introduction of harmonic distortion in the vertical component of the reproduced signal. The vertical angle of stylus motion was mentioned briefly in discussing the recorder and was stated to be nominally 23°. The reproducer has been designed to provide a corresponding angle of deviation from the vertical plane.

The angle difference between recorder and reproducer introduces tracking angle errors and is subject to the same mathematical analysis as tracking angle errors in lateral disk systems. Considerable literature has been published on this subject[6].

The applicable equation for a velocity-sensitive reproducer to a close approximation is:

$$D = \frac{0.2 \times A \times M}{R}$$

where

D = % 2nd harmonic distortion at 33-1/3 rpm

A = Tracking-angle error in degrees

M = Modulation velocity (cm/sec. peak)

R = Distance of groove from record center (inches)

Figure 12 is a family of curves obtained by the use of this equation which shows the amount of second-harmonic distortion for various amounts of tracking angle error. The curves are for grooves cut at constant velocity and at different distances from the record center. In Figure 13 the second-harmonic distortion is shown for various groove radii for a constant tracking angle error and at two recorded velocities. In both cases the amount of distortion is reduced to 63 per cent of the values shown, by the average 4-db-per-octave slope in the RIAA reproducing characteristic.

Referring to Figure 11, the magnetic path consists of a center pole piece, two magnet pole pieces and a magnet. The latter is not shown. One edge of each coil is disposed in one of the gaps between the center pole piece and the magnet pole pieces. The polarity of the coil leads is so arranged that a laterally modulated groove will produce signals which are in phase in the output circuits.

Figure 14 is an outside view of the reproducer. It is entirely enclosed except for the small clearance hole for the diamond stylus and it mounts in a standard reproducer arm. It tracks properly with six grams of stylus force. The stylus compliance is nominally 2.6×10^{-6} cm/dyne and the dynamic mass is 3.0 milligrams. To avoid the difficulties involved in winding the moving coils with a large number of turns of extremely fine wire, a wire size has been chosen that results in a low output impedance of 2.4 ohms and a correspondingly low output voltage of 2 millivolts rms per channel for 10-cm-per-second peak velocity. A pair of transformers is therefore required for good signal-to-noise ratio.

A typical frequency response of one channel of the reproducer is shown in Figure 15. This characteristic was obtained by reproduction from an optically calibrated pressing containing single-channel test frequencies recorded at 45°. The slight rise at the low-frequency end of the curve is the result of the resonance of the arm mass with the stylus compliance and it will vary somewhat with the mass of the arm employed.

Tracing Distortion

Recordings made with the 45-45 method may vary all the way from single-channel 45° records to all vertical or all lateral types. Therefore, a single 45° channel may receive the same information from any of these types. However, tracing distortion may be expected to exhibit different characteristics in all three cases due to the differences in tracing distortion between vertical and lateral or combination recordings thereof.

The following distortion measurements were made from records recorded under all three conditions. The input level per channel was the same in all cases. The recorded velocity was 2.5 cm per sec. (peak) per channel. This resulted in an amplitude variation of ± 0.55 thousandths of an inch at the 400 cycle fundamental frequency.

Figure 16 shows the total harmonic distortion as measured through the usual high pass filter measuring device.

Figure 17 shows the intermodulation distortion measured with the RA-1258 Intermodulation Generator and the RA-1257 Intermodulation Analyzer.

The system of stereophonic recording on disk described in this paper appears to give highly satisfactory two-channel stereophonic reproduction. The cross-talk of the over-all system shown in the lower curve in Figure 15 appears to be entirely satisfactory for stereophonic listening and the frequency response is entirely adequate for high-fidelity reproduction in the home. Experience to date indicates little if any added problems in producing pressings with a low noise level. The system has been demonstrated widely to representatives of the disk recording studios and reproducing equipment industries both here and in Europe. A limited number of 3A Recorders has been placed in the hands of several recording companies and experimental disk recordings have been made available by some of them. Interest shown by industry and the public in the demonstration of the 45-45 StereoDisk System would appear to confirm the soundness of this approach to the application of stereophonic recording principles to the field of disk recording.

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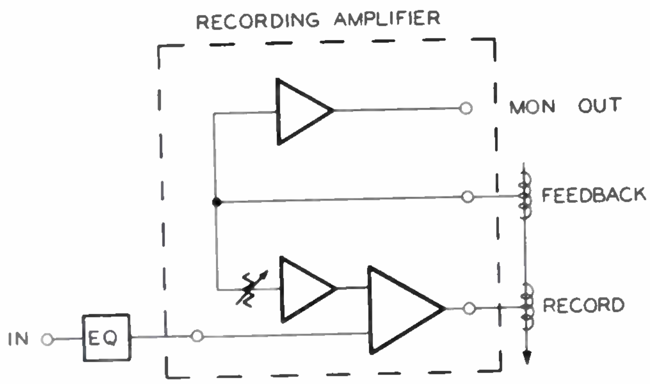

Fig. 1 Block schematic of recording amplifier.

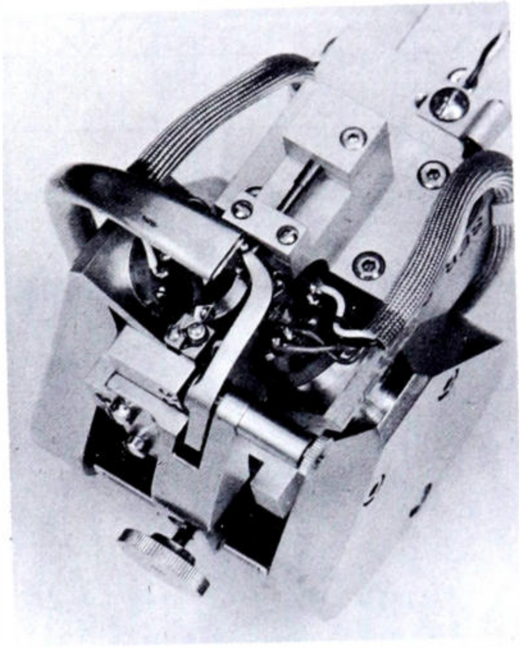

Fig. 3 Bottom view of 3A StereoDisk recorder.

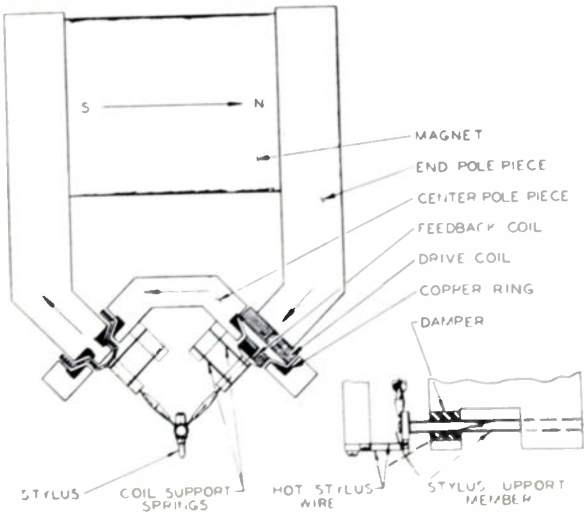

Fig. 2 Simplified recorder illustration.

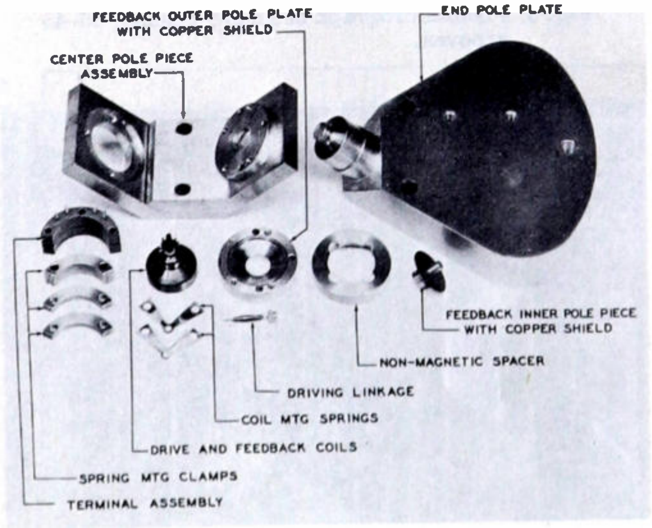

Fig. 4 View of significant parts of recorder.

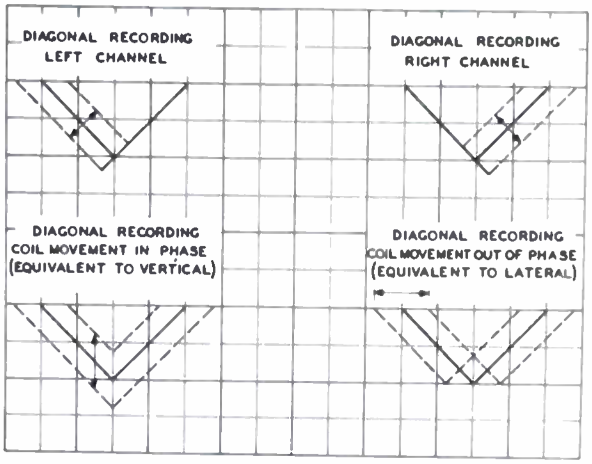

Fig. 5 Cross section of 45-45 grooves for four limiting conditions.

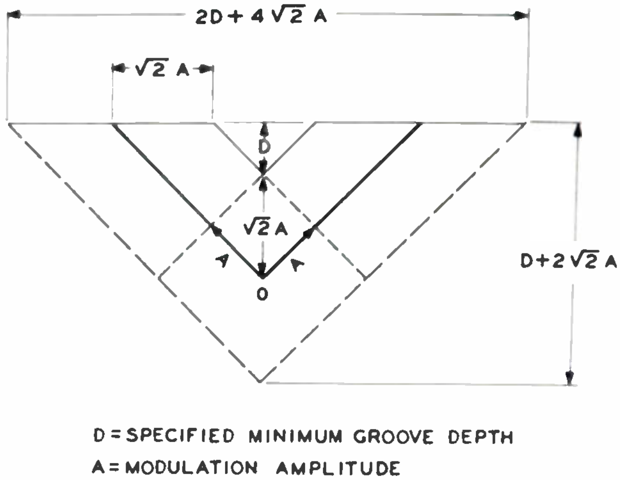

Fig. 8 Cross section of 45-45 groove for maximum groove excursion.

Fig. 6 Photomicrograph of single-channel 45-45 grooves.

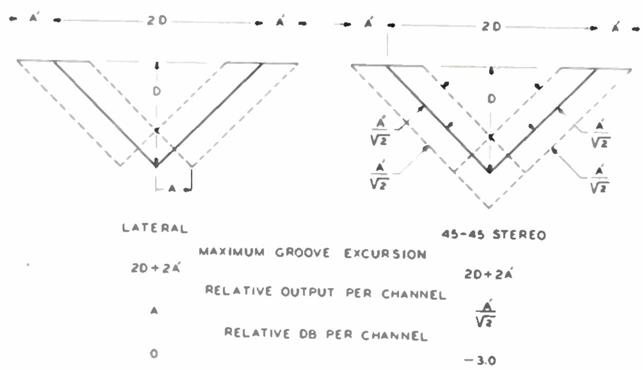

Fig. 9 Comparison of 45-45 with standard lateral groove.

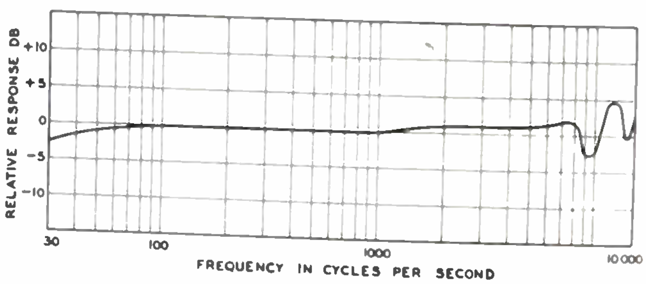

Fig. 10 Typical recorded frequency characteristic.

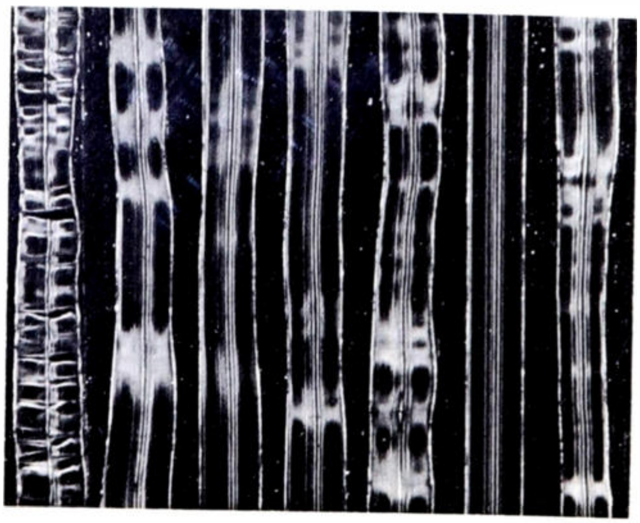

Fig. 7 Photomicrograph of 45-45 grooves with stereophonic program material.

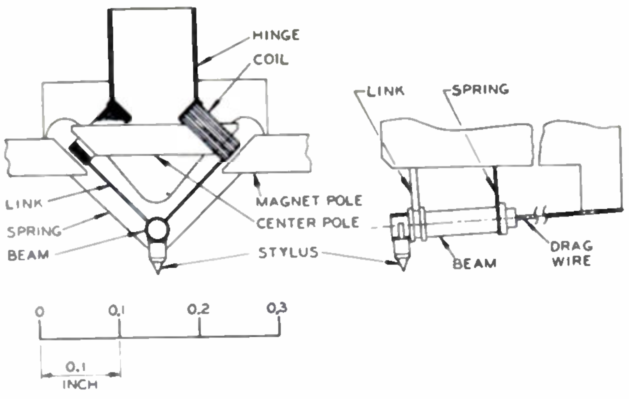

Fig. 11 Simplified reproducer illustration.

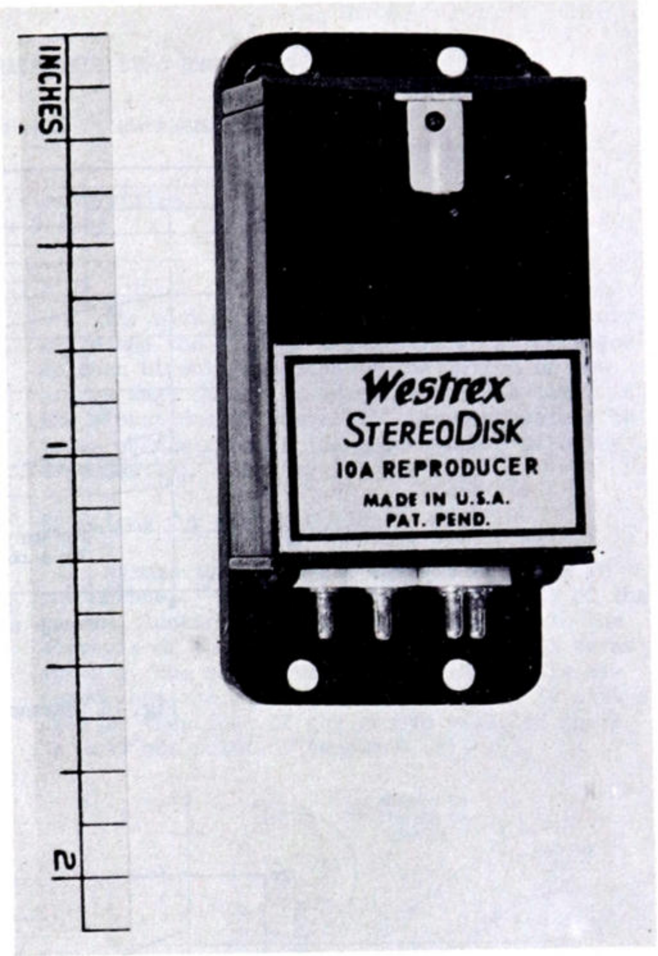

Fig. 14 View of 10A StereoDisk reproducer.

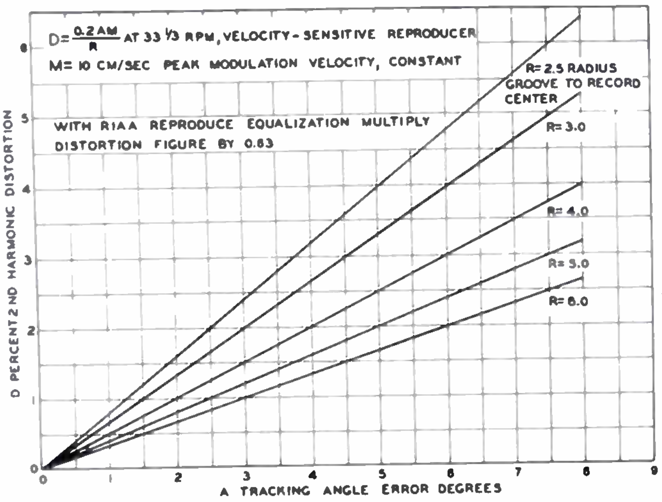

Fig. 12 Distortion vs. tracking-angle error.

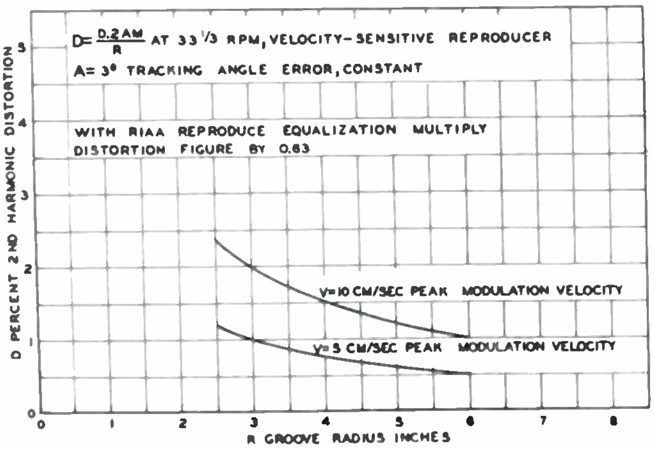

Fig. 13 Tracking-angle distortion vs. groove radius and recorded velocity.

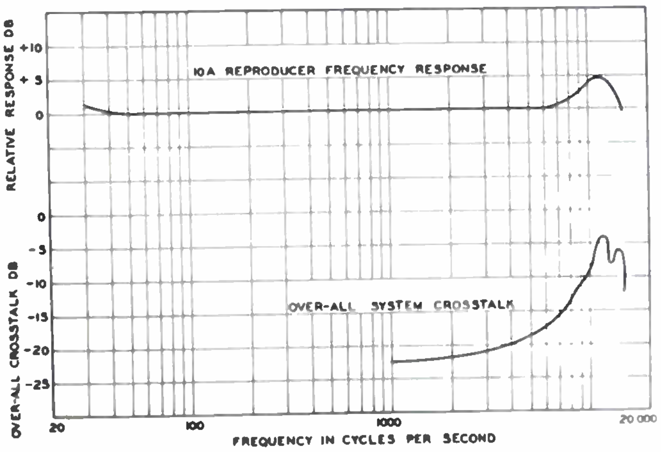

Fig. 15 Reproducer frequency characteristic and system cross-talk.

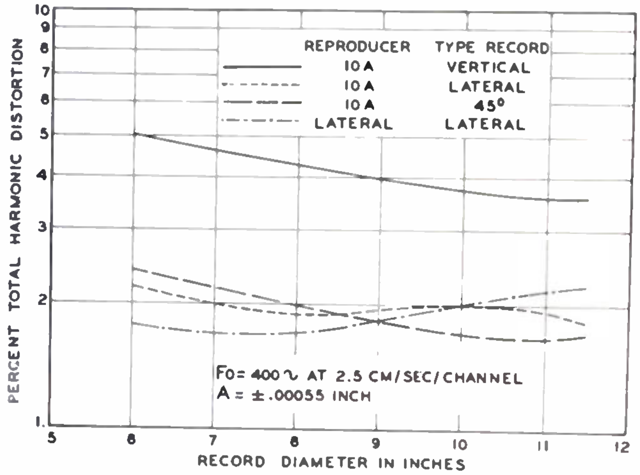

Fig. 16 Harmonic distortion curves.

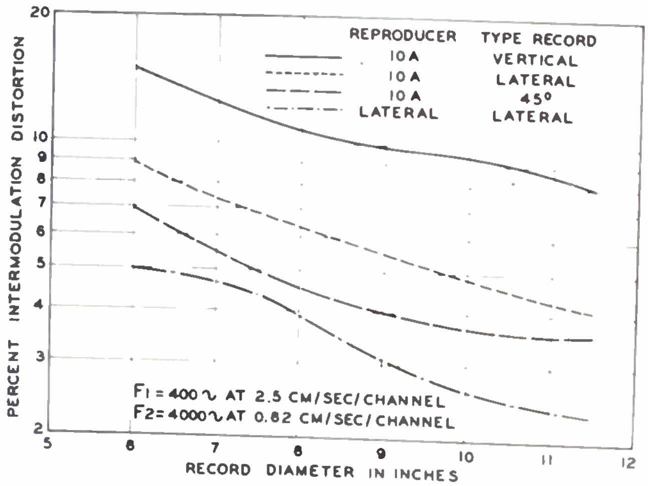

Fig. 17 Intermodulation distortion curves.

TRACING DISTORTION IN STEREOPHONIC DISC RECORDING

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Introduction

Although the theory of tracing distortion in vertically and laterally cut records has been known for many years, it will be reviewed briefly to show the method of computation and the modifications required for stereophonic recordings. In previous work, the tracing distortion was obtained by a harmonic analysis of the curve traced by the center of a spherical stylus. The amplitudes of the harmonic frequencies were found by (1) finding the coordinates of the curve and making a numerical harmonic analysis[1] or (2) by expanding an explicit expression for the curve in a power series.[2,3] In either event, the amount of work necessary was prodigious when more than one frequency was involved. When the record groove is modulated laterally and vertically simultaneously, the amount of labor required to compute the various distortion components, using a desk calculator, would be too great to attempt. In view of this, it was decided to make the computations by programming an electronic digital computer to use method (1).

No Crosstalk Between Channels

In the 45°-45° stereophonic record system with a groove angle of 90° as shown by Fig. 1,

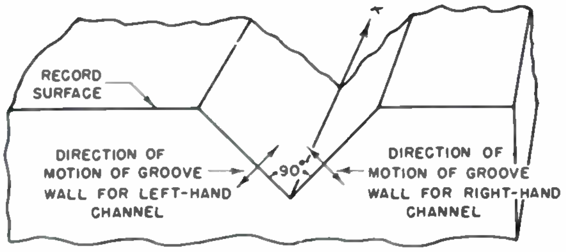

FIG. 1 GROOVE WALL MOTIONS FOR 45°-45° SYSTEM

the outer wall of the groove is modulated by moving it parallel to itself with amplitude proportional to the signal in the right-hand channel. The inner groove wall is moved in a similar manner in accord with the signal in the left-hand channel. If identical signals are fed into both channels, the phasing is chosen so the groove motion is lateral only.

The pickup is designed with two output circuits and the axes of the elements are arranged so each circuit responds to the motion of one groove wall only. If adequate care is taken in the pickup design there will be no crosstalk between channels since the groove walls move independently.

Equations for Stylus Motion

Assume that a cosine wave is recorded in one channel. A cross section of the wall of the groove, looking in a direction parallel to the elements of the cylindrical surface, will be as shown by the solid curve of Fig. 2. It is assumed that the curvature of the stylus is always greater than that of the groove wall, so there is only one point of tangency (x_1, y_1).

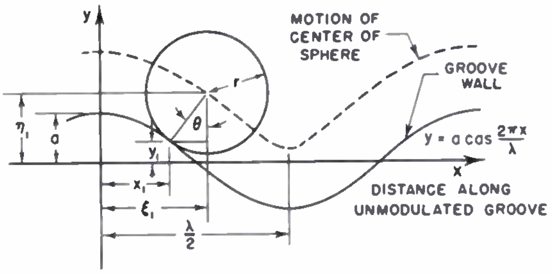

FIG. 2 SPHERICAL STYLUS TRACING A COSINE WAVE

The equation of the center of the sphere (ξ_1, η_1) using the coordinate system shown is

$$\xi_1 = x_1 + r \sin\theta \tag{1}$$

$$\eta_1 = y_1 + r \cos\theta \tag{2}$$

where $y_1 = a \cos k x_1$ and $k = 2\pi/\lambda$. The angle θ is defined by

$$\tan\theta = - \left.\frac{dy}{dx}\right|_{x=x_1} = - y_1' \tag{3}$$

so that $\sin\theta = - y_1'/\left[1 + (y_1')^2\right]^{\frac{1}{2}}$ and $\cos\theta = 1/\left[1 + (y_1')^2\right]^{\frac{1}{2}}$.

When the curve traced by a spherical stylus of radius r is an arbitrary function $y = f(x)$, the coordinates of the center of the sphere can be written

$$\xi_1 = x_1 - \frac{r\, y_1'}{\sqrt{1 + (y_1')^2}} \quad (4)$$

$$\eta_1 = y_1 + \frac{r}{\sqrt{1 + (y_1')^2}} \quad (5)$$

where $y_1' = - \left.\frac{df}{dx}\right|_{x=x_1}$.

The coordinates of the center of the sphere for the cosine wave $y = a \cos kx$ become

$$\xi_1 = x_1 + \frac{r\, a\, k \sin k\, x_1}{\sqrt{1 + a^2 k^2 \sin^2 k\, x_1}} \quad (6)$$

$$\eta_1 = a \cos k\, x_1 + \frac{r}{\sqrt{1 + a^2 k^2 \sin^2 k\, x_1}} \quad (7)$$

If equations (6) and (7) are each multiplied by k and the substitutions

$$A = ka \qquad H = \eta_1 k$$
$$R = kr \qquad \Xi = \xi_1 k$$
$$X = kx_1 \quad (8)$$

are made, the normalized form of equations (6) and (7) are

$$\Xi = X + \frac{R A \sin X}{(1 + A^2 \sin^2 X)^{\frac{1}{2}}} \quad (9)$$

$$H = A \cos X + \frac{R}{(1 + A^2 \sin^2 X)^{\frac{1}{2}}} \quad (10)$$

Solution of Equations for Stylus Motion

To obtain a table of corresponding values of Ξ and H, equations (9) and (10) must be solved simultaneously by eliminating X, the normalized distance down the unmodulated groove. H should be tabulated at equidistant steps in Ξ to simplify the numerical harmonic analysis.

Equation (9)

$$X + \frac{R A \sin X}{(1 + A^2 \sin^2 X)^{\frac{1}{2}}} - \Xi = 0 \quad (11)$$

is solved for X for each value of Ξ by Newton's method[4] of successive approximations. If X_1 is an approximate value of the desired root, Newton's formula for a more accurate value, X_2, for the root of the equation $f(X) = 0$ is

$$X_2 = X_1 - \frac{f(X_1)}{f'(X_1)} \quad (12)$$

where X_1 is the first approximation and $f'(X_1)$ is the value of the first derivative at X_1. This formula can be used successively until the desired degree of accuracy is obtained. When X has been found, equation (10) is used to find the value of H corresponding to the assumed value of Ξ. The result is a table of values of H for uniformly spaced values of Ξ as shown by Table I. The increment in Ξ is determined by how rapidly the modulation varies and by the accuracy required.

TABLE I

Ξ	H
0°	—
5°	—
10°	—
15°	—
—	—
—	—

Harmonic Analysis of Stylus Motion

From the table of values of Ξ and H just obtained, the Fourier series representation

$$H = \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos n\Xi \quad (13)$$

where

$$A_n = \frac{2}{\pi}\int_0^{\pi} H \cos n\Xi \; d\Xi \quad (14)$$
$$n=0, 1, 2, \ldots$$

can be obtained by the Newton-Cotes formulas[5] for numerical integration. The evaluation of integral (14) for the harmonic amplitudes was programmed so that the entire calculation was done by the computer. Although equations (9) and (10) were derived on the basis of a single tone, it is obvious that the analysis can be extended to multitone operation by using equations (4) and (5).

Analysis of Vertical-Lateral System

Cross Modulation Obtained

In the vertical-lateral system of stereorecording, the cutting stylus is modulated vertically and laterally simultaneously. The lateral channel produces a second-harmonic term in the vertical channel because of the pinch-effect. The tones in the vertical and lateral channels beat together to produce sum and difference frequencies in the lateral channel.

Equations for Stylus Motion

The vertical-lateral system can be analyzed in the following manner: Let the vertical and lateral modulation of the cutting stylus be given by $F_1(x)$ and $F_2(x)$ respectively, where x is the distance down the unmodulated groove. The vertical and lateral motions may be resolved into components $f_1(x)$ and $f_2(x)$ respectively perpendicular to the groove walls OB and AO as

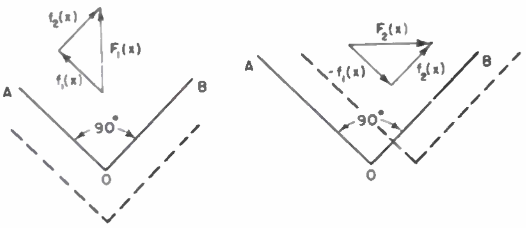

FIG. 3 VERTICAL AND LATERAL MOTION OF CUTTING STYLUS

shown in Fig. 3. In terms of $f_1(x)$ and $f_2(x)$, the vertical and lateral displacements become

$$F_1(x) = \frac{1}{\sqrt{2}}\left[f_1(x) + f_2(x)\right] \tag{15}$$

and

$$F_2(x) = \frac{1}{\sqrt{2}}\left[- f_1(x) + f_2(x)\right]. \tag{16}$$

Solving equations (15) and (16) for $f_1(x)$ and $f_2(x)$ gives

$$f_1(x) = \frac{1}{\sqrt{2}}\left[F_1(x) - F_2(x)\right] \tag{17}$$

$$f_2(x) = \frac{1}{\sqrt{2}}\left[F_1(x) + F_2(x)\right] \tag{18}$$

By use of equations (4) and (5), the distance from the center of the sphere to the unmodulated sidewalls OA and OB may be found as a function of the distance, x, along the unmodulated groove.

Single Tone in Each Channel: Let the cutting stylus be modulated by the tones

$$F_1(x) = a_1 \cos k_1 x \tag{19}$$

and

$$F_2(x) = a_2 \cos k_2 x \tag{20}$$

in the vertical and lateral directions respectively, where a_1 and a_2 are the amplitudes of motion and k_1 and k_2 are the angular frequencies of the tones. Using equations (17) and (18), the equations for the sidewall motion are

$$y_1 = f_1(x) = \frac{1}{\sqrt{2}}\left[a_1 \cos k_1 x - a_2 \cos k_2 x\right] \tag{21}$$

and

$$y_2 = f_2(x) = \frac{1}{\sqrt{2}}\left[a_1 \cos k_1 x + a_2 \cos k_2 x\right] \tag{22}$$

The coordinates for the center of the sphere are obtained by substituting equations (21) and (22) into equations (4) and (5). After normalization, the equations of the stylus motion are given by

$$\Xi_1 = X_1 - \frac{RW_1}{\sqrt{1 + W_1^2}} \tag{23}$$

$$H_1 = A \cos X_1 - \beta A \cos \alpha X_1 + \frac{R}{\sqrt{1 + W_1^2}} \tag{24}$$

$$\Xi_2 = X_2 + \frac{RW_2}{\sqrt{1 + W_2^2}} \tag{25}$$

$$H_2 = A \cos X_2 + \beta A \cos \alpha X_2 + \frac{R}{\sqrt{1 + W_2^2}} \tag{26}$$

where

$$W_1 = - A \sin X_1 + \alpha\beta A \sin \alpha X_1,$$

$$W_2 = A \sin X_2 + \alpha\beta A \sin \alpha X_2,$$

$\alpha = k_2/k_1$, $A = a_1 k_1/\sqrt{2}$, $X_1 = k_1 x_1$, $\Xi_1 = k_1 \xi_1$, $H_1 = k_1 \eta_1$, $\beta = a_2/a_1$, $R = k_1 r$, $X_2 = k_1 x_2$, $\Xi_2 = k_1 \xi_2$, $H_2 = k_1 \eta_2$. x_1 is the distance to the point of contact on groove wall OA.

Since the normalized coordinate of the center of the sphere measured along the unmodulated groove is the same for both sidewalls, $\Xi_1 = \Xi_2 = \Xi$. H_1 and H_2 are therefore each expressible as a function of Ξ. $H_1(\Xi)$ is found by solving equation (23) for X_1 for each successive assumed value of Ξ by Newton's method and substituting this value of X_1 into equation (24). $H_2(\Xi)$ is found in the same manner using equations (25) and (26).

Stylus Displacement: Fig. 4 shows the position

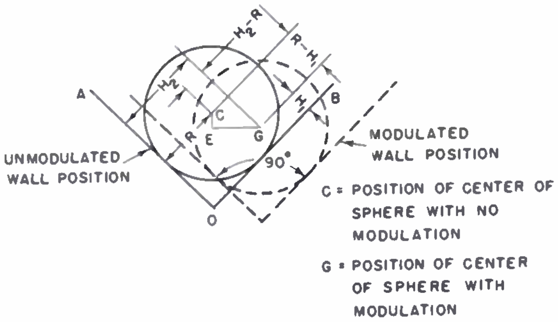

FIG. 4 COORDINATES OF MOVEMENT OF SPHERE

of the groove in the modulated and unmodulated states. In this figure, H_1 represents the distance of the center of the sphere from the unmodulated right-hand groove wall OB at any displaced position. The distance of the center of the sphere from the other unmodulated sidewall AO is given by H_2. Let the quantities D_1 and D_2 denote the normalized vertical and horizontal displacements of the center of the sphere. In Fig. 4 these distances are given by D_1 = CE and D_2 = EG respectively. When expressed in terms of H_1 and H_2 with the aid of Fig. 5, the vertical and horizontal displacements are

$$D_1 = \frac{H_1 + H_2 - 2R}{\sqrt{2}} \qquad (27)$$

$$D_2 = \frac{H_2 - H_1}{\sqrt{2}} \qquad (28)$$

A harmonic analysis of equations (27) and (28) by the method previously described can be made to find the harmonic and crosstalk distortion in the two channels. To solve for the

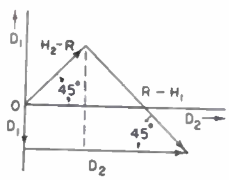

FIG. 5 RELATION BETWEEN H_1, H_2, D_1 AND D_2

harmonic coefficients, the integration indicated by equation (14) must extend over π radians of the difference frequency $k_1 - k_2$, which may be several cycles of each signal frequency.

The analysis described above can be extended to cases where there is more than one tone in each channel. The computation time will increase rapidly, since the integration interval must be one half cycle of the lowest possible beat frequency of the tones involved.

Curves of Calculated Distortion

To obtain curves of harmonic distortion in terms of the tangential groove velocity v and the recording velocity u, the normalized amplitude A and radius R in equations (9) and (10) must be expressed in terms of these quantities. Since the tangential groove velocity for a frequency f is given by $v = \lambda f$ and the amplitude a for a recording velocity u is $a = u/2\pi f$, the normalized amplitude is

$$A = ka = \frac{2\pi a}{\lambda} = \frac{u}{v}. \qquad (29)$$

The normalized radius is then

$$R = kr = \frac{2\pi r}{\lambda} = \frac{2\pi f r}{v} \qquad (30)$$

Harmonic Distortion in Vertical and Lateral Systems

Using the proper values of A and R in equations (9) and (10), the coordinates of the stylus motion and its harmonic amplitudes have been found for a recording frequency of 400 cycles per second. The percent second harmonic amplitude is shown in Figs. 6(a), 6(b), and 7(a) for recording velocities of 7, 14, and 22 cm/sec for three different stylus radii. These curves correspond to the tracing distortion obtained in each channel of the 45°-45° system using an amplitude sensitive pickup. For a velocity sensitive pickup, the second harmonic amplitude would be multiplied by 2, the third by 3, etc. The same curves also apply to the vertical channel of the vertical-lateral system when the crosstalk from the lateral channel is less than 0.1 percent. Higher order harmonics which are not shown in these figures have amplitudes which are in the order of or less than 0.1 percent. With a recording velocity u = 22 cm/sec the magnitude of the third component is more than 20 db down from the amplitude of the second harmonic component on an amplitude basis.

The harmonic distortion in a laterally-cut record, recorded at a velocity of 22 cm/sec, is shown in Fig. 7(b) for comparison purposes. In this case the only significant term is the third harmonic component which is more than 20-to-1 down in amplitude from the second harmonic component in the corresponding vertical recording.

Intermodulation Distortion in Vertical and Lateral Channels: The intermodulation distortion with two tones of frequencies 400 and 4,000 cycles in one channel has been plotted in Figs. 8(a) and 8(b) for the vertical and lateral channels.

For a vertically-cut record the intermodulation, I, has been defined as

$$I = \frac{\text{Amplitude of 3600 cycle tone} + \text{amplitude of 4400 cycle tone}}{\text{amplitude of 4000 cycle tone}} \qquad (31)$$

For a laterally-cut record the intermodulation is

$$I = \frac{\text{Amplitude of 3200 cycle tone} + \text{amplitude of 4800 cycle tone}}{\text{amplitude of 4000 cycle tone}} \qquad (32)$$

A recording velocity ratio of u_1/u_2 = 4 has been assumed for the 400 and 4000 cycle tones so that the amplitude of the 4000 cycle tone for a constant velocity recording is 1/40 of that for the 400 cycle tone.

It is noted that the intermodulation in the vertical channel is more than 10 times that in the lateral channel for the same recording conditions. The intermodulation limitations are the same in the vertical-lateral system as in the 45°-45° system since each uses a vertical channel. To decrease the intermodulation distortion it is necessary to reduce the recording level or the stylus radius.

A close approximation to the intermodulation is given by the formulas

$$I = \frac{800\,\pi\, u_1 r}{v^2}, \text{ (vertical recording)} \quad (33)$$

$$I = \frac{800\,\pi^2 u_1^2 r^2}{v^4}, \text{ (lateral recording)} \quad (34)$$

where

I = percent intermodulation,

u_1 = recording velocity in inches/sec (400 ~),

r = stylus radius in mils,

v = groove velocity in inches/sec.

Equations (33) and (34) are accurate to within 5 percent at 75 percent intermodulation and within 1 percent at 25 percent intermodulation. These formulas are based on the relation between the harmonic amplitudes and the intermodulation distortion, and on the formulas derived by Corrington.[3]

Combination Tone Amplitudes for Vertical-Lateral System: Combination tones include all possible frequencies, harmonic, sum, and difference, etc., that are produced in a given channel. The curves shown in Figs. 9(a) and (b) were calculated with equal amplitude sine waves of frequencies 400 cycles and 300 cycles recorded in the vertical and lateral channels respectively. Combination tone amplitudes in the vertical channel include the 600-cycle pinch-effect crosstalk from the lateral channel and the 800-cycle second-harmonic distortion generated within the channel. Two recording amplitudes were chosen; these are u = 7 cm/sec and 14 cm/sec.

In the lateral channel the significant distortion frequencies are the sum and difference frequencies of 100 and 700 cycles per second. All other distortion terms are less than 0.1 percent. These distortion components are seen to be larger than the harmonic distortion or the pinch-effect component in the vertical channel.

The curves of Figs. 10 and 11 show the variation of the combination-tone amplitudes when the radius of the stylus is made 0.50 mils and 0.25 mils respectively.

Relation Between Groove Velocity and Record Diameter

The groove velocity was given as the abscissa of the preceding curves so that they would be independent of the particular choice of revolutions per minute. Fig. 12 can be used to find the record diameter corresponding to a given groove velocity.

Conclusions

Each wall of the groove of the 45°-45° system is equivalent to a channel with vertical recording. Since this is a single-sided system there will be second and third harmonic distortion. If the pickup is properly designed, there will be no cross modulation between channels. Curves are given for the harmonic and intermodulation distortion in either channel for various levels and stylus radii. The second harmonic component is the dominant distortion term and is approximately directly proportional to the stylus radius, the recording velocity and inversely proportional to the square of the groove velocity.

In either the 45°-45° or the vertical channel of the vertical-lateral system, the percentage distortion is the same, therefore the distortion limitations are the same for either system when the channels are considered individually.

To keep the distortion as low as in lateral recordings, the level should be lower than that presently used for single-channel lateral recordings, and the stylus radius should be less than the 1 mil commonly used. There is a limit to the allowable reduction in the recording level since the signal-to-noise ratio is also reduced as the level is decreased.

When the vertical-lateral system is used, the distortion will be different in the two channels. The second harmonic of the tone in the lateral channel will appear in the vertical channel because of pinch-effect. The vertical channel will also contain second and third harmonics of the modulation in the vertical channel.

The tones in the vertical and lateral channels will beat together and produce sum and difference tones in the lateral channel. These sum and difference amplitudes in the lateral channel are larger than the pinch-effect term produced in the vertical channel. There will also be third harmonic distortion in the lateral channel, however, this harmonic will be quite small in amplitude compared to the second harmonic in the vertical channel.

No attempt has been made to set maximum allowable limits for the distortion in a stereophonic system. This should be done after extensive listening tests where the levels and stylus radii are varied systematically. On the basis of a limited number of listening tests made to date, it may be that the present limits for single-channel systems are unnecessarily low for a pleasing stereophonic system.

Acknowledgment

The program for the computer was prepared by R. F. Kolar. Without his skill, persistence and efficient use of the computer, it would have been difficult to study so many cases.

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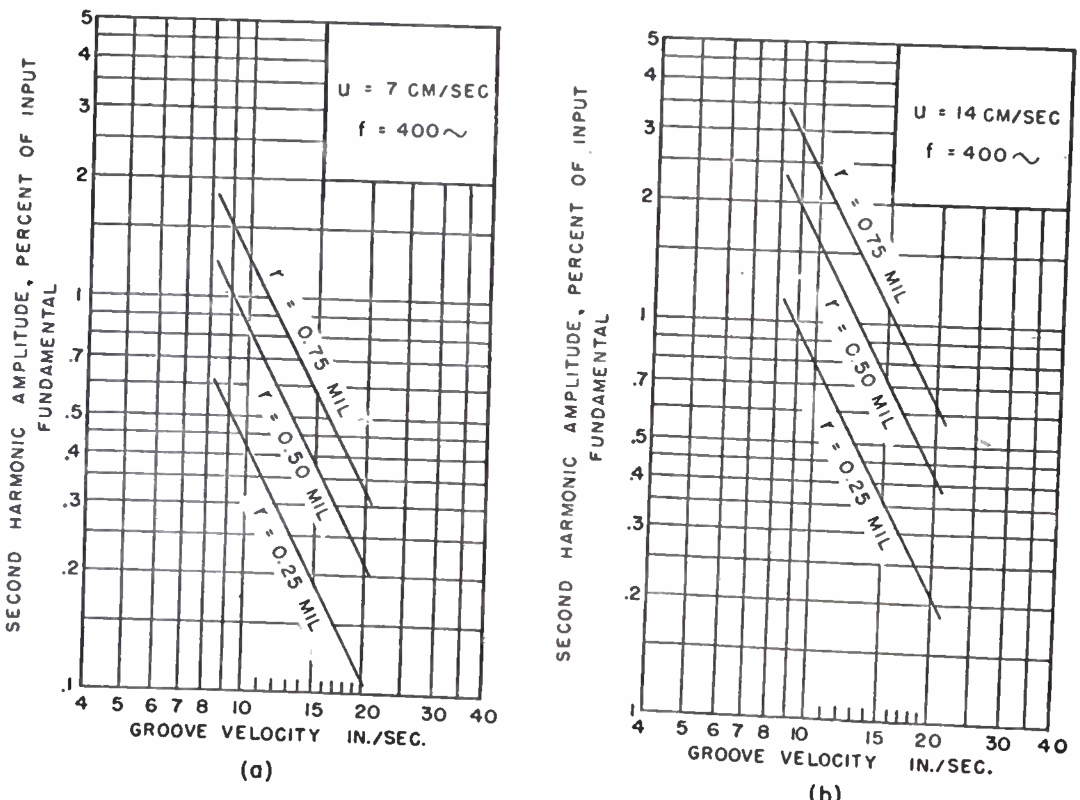

FIG. 6 SECOND HARMONIC AMPLITUDE FOR VERTICAL CUT RECORDING

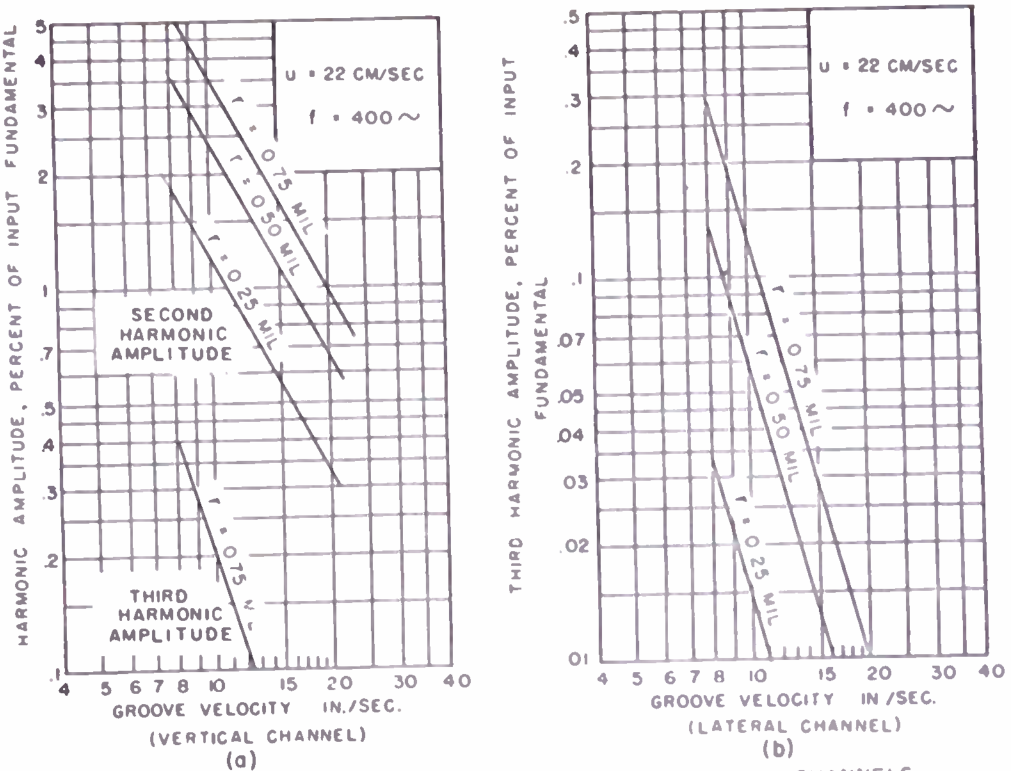

FIG. 7 HARMONIC DISTORTION IN VERTICAL AND LATERAL CHANNELS

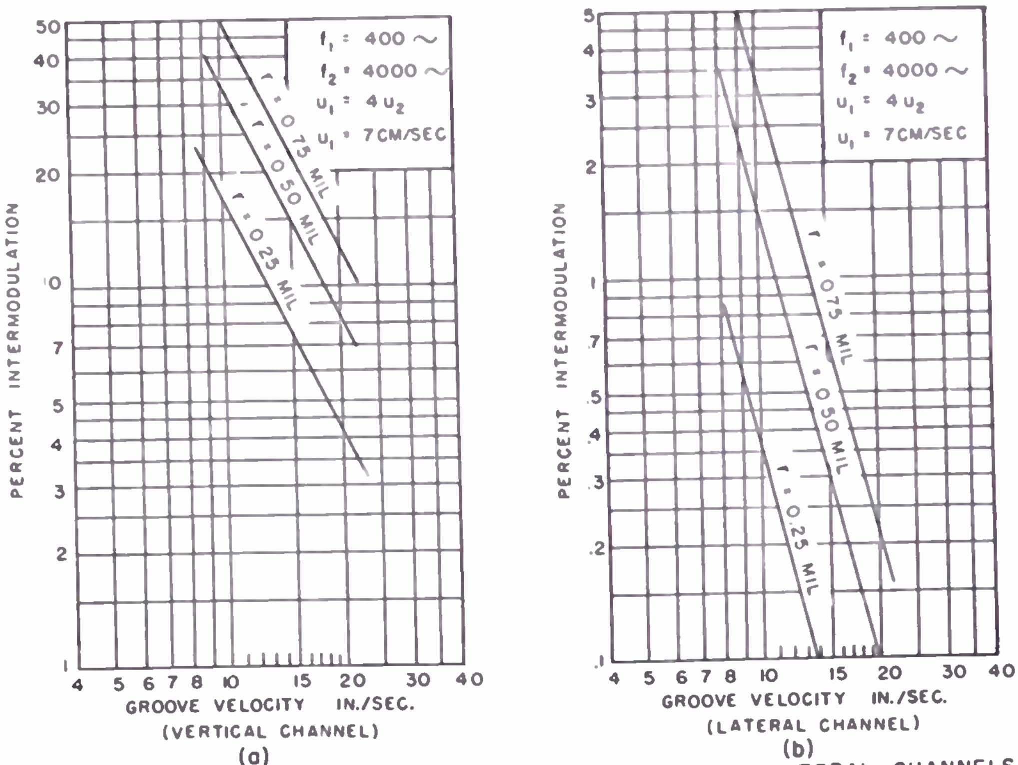

FIG. 8 INTERMODULATION DISTORTION IN VERTICAL AND LATERAL CHANNELS

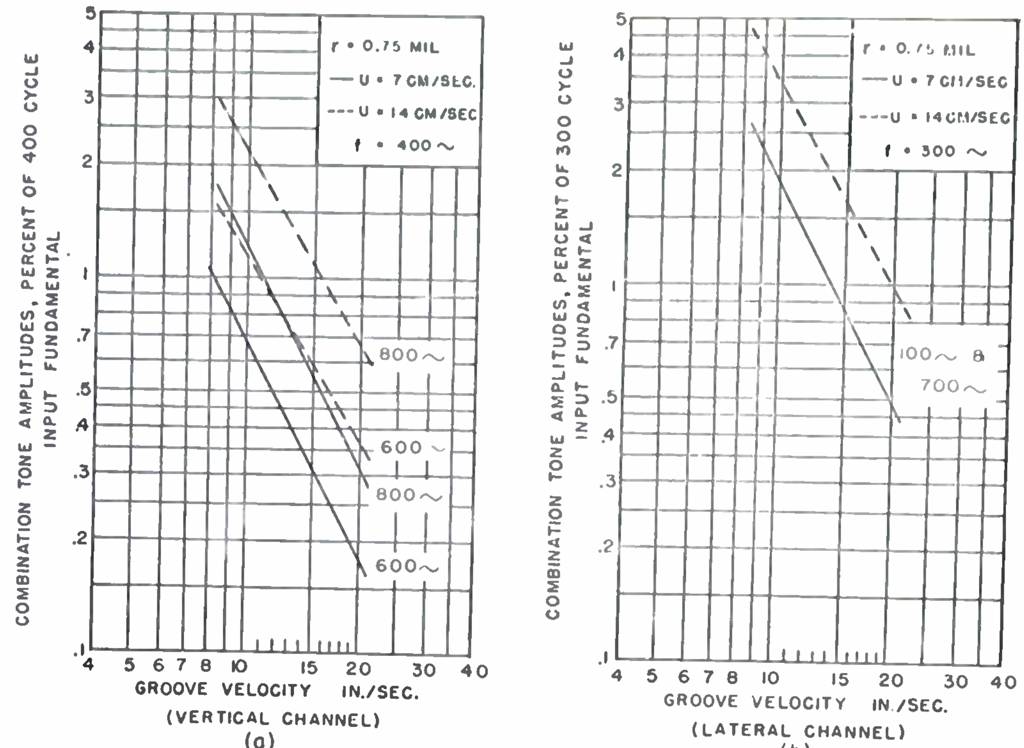

FIG. 9 COMBINATION TONE AMPLITUDES FOR VERTICAL-LATERAL SYSTEM

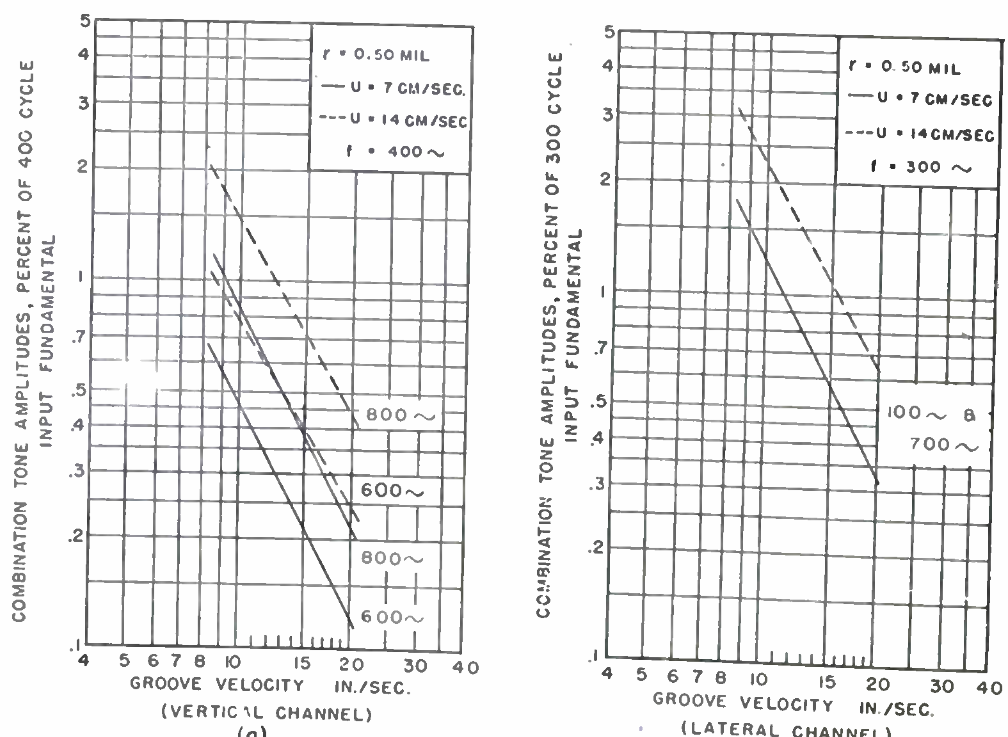

FIG. 10 COMBINATION TONE AMPLITUDES FOR VERTICAL-LATERAL SYSTEM

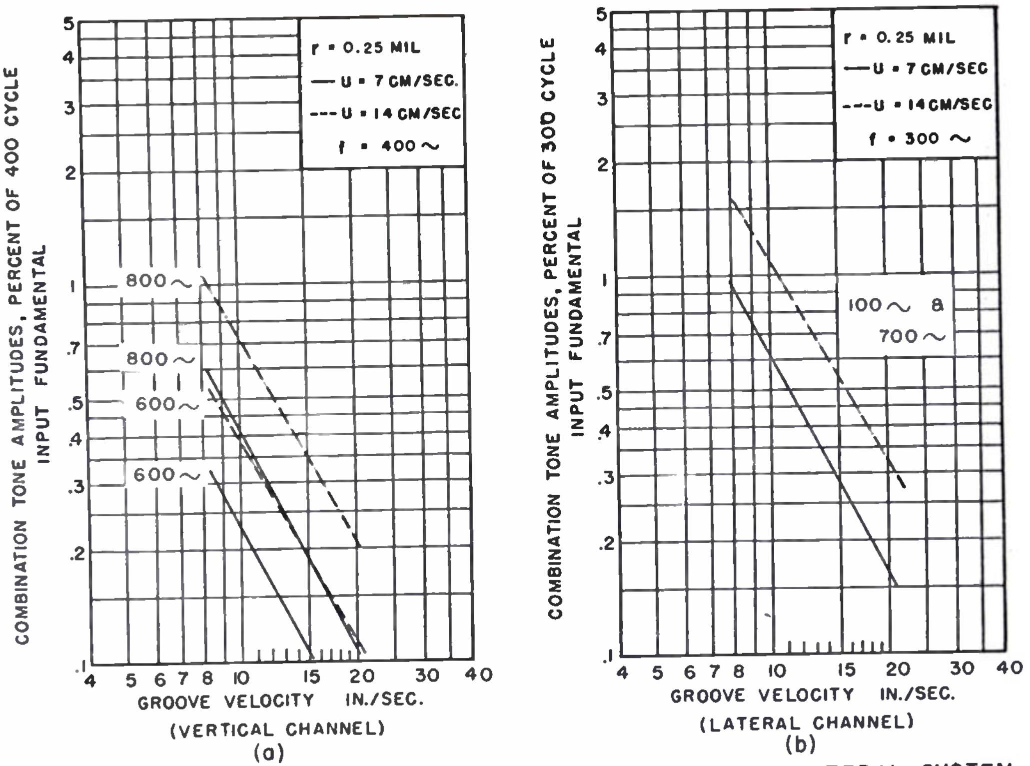

FIG. 11 COMBINATION TONE AMPLITUDES FOR VERTICAL-LATERAL SYSTEM

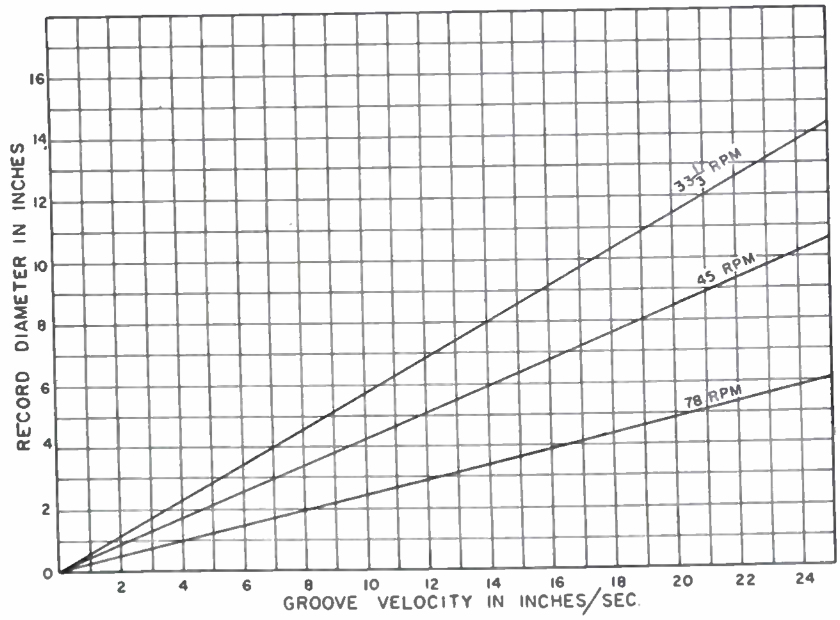

FIG. 12 RELATION BETWEEN RECORD DIAMETER AND GROOVE VELOCITY

COMPATIBILITY PROBLEMS IN STEREOPHONIC DISC REPRODUCTION

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ABSTRACT

The addition of stereophonic information to a record groove introduces a vertical component of groove modulation which generates tracking problems with present day phonograph pickups. A special set of test records was pressed to study the tracking ability of phonograph pickups for vertical modulation. A number of commercial pickups were tested. The results give a measure of the vertical impedance of commercial pickups and provide means for ascertaining the maximum modulation usable in a compatible record.

PHONOGRAPH PICKUPS FOR STEREOPHONIC RECORD PRODUCTION

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ABSTRACT

The introduction of stereophonic records will require the design of new pickups capable of translating record groove modulation in orthogonal modes. Several schemes are described, together with the advantages and disadvantages of each.

THE REQUIREMENTS OF A RECORD CHANGER, THE COMPONENT PART AND ASSOCIATED EQUIPMENT FOR STEREOPHONIC RECORD REPRODUCTION.

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Stereophonic music is probably one of the greatest advances in music listening pleasure that has been introduced since the beginning of the recording industry. Due to the distorted frequency response of the ear, and to the reduction of high frequency response with age, the full benefit of high fidelity cannot always be appreciated; but, any person with two ears can hear and appreciate the stereo recordings which are now becoming available on discs. To make these possible, it is necessary that all the components in the system be designed with this thought in mind.

There have been certain recommended standards suggested to the EIA and because this system is new, some of the requirements of the system will be discussed.

Stereo Records

The stereo records will be recorded at 33 1/3 and 45 RPM. The physical size of the record will be the same as the present LP and 45 - 7" discs. The information from the right hand microphone will be recorded on the outside wall of the groove, and this should correspond to the right hand speaker when the recording is being played back. The inner groove will have the information from the left hand microphone and for the right hand speaker. The recorded level of stereo disc will be down from 3 to 5 DB.

Needle

It has been recommended that a .7 mil needle ± .1 mil at 6 grams ± .1 gram pressure be used for record changers. This will permit the same needle to be used on LP and 45" records. A smaller needle may be used to reduce distortion for stereo records only. It was felt that the smaller needle would not be practical for the average person to handle, and would be quite difficult for the needle manufacturers to produce. All the record changers on the market today will function properly at 6 grams needle pressure. The smaller tipped needle would not be used with LP or 45" records, as it would be riding the bottom of a groove. The .7 mil needle at 6 grams would be comparable to the present day .1 mil needle playing LP records, as far as record and needle wear are concerned, providing the compliance of the cartridge is satisfactory.

Cartridges

The output level of the two elements should be equal to each other with 3 DB, also the frequency response of each element should be between a 3 DB envelope. Signal separation between elements should be a minimum of 15 DB between 500 to 6000 cycles for enjoyable orchestration reproductions. For special sound effects, it may be necessary to increase the separation and extend the frequency response. The vertical versus horizontal output of each element of the cartridge should be equal; otherwise amplitude distortion will result. Temperature characteristics of the crystal cartridge should be considered and the circuitry designed for a maximum output when the cartridge is at the temperature of the enclosure in which the changer is to operate; but this is liable to cause a mechanical feedback condition when the unit is first turned on. The termination of the leads on the back of the cartridge should be identified as to right and left channels. It is recommended that for stereo phonograph cartridges with 3 or 4 pin termination; (1) the common terminal be center, black color or identified with the letter "C" located on or near the terminal; (2) the right channel terminal to be copper or brass color, or identified with the letter "R" and (3) the left channel terminal be silver or nickel color or identified with the letter "L". Also, it is proposed that a standard size terminal be adopted for stereo phonograph cartridges which will accept a .047" to .051" diameter pin jack.

If the unit is to be played through, or with an AC-DC type amplifier, that is, one without transformer isolation between the main line and amplifier circuit, the 4 pin cartridge should be used for two reasons: (1) to reduce hum (2) to reduce shock hazards and comply with underwriters approval, or requirements.

Tone Arm

It has been found that the tone arm has a definite bearing on the overall rumble transmitted through the changer. It is best to select a cartridge and tone arm combination that will reproduce with a minimum of rumble. I have found as much as 6 DB rumble variation between combinations of tone arms and cartridges. This is due to the resonant frequencies of the arm and cartridge combination. Both horizontal and vertical friction should be reduced to a minimum, to insure better tracking. The tone arm should be so designed that it has a tracking error of not more than 3 degrees.

Changer

It is necessary to have better vertical compliance in the complete changer to compensate for the vertical output of the cartridge. This may be accomplished by more weight in the changer itself, or more compliant spring mounting. This will reduce the tendency towards mechanical feedback and also improve rumble conditions. On 2 pull motors, the motor should be as far as possible from the changer base. This will help reduce 120 cycle rumble. The rotor in the motor itself should be balanced to a very high degree for elimination of 60 cycle rumble. On a 4 pull motor, the rotor should also be balanced to a fine degree for elimination of 30 cycle rumble. Improved rubber mounting of the motor itself may be quite advantageous.

Both mechanical feedback and rumble are more pronounced in the stereo system, because the stereo system is recorded at a lower level; therefore, at equal volumes the rumble figure will be increased equal to the amount of the reduced recorded level. Also, the stereophonic system will have better air coupling at the lower frequency response because by doubling the amount of speakers, low frequency rumble and hum are more apparent.

Capacity and Impedance Match Between Cartridge and Amplifiers

There are several points of interest regarding the electrical connections of a stereo cartridge. At the present time, the ceramic and crystal stereo cartridges appear to be the most popular types with the cartridge manufacturers. The magnetic cartridge may have an advantage performancewise, but will also present it's own problems of extra amplification. The crystal or ceramic cartridge is basically a capacative generating device.

With the 3 element design of a stereo cartridge, the generating capacitance will run as low as 400 micro-micro farads, for each element. Since some of the proposed stereo cartridges will be of higher impedance than the conventional lateral cartridge, a review of the loading problem is in order. High capacitance or long cable, will reduce the output level of the cartridge, since the generating capacitance and the cable capacitance will constitute a voltage divider. The cable capacitance, however, will not have a direct effect upon the frequency response. With the stereo cartridge, care must be taken that the cable capacitance on each element is equal and as small as possible.

There will be a measurable capacitance between conductors as well as to the ground; particularly where a common shield is used. To prevent excessive crosstalk, the capacitance between conductors must be small, as compared to the combined capacitance of one element and one conductor to ground. Since the element capacitance must be considered, it should be noted that the crosstalk will increase with length of cable until the conductor to shield capacitance is large compared to that of the element. This will result in an unsatisfactory condition.

The bass response of the cartridge will roll off at 6 DB per octave at frequencies below which the load impedance equals the combined element and cable capacitance. Therefore, with a given cartridge and cable, a decrease in load impedance will cause a decrease in bass responce; conversely, an increase in load impedance will cause a bass boost with corresponding rumble and mechanical feedback difficulties. With a given cartridge and load, a higher capacitance or longer cable will give a comparative bass boost in that the bass roll off will occur at a lower frequency and that the higher frequency output level will decrease. It is apparent that the loading conditions upon both elements should be as nearly matched as possible.

Amplifiers

At this point in the system, there are several approaches that may be considered. (1) A complete stereo amplifier on a common chassis with a common power supply. In designing this type of amplifier, it is necessary to consider the input circuit as having very small interchannel capacity, and also to consider the separation of the B voltage filter unit. At these two points, it is possible to have interchannel reaction, thus reducing separation of the two channels. A 3 terminal cartridge works very well with this type system. (2) Another approach is to have two complete and separate amplifiers. Crosstalk in this type of system is less likely to appear. (3) Still another system that sounds very well is where the bass speaker is in the center and mid-range and high frequency speakers are to the left and right. With proper equalization, this system can be made to sound like 3 channel stereo and a larger separation may be obtained without the apparent loss of sound between. Probably the most common system will be the 2 speaker system, which has the most versatility of placement in the room.

Cabinet

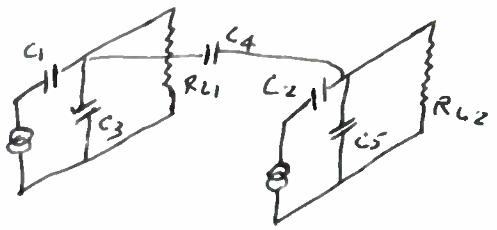
The cabinet for containing one leg of the system and the record changer should be well constructed. As the changer itself is more subject to mechanical feedback, it may be well to consider that the changer mounting board be glued and fastened securely to the cabinet, so that is is an intrical part of the cabinet. This will add to the strength of the complete unit. Baffles should be securely fastened to the cabinet, again making a stronger overall unit, with less chance of mechanical feedback. The speaker placement in the cabinet should be considered in the overall design, and speakers placed in such a way as to have minimum air coupling from the back of the speaker cone to the bottom of the record changer itself. If this is impossible to do, it is well to investigate the possibility of porting the cabinet in such a way as to have equal sound pressure on top and bottom of the record changer base.

There will probably be other things to consider which cannot be determined at this time, but will become apparent when we have mass produced cartridges, changer, and complete systems. The industry has been perfecting the monaural cartridge over a period of some 20 years, so it is not likely that we will have perfection of the stereo disc system for sometime; however, at it's present stage it is quite satisfactory and makes very enjoyable listening.

I would like to extend my thanks to Mr. Robert Hammond and Mr. Robert Van Antwerp for their assistance in the compiling of the above notes.

DIAGRAM FOR CAPACITY AND IMPEDANCE MATCH BETWEEN CARTRIDGE AND AMPLIFIERS:

Consideration of the diagram shown below will facilitate any necessary computation.



Wherein: C1 and C2 are element capacitance.
C3 and C5 are cable capacitance.
RL1 and RL2 are amplifier input impedances.

DISTORTION IN AUDIO PHASE INVERTER AND DRIVER SYSTEMS

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SUMMARY

The distortion produced by certain widely used audio phase inverter and driver systems may be reduced considerably by changing the type of tubes used in the circuit or by making small changes in circuit component values. The decrease in driver distortion following these changes can result in a lower distortion output from the complete amplifier and a greater tolerance of the amplifier to a change in operating voltages or a change of tube characteristics.

A great deal of effort and money are spent on output transformers and associated output amplifier components in order to secure audio power that has a minimum of distortion. This effort and expenditure is negated to a degree when the remainder of the system is not as highly refined. This is an unfortunate situation because except for the transducers, other parts of the system are reasonably inexpensive and their refinement can be accomplished without much change in cost.

Fig. 1 shows the distortion characteristic of a Williamson type amplifier which was constructed with 12AU7 tubes in the driver circuit in place of the 6SN7's usually employed. The measured distortion in this amplifier was considerably higher than might be expected in a high quality amplifier. New tubes were substituted for the original ones without producing any significant change in the distortion. Other circuit components were checked for value and were found to be correct. Having previously found that 12AU7's produce more distortion at a given output than do 6SN7's, a pair of 6CG7's was substituted for the 12AU7's with the improvement which can be noted in the figure.[1] 6CG7's are exact electrical equivalents of 6SN7's and with only a change of heater connections at the sockets they may be substituted for 12AU7's. It should be kept in mind that they draw 300 ma. more heater current but in most cases this will not be a matter of importance.

A change of duty station soon after working on the amplifier just mentioned, delayed further study of the problem. About a year and a half later a more thorough study of the reduction of distortion in phase inverter and driver systems was begun. The study was limited to systems which would be capable of driving the newer high power tubes. A supply voltage of 400 was used and except for one circuit, the output voltage was applied to 100 K grid leaks. The input signal was 60 and 6000 cps at a 4 to 1 voltage ratio. The distortion is plotted against peak output thus allowing easy reference to the level needed for maximum amplifier output. The peak voltage required for maximum amplifier output is, of course, equal to the bias on the output tubes.

The system set up for the tests is shown in Fig. 2. An intermodulation signal generator is used to furnish the test signal, which is fed into the circuit under test. The output of the circuit under test was fed into a differential amplifier, which was designed to discriminate against any in phase components of the two signals which were applied to its input terminals. By this means this unit eliminated the portions of the even harmonics, which in a complete amplifier unit would be balanced out in the output transformer. The output of the differential amplifier was connected to the intermodulation analyzer from which the readings were taken. The voltmeters and the oscilloscope were used to measure input and output voltages and to make visual observations of the waveforms.

A test model of a Williamson driver system as shown in Fig. 3 was constructed so that either octal or miniature tubes could be plugged into the circuit. The distortion characteristics of this circuit with 12AU7's and with 6SN7's installed are shown in Fig. 4. The curves show that the 6SN7 gives up to a 3 to 1 improvement over the 12AU7. An increase in the bias resistor on the push-pull driver stage helps a little in the case of the 12AU7's but the 6SN7's still show a great advantage.

Figs. 5 and 6 show comparisons of the distortion present in each of the halves of the driver system with the distortion present in the push-pull output. The great amount of distortion which must be balanced out emphasizes the necessity for maintaining balance in the push-pull output stage including the output transformer.

On the assumption that the simplest circuit which will do a job well is the best, the next circuit to be tested was the 6AN8 connected as a pentode voltage amplifier and a triode split load phase inverter. The circuit is shown in Fig. 7. This circuit was built up and a series of measurements were made. The circuit was modified to work with other types of tubes and then reconnected for the 6AN8 to get some additional information.

Some of the results obtained during the second series of measurements were as much as 300% different from similar measurements made during the first series. Such an occurence is very discouraging to the experimenter unless a reason for the difference can be found. After considerable investigation it was found that the difference between the two series of measurements was the heater voltage applied to the tube. A new series of measurements were made with three different heater voltages. As can be seen from Fig. 8 this particular circuit using the 6AN8 is sensitive to changes in heater voltage.

Since it is difficult to control heater voltage in the usual set up, other means were sought to keep the distortion at the lower levels. It was found that increasing the pentode cathode resistor to 1.2K reduced the distortion over most of the useful range. Fig. 9 shows the characteristics of the circuit with the higher value cathode resistor.

The higher value of cathode resistor reduces the sensitivity of the circuit so that the driving voltage necessary to reach full output is increased by about 25%.

Other tube types were tested in this general circuit configuration and although some of them were less sensitive to heater voltage variations, none of them offered any significant advantage over the 6AN8.

The third complete system tested was an American variation on the Mullard circuit as shown in Fig. 10. This circuit consists of a pentode voltage amplifier driving a long-tailed pair phase inverter. As originally developed in England, the circuit was used to drive a pair of output tubes having a very low driving voltage requirement and having a high value of grid resistor. Also the long tailed pair used a high mu twin triode. The modification used in this country substitutes a medium mu twin triode for the high mu twin triode and used the resultant circuit to drive a pair of the new high power pentodes which require something on the order of 40 peak volts to be driven to full output. The result of these changed conditions is to require a much higher voltage out of the pentode voltage amplifier. This in turn raises the amount of distortion produced by the pentode amplifier.

As can be seen in Fig. 11 a great deal of this distortion can be eliminated by disconnecting the pentode amplifier cathode by-pass capacitor. An even greater reduction may be secured by also increasing the value of the cathode bias resistor to 3.9K. As in the case of the 6AN8 this increased bias resistor lowers the gain of the circuit thus requiring a greater input voltage for a given output. There seems to be enough gain remaining to allow a complete amplifier to be driven from any usual pre-amplifier.

Some work as done to determine the contribution of individual stages to the total distortion of a system. It was, of course, simple enough to determine the distortion produced in the individual stages but no way was found to predict the combined effect of two or more stages in cascade. Fig. 12 gives the distortion produced by various tubes connected as split load phase inverters. It can be seen that the 6SL7 gives only 1/10 to 1/4 as much distortion as the 6AN8 triode, however, when the 6SL7 was combined with the 6AN8 pentode the resulting combination gave more distortion than did the original circuit using the 6AN8 triode. The 6SN7 long tailed pair phase inverter alone produced less distortion with circuit component values different from those used in the circuit in Slide 11, however, when it was combined with the EF86 amplifier, these different values did not give any appreciable reduction in distortion.

Some measurements were made of high mu triode voltage amplifiers with split load phase inverters and with long-tailed phase inverters. Some of these combinations gave reasonable good distortion characteristics but in general the high mu triodes operating at low plate voltages were very sensitive to changes of signal source impedance and to the value of the first grid resistor. In connection with this the possibility of grid circuit distortion made these circuits relatively unattractive.[2]

Overall feedback in the finished amplifier has no effect on the grid circuit distortion contributed by the first tube and therefore on the assumption the 20 db. of feedback will be used, it is about ten times as undesirable as other types of distortion of the same general order.

If we assume that 1% IM in the driver system is a satisfactory level it can be seen that any of the three systems investigated will serve satisfactorily if the recommended modifications are made. On the basis of the difficulty of stabilization when feedback is applied the Williamson type loses out to the other two because it has an additional time constant or phase shift at both high and low frequency. Also it places the burden of the cancellation of even harmonic distortion on the output stage. The split load phase inverter circuit will have less phase shift than the long tailed pair however on the other hand the long tailed pair puts less of a requirement on the pentode voltage amplifier driving it than does the split load circuit.

A study of the information in Fig. 12 showed that it should be possible to build a driver system that would not produce more than 0.1% IM distortion which is the minimum measurable on the analyzer used in these tests. This objective was compromised somewhat in order to use only two tube envelopes in the system. The resulting circuit is shown in Fig. 13. It consists of a 6SH7 pentode voltage amplifier, a 6SN7 triode voltage amplifier and a 6SN7 split load phase inverter. Approximately 20 db. of feedback is applied from the cathode of the phase inverter. Fig. 14 shows the distortion characteristic of the feedback driver circuit. The use of only one half of the 6SN7 for the phase inverter and coupling it to 100K grid resistors did not give as high a voltage output as was planned. The addition of another tube or the substitution of a 6BL7 for the 6SN7 might increase the voltage output that can be reached before the distortion exceeds 0.1%.

To prove that the feedback type driver is practical, it was substituted for the driver in a Williamson type. Fig. 15 shows the distortion curve for the entire amplifier. The second curve on Fig. 15 shows the effect of omitting the cathode by-pass capacitor on the output stage. It appears that the amplifier can be stabilized as easily as can others. The response and phase characteristic of the driver can be shaped by connecting an RC network from the plate of the triode voltage amplifier to the cathode of the pentode.

Some of the pitfalls confronting the experimenter might well be mentioned. A diode rectifier type voltmeter connected across a high impedance circuit will generate a considerable amount of IM. An amplifier type voltmeter should be used on all critical circuits. As mentioned previously certain tubes may create distortion in the grid circuit even though the grid is at a negative potential. In some cases the moving contact in a potentiometer will offer a non-linearity. If a very low variable voltage is required, it is preferable to operate the potentiometer at a reasonably high level and then to divide the output from the potentiometer with a network of fixed resistors.

Although this study is far from complete, it is hoped that it clearly demonstrates that further progress may be made in this field. It is hoped that interest has been generated to cause others to carry on work on this subject.

1. Bernard, Distortion in voltage amplifiers, Audio Engineering, Feb. 1953

2. Watkinson, Grid circuit distortion, Electronic & Radio Engineer, June 1957

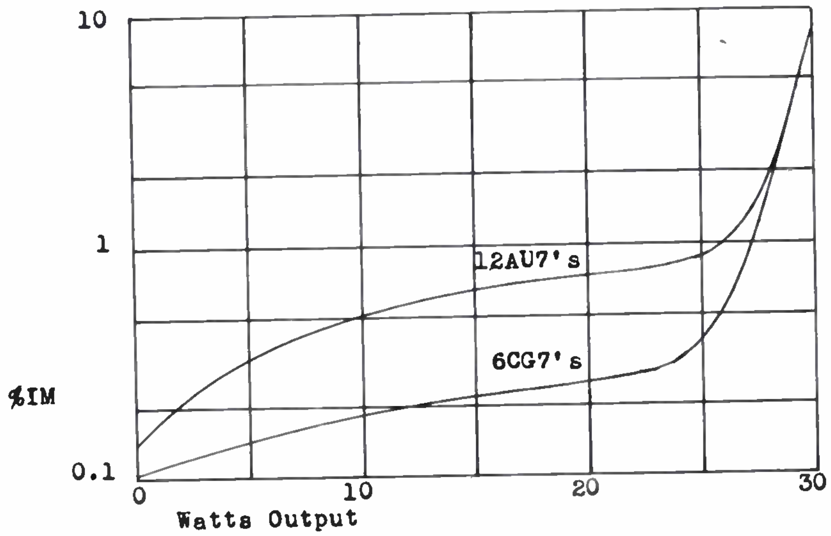

Fig. 1 Distortion in Williamson-type amplifier showing effect of changing tube types in driver circuit.

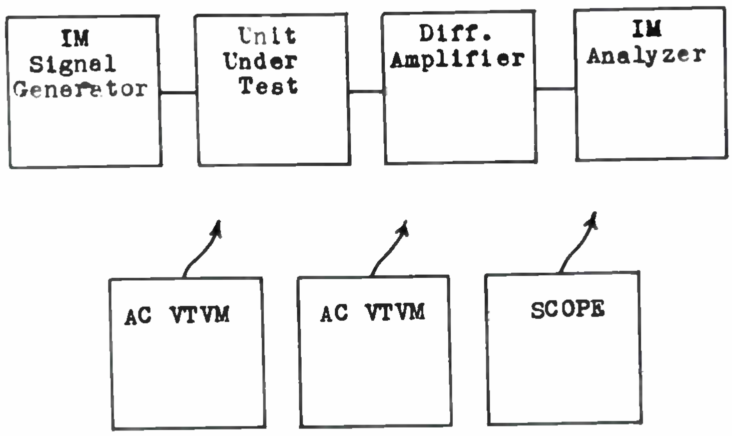

Fig. 2 Test setup for measurements.

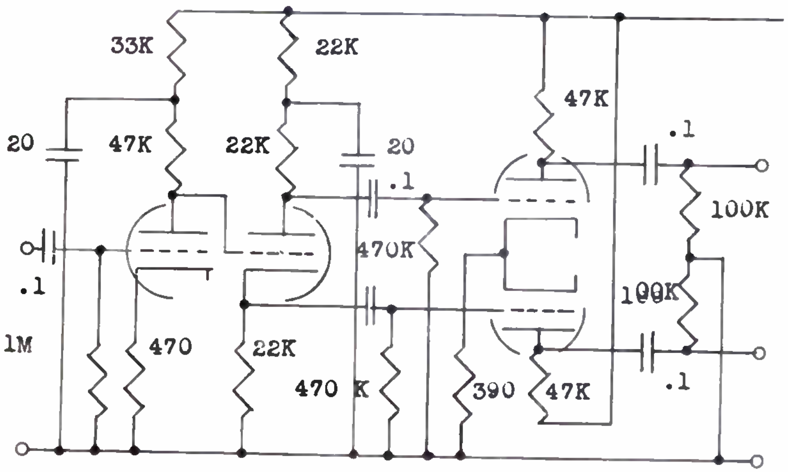

Fig. 3 Williamson-type driver circuit.

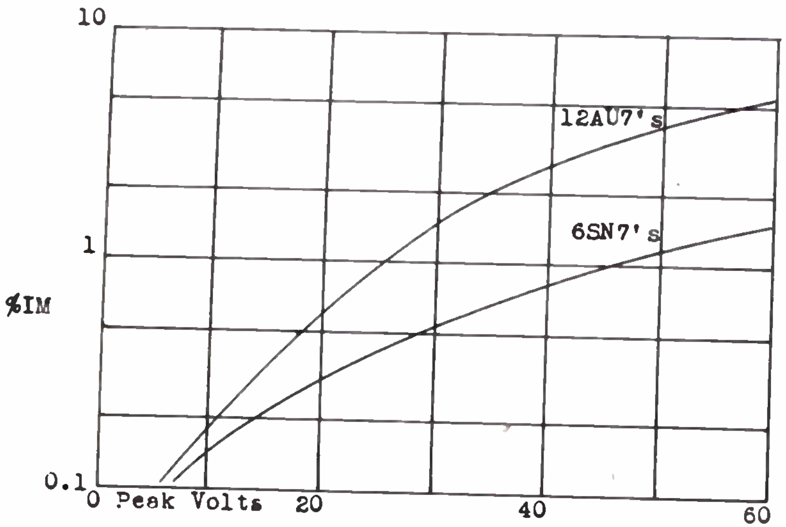

Fig. 4 Distortion in Williamson-type driver circuit showing effect of changing tube type.

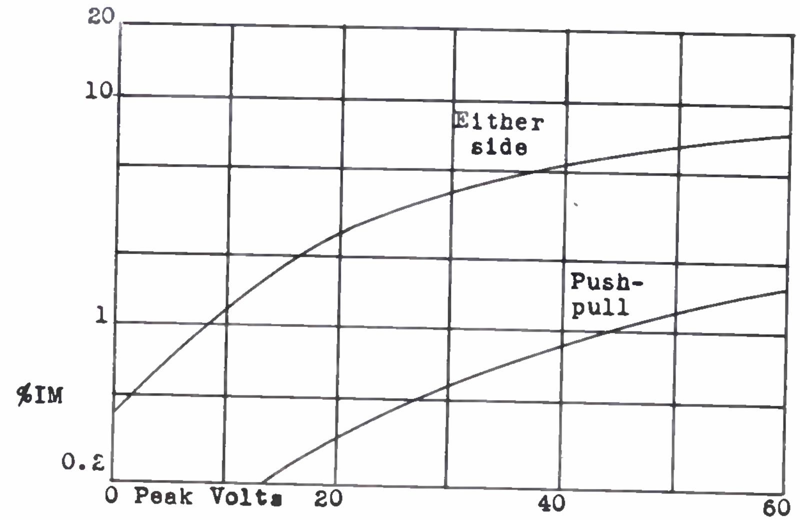

Fig. 5 Distortion in Williamson-type driver system employing 6SN7 tubes showing the cancellation of distortion due to push-pull action.

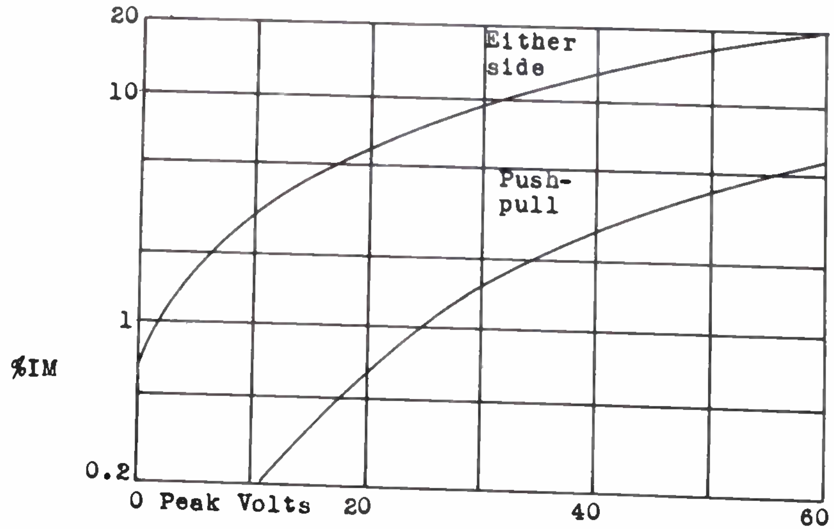

Fig. 6 Distortion in Williamson-type driver system employing 12AU7 tubes showing the cancellation of distortion due to push-pull action.

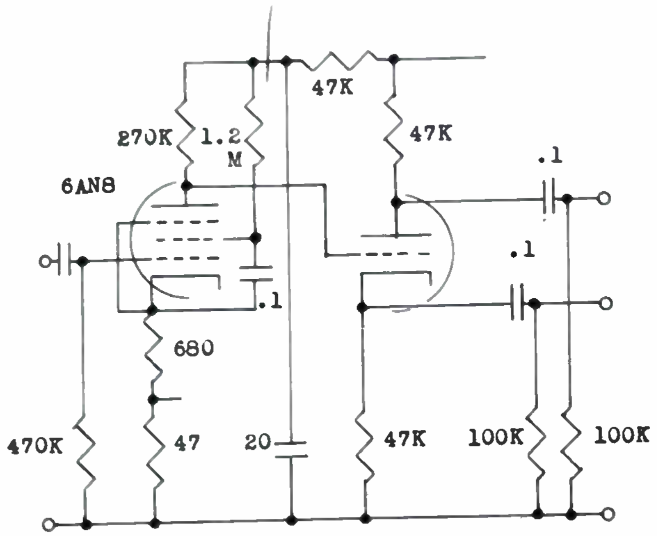

Fig. 7 6AN8 driver circuit.

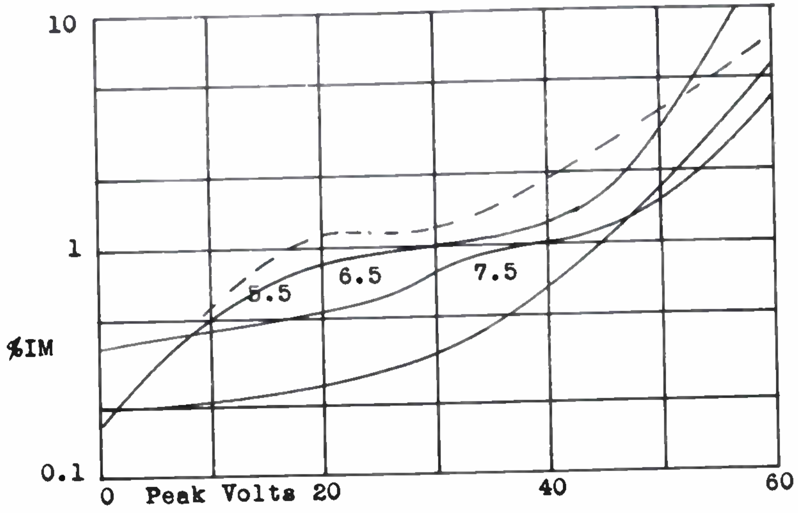

Fig. 9 Distortion in 6AN8 driver circuit showing the effect of variation of heater voltage, RK 1.2K. Dotted curve is 6.5 v curve for RK 680 ohms.

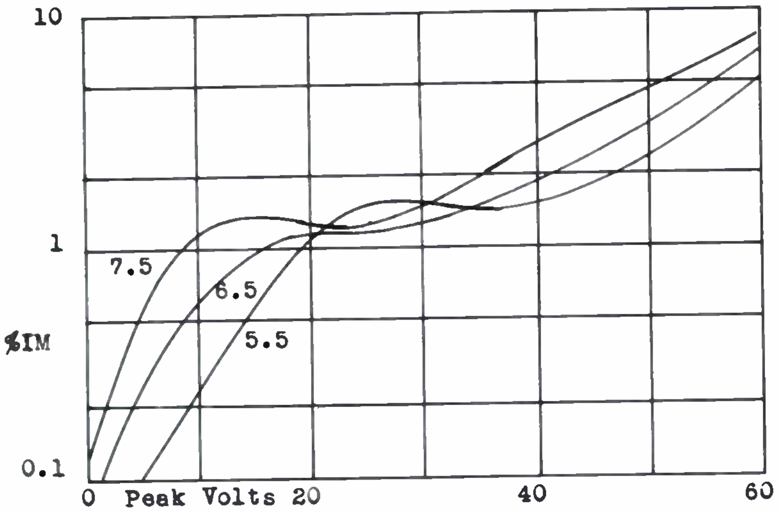

Fig. 8 Distortion in 6AN8 driver circuit showing the effect of variation of heater voltage, RK 680 ohms.

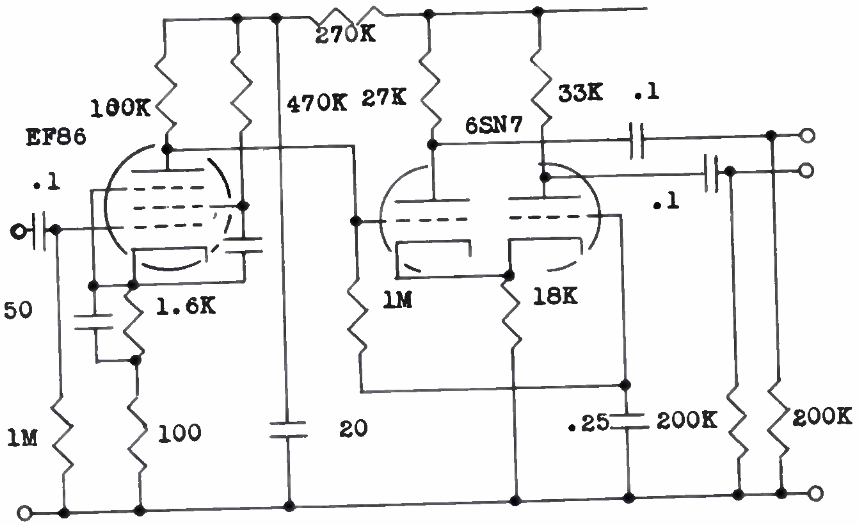

Fig. 10 EF86 amplifier with 6SN7 long-tailed pair phase inverter.

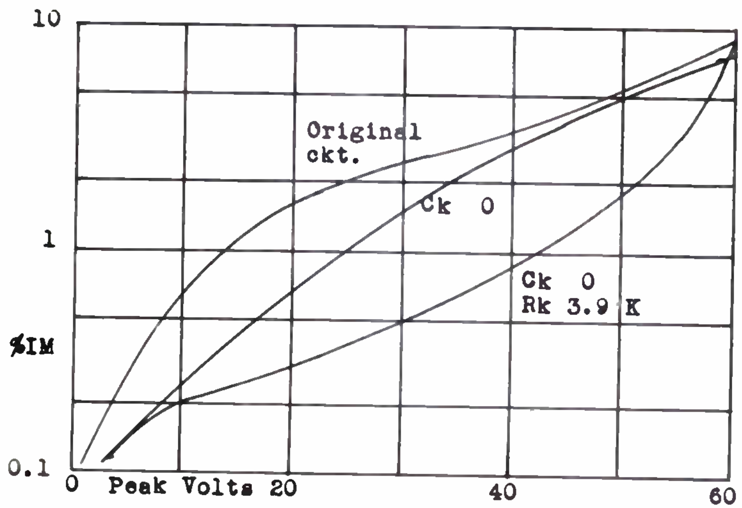

Fig. 11 Distortion in EF86, 6SN7 circuit showing effect of changing component values.

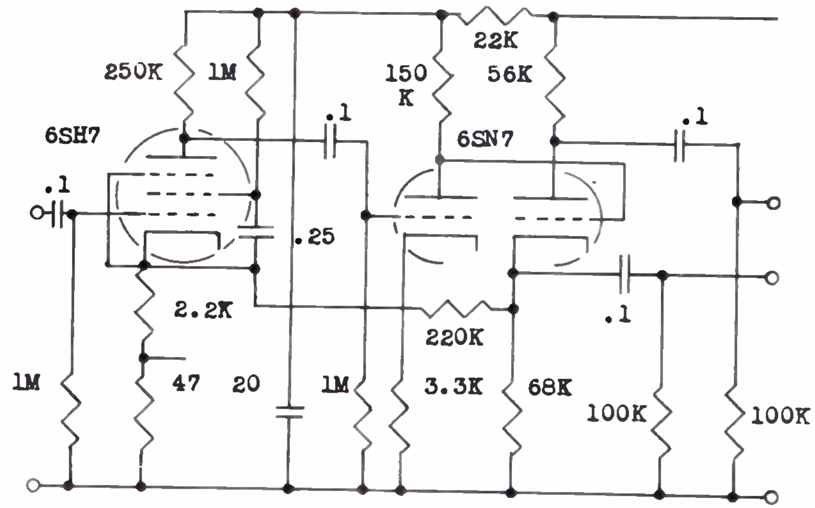

Fig. 13 Driver circuit consisting of feedback pair and split load phase inverter.

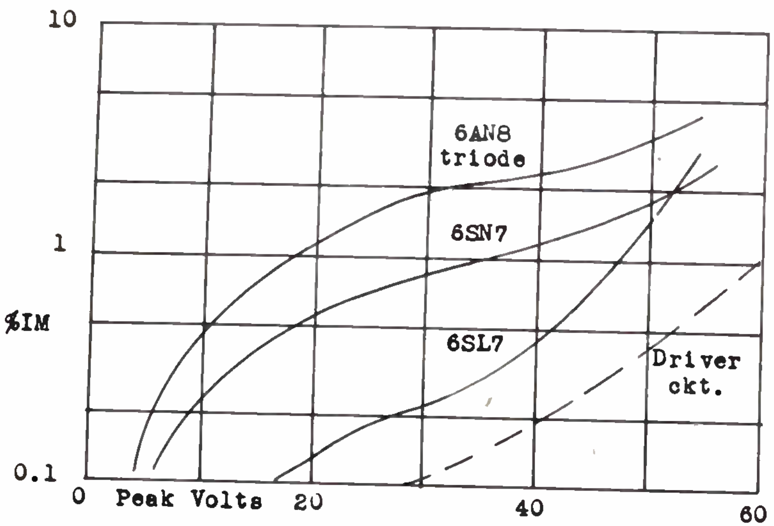

Fig. 12 Distortion in split load phase inverters.

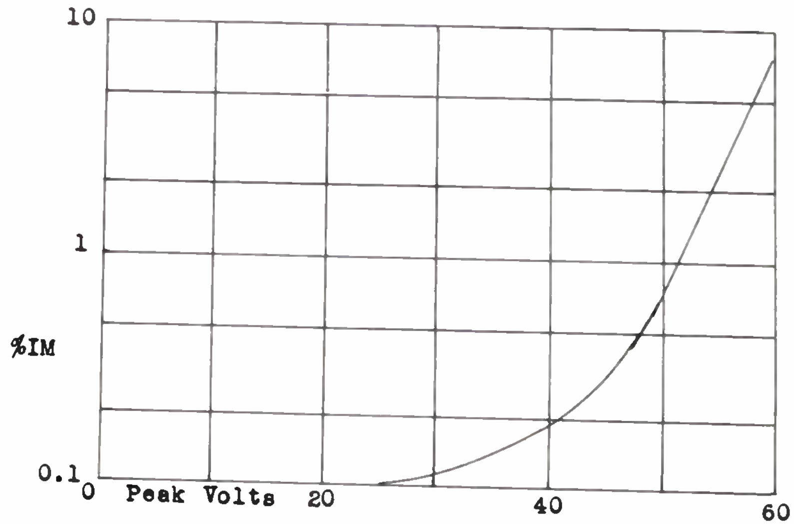

Fig. 14 Distortion in feedback-type driver circuit.

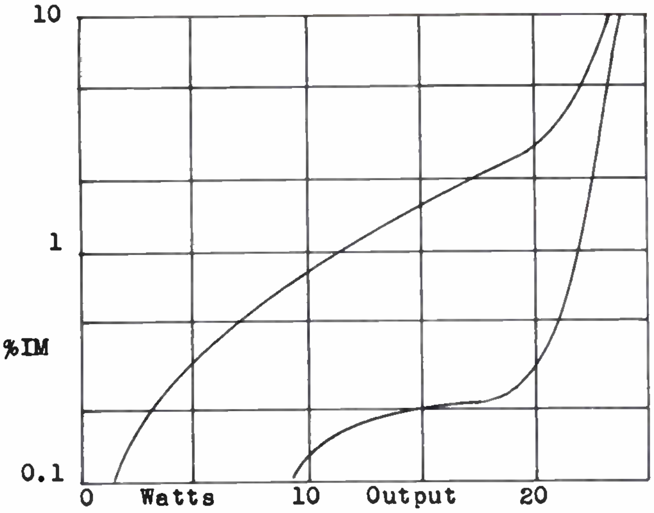

Fig. 15 Distortion in double feedback loop amplifier.

THE CBS COMPATIBLE STEREOPHONIC DISC

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CBS Laboratories
New York, New York

W. S. Bachman
Columbia Records, Inc.
New York, New York

(The text of this paper was not available at publication time.)

DESIGN OF A TRANSISTORIZED RECORD-PLAYBACK AMPLIFIER FOR DICTATION MACHINE APPLICATION

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Summary

The design of a transistorized dictation machine amplifier is discussed, with particular reference to the problems encountered in its development. Information on transistor noise, stabilization and feedback, and Class A and B power stages is included.

Introduction

In discussing the evolution of a high-gain special purpose transistor audio amplifier for the dictation instrument field, this paper is presented primarily from the product development viewpoint. Most of the problems to be discussed here are fundamental in nature and are likely to be encountered in the course of developing many types of transistorized equipment. Therefore, the emphasis will be placed primarily on the general type of problems and less on the details of this specific design. We realize that many of the problems under discussion are to be expected in the relatively new and fast-moving field of transistors and their many circuit applications. To those who have turned the transistor into a useful citizen in their electronics world, this paper may be unnecessary. To others, it is hoped that this brief case history of a transistor project will provide at least sufficient clues to point the way toward the high performance and reliability available in present-day transistors.

With an eye toward acquiring the experience necessary to encourage wider use of transistors in commercial equipment, three basic recommendations are made:

1. Institute a long-range transistor study project. Two objectives of such a project should be to (a) provide detailed information concerning trnsistors, transistor circuit performance, and associated circuit components; and (b) acquire within the organization the necessary first-hand experience so vital to the satisfactory application of transistors to a particular product design.

2. Application engineering services of transistor manufacturers are particularly helpful in preventing "misapplication" of transistors, correlating test and inspection techniques, and selection of the best transistor for a specific circuit use. Such application services should be used in conjunction with, not in lieu of, know-how within the organization itself.

3. Provide an internal training program so that design, product, and manufacturing engineering groups, as well as inspection and service personnel, can acquire a practical understanding of both transistors and transistor circuits, as well as specific information pertinent to the functions.

Equipment Design

That branch of audio which deals with the recording and reproduction of sound has, as a sub-group, the field of dictation and central-office recording. This general field has its own set of special requirements. Among these are simplicity of operation, reliability, uniformity of performance, ease of servicing, and rather surprising extremes in environmental conditions.

In most cases, the amplifiers have to "double in brass"; that is, they play the dual role of both record and reproduce amplifiers, usually with a wide variety of transducers. The input devices may span a range of impedances from a few ohms to several hundred thousand ohms. In many cases, the transducers are bilateral; a microphone in record becomes a receiver in playback. This immediately implies the necessity for switching input and output devices, gain controls, and equalization circuits. This, coupled with the high gain of these amplifiers and the extreme compactness usually necessary, creates many problems. Among them are stability, hum and noise pick-up, and temperature rise.

One of the most important requirements of the audio system in dictation instruments is that the highest possible degree of intelligibility and naturalness of speech recording and reproduction be provided. While the bandwidth is modest compared to that of a high-fidelity system, the questions regarding recording time, small loudspeakers, uncontrollable microphone techniques, efficiency, reliability, small size, low cost, etc. create a set of design problems as challenging as those encountered in other audio systems.

The first illustration is a composite internal-external photograph of an ultra-compact transistorized dictation instrument known as the Gray "Key-Noter". This unit, developed in the Gray engineering laboratories during the past year, measures approximately 8-1/2" x 6-1/2" x 2-3/4" and weighs 5-1/2 pounds. Employing the Gray constant linear groove velocity embossed recording technique, 20 minutes of recording time are provided using a 10 mil vinyl disc less than 6" in diameter.

General Design Specifications

Specifications covering several types of amplifiers developed for this unit are listed in Fig. 1.

Class A, 4 Stage Amplifier

The amplifier circuit to be discussed first (Fig. 2) uses four transistors, all common emitter stages. The output stage is operated in linear Class A and uses a power transistor of the automobile radio type. Each stage is stabilized, employing both D.C. and signal negative feedback. As stated previously, the overall power gain of this system is in the order of 93 db. This amplifier can be operated directly from the power line, by means of its own power supply, directly off a nominal 14.4 volt D.C. supply or a 12 volt automobile system. The first three stages are NPN transistors; the fourth stage uses a PNP power transistor.

In both this and the Class B amplifier to be discussed later, individual stage feedback is used, proportioned so as to serve for bias stabilization and signal feedback. The thermal stability of each stage of this amplifier is excellent without resorting to the use of non-linear elements in the bias networks. A few points worthy of note in the Class A power output stage, regarding both thermal and signal stabilization, may be of interest.

Emitter degeneration is employed in all stages. Collector-to-base feedback, D.C. and signal, is used. The transistor is operated with a collector dissipation that does not exceed 3 watts at zero signal conditions. It is mounted on a substantial aluminum bracket which, as may be seen in Fig. 1, is thermally coupled to an aluminum frame, the entire assembly serving as a heat-sink.

We found it advisable to use mica insulators between the mounting flange of the power transistor and the aluminum heat-sink. The small reduction in heat transfer was more than compensated for by the electrical reliability provided by the mica spacer. Anodized aluminum heat-sinks appeared to be vulnerable to any burrs on the transistor mounting surface.

The high degree of stabilization employed in this amplifier has permitted the use of transistors with fairly wide variations in current gain, while at the same time achieving extremely good uniformity in overall gain with production units. For example, a large group of production amplifiers showed a gain spread of only about 1.5 db at 1 kc. The uniformity in gain and stability, in spite of the high power gain, is achieved by the use of feedback, a printed wiring board, a carefully laid-out grounding system, and good component control. To those experienced with very high-gain amplifiers, there is no need to emphasize the need for a carefully designed grounding system to achieve low hum levels and freedom from oscillation.

We believe that it would not have been practically possible to meet the requirements of this compact, light-weight instrument with a vacuum-tube amplifier.

Class B, 3 Stage Amplifier

Fig. 3, which illustrates a 3 stage Class B transistor amplifier, is pointed up for many reasons. The output stage is a linear Class B push-pull stage using two PNP transistors, the driver and input stages use NPN transistors, and the entire amplifier can be substituted for the one shown in Fig. 2.

There are several points of interest in this Class B amplifier. We will start by stating that the same overall power gain and bandwidth are required, and the same transducers are used. In general, amplifier "A" can be replaced with amplifier "B", and one stage is eliminated. To provide virtual interchangeability of amplifiers and yet eliminate one stage means that additional gain must be picked up; yet stabilization has to be maintained. This amplifier is different from the Class A amplifier of Fig. 2 in that little feedback is used except in the Class B stage. The biasing circuit of the first two stages is conventional and will be discussed under the subject of "stabilization and feedback". Feedback is removed so that the first two stages largely make up the gain increase required. Due to the low dissipation of the first two stages and D. C. emitter degeneration, thermal stability is not a problem in these stages. The Class B stage is operated with a quiescent collector current of approximately 1.5 to 2.0 milliamperes. The base emitter bias consists of a temperature-sensitive germanium diode in a voltage divider circuit. Additional degeneration is employed in each emitter leg by means of 6.8 ohm resistors. We believe that, with diode biasing, much improved temperature stability is attained.

The resistors in the emitter legs and heat-sink mountings for the transistors provide good stability under conditions of 2/3 to maximum power output. Unlike the Class A stage, whose greatest thermal problem is at zero signal, the Class B stage should be evaluated at power levels between approximately 1/2 power and full power. It would be well to plot an efficiency curve of the Class B output stage. This amplifier has been driven close to full output from a sine-wave source continuously for many hours without any tendency for run-away.

When comparing the 4 stage Class A with the 3 stage Class B output circuit, we must point out that we have both gained and lost by eliminating one stage while maintaining the overall gain requirements. First of all, the amazing uniformity in production units attained by the 4 stage amplifier cannot be expected with the 3 stage amplifier by virtue of the fact that signal feedback has been nearly eliminated. When this high degree of uniformity is not essential, the 3 stage unit may be completely satisfactory. The main point is to examine what has been gained and lost in comparing the two amplifiers. The big gain obviously is the cost savings realized by eliminating one stage with its associated components.

In view of the foregoing, a 4 stage Class B amplifier was designed, utilizing essentially the same stabilization embodied in the Class A unit. For the Gray "Key-Noter", uniformity and stability are considered to outweigh the cost of the extra components. This approach permits reasonable current gain tolerances in the transistors.

Components

It is beyond the scope of this paper to embark on a discourse on components. We will restrict ourselves to those components that created the biggest design problems. In order of importance, we would list first the transformers and second the transistors.

Class B Output Transformer. Without a doubt, our biggest single component problem was the Class B output transformer. This put a restriction on not only the power output from our Class B stage but distortion and bandwidth as well. It is a sad commentary that we must make regarding the claims for commercial transformers, advertised and specified for use in Class B circuits, of the 3/4 watt power output class. Of the many makes and modèls bought in the open market for this type of output circuit, not one lived up to the manufacturer's claims. Not until we had a particular transformer manufacturer design a unit exactly to our requirements were we able to obtain the high efficiency, low distortion, bandwidth, and power output that should be obtained with the transistors used. Apparently, this has been a universally sad experience; and, for that reason, we point up this problem since many others may encounter it. The moral of the story is: do not be discouraged with the transistors until it is certain that the transformer performance is adequate for the application.

An illuminating way to illustrate the point is as follows: If the Class B stage in Fig. 3 is connected as a shunt-fed output stage properly terminated and then used as a conventional transformer-coupled stage, the difference in performance may be startling. With transformers of this type, we would suggest that the following points be carefully determined so that satisfactory design specifications can be written.

The first known factors are assumed to be power output, collector supply voltage, collector-to-collector load impedance, bandwidth, and distortion at rated output at specified conditions. The transformer specifications should include at least the following data:

Primary Winding. Impedance source and load; taps, if any; D.C. resistance (restrict to 5% of impedance); bi-filar or standard winding; incremental inductance; coupling and leakage inductance; primary current; maximum average D.C. current; peak current per winding; and degree of primary current unbalance.

Secondary Winding(s). Impedance ratios; termination; power distribution; D.C. resistance (10% of impedance); taps and electrical balance; and D.C. currents.

Power Rating. Power output - level and termination of each winding; power response and distortion (type, operating conditions).

General. Core area; lamination size; winding polarities; impregnation; dielectric tests; finish; lead termination; and circuit application.

In general, we would recommend a high grade lamination steel (grain-oriented XXX for example) since the modest cost increase is outweighed by the better performance.

The placement of windings in relation to the core and the coupling between windings is extremely important in the Class B amplifier due to the switching problem between transistors. It may take some experimenting to determine the best location of each winding with respect to each other and the core. The only conclusive way to determine results is by careful study in the laboratory with the actual circuit arrangement. Having once finalized on prototype transformers, it then becomes necessary to clearly specify the transformer requirements and ensure that realistic test methods determine that these requirements are met.

Driver Transformer. The driver transformer problem is similar to that of the output transformer. We might add these two comments: The amount of D.C. magnetizing current in the primary should be determined on the basis of optimum operation with the driver transistor involved. The secondary impedance should be carefully determined from the standpoint of good transfer and low distortion when driving the push-pull inputs of the power stage.

Power Transformer. It may be interesting to note from Fig. 3 that this entire amplifier operates from a conventional-looking power supply. When space, cost, efficiency, and temperature are as important as they are in a compact portable instrument of the Gray "Key-Noter" type, it is seldom possible to resort to the luxury of a regulated power supply.

Fig. 4 illustrates what can be done with a conventional transformer, bridge rectifier, R-C filtered supply to improve regulation and efficiency - in fact, to achieve regulation adequate for the Class B amplifier mentioned previously. With instruments of this type, it is not uncommon to use small, governor-controlled D.C. motors. These motors usually have governing which is greatly aided by good supply voltage regulation.

To cite a specific case, Fig. 4 illustrates the regulation curve (a) for a power supply satisfactory for an application such as the Class A amplifier and yet not adequate for the Class B power amplifier of Fig. 3. Better than a two-to-one improvement was achieved, as may be seen by comparing curves (a) and (b) of Fig. 4.

Transistors

There are a few general comments to be made before we get into specific problems with transistors. We would like to encourage all efforts towards standardization with transistors. For example, in the Class A output stage illustrated in Fig. 2, there are about six transistors manufactured by as many companies that could be used in this stage. It is rather like having a non-standard 12AU7 vacuum tube. The parallel drawn here with power transistors is the fact that, while these six power transistors may be practically interchangeable electrically, they have different base-flange thicknesses, overall heights, etc. Each manufacturer seems to have his own manner of specifying performance or measuring parameters. This leaves equipment manufacturers who live by the philosophy of two sources of supply faced with the problem of setting up their own composite specification for a particular transistor. There are several other examples of transistors that are electrically interchangeable but physically different in housing, basing, case size, and the method of specifying performance. We can only encourage the standards groups in their worthwhile efforts.

Transistor Tests. The following comments on transistor testing are made with the particular interest of receiving-inspection departments in mind. In most cases, it is not necessary to measure the hybrid r or π parameters. It is more important to measure such factors as small-signal current gain (A.C. beta), input impedance, collector-to-base leakage current (I_{co}), emitter-to-base leakage currents (I_{eo}), collector-to-emitter leakage current with a specified resistance between the base-emitter junction ($I_{c'}$), large signal current gain (D.C. beta), power gain under given operating conditions, and the cut-off frequency.

Fig. 5 illustrates, in schematic form, a test set suitable for measuring small-signal beta, input impedance, I_{co}, I_{eo}, $I_{c'}$, and D.C. beta for transistors up through the medium power types. A description of the test methods may be in order. The small-signal beta measurement and H_{11} are made with the collector circuit A.C. short circuited, a constant collector voltage, and a specified collector current. The input signal current is coupled in from a high-resistance source. I_{co} and I_{eo} are measured in the conventional manner with provisions for varying V_{cb} or Veb over wide limits. The D.C. beta is measured in the common emitter configuration and is the ratio of I_c/I_b for a specified collector-emitter voltage. The advantages of such a test set are its simplicity of operation, the fact that it requires commonly available meters and generators, and it provides the facility of measuring any one or all of six parameters in one set-up. It is obvious that modifications in biasing arrangements can be made. Either PNP or NPN transistors can be used by reversing the supply voltage and the D.C. meters.

Fig. 6 is a simplified break-down of the transistor test set shown schematically in Fig. 5, and is self-explanatory.

Power Transistor Tests. Fig. 7 illustrates in three sections the tests necessary for the receiving-inspection evaluation of power transistors. Section (a) of Fig. 7 shows the test circuit for the measurement of power gain, power output, cut-off frequency, and distortion at rated input. It can also be used for D.C. beta measurements with minor modifications. The test circuit, as a part of the power gain measurement, also determines the A.C. input impedance of the transistor at a rated load and operating point. D.C. base-emitter input resistance can be measured with small modifications. Fig. 7b is a collector-to-emitter break-down voltage (V_{cer}) with a specified base-to-emitter resistance, collector load resistance, and collector-to-emitter source voltage. A minimum collector-to-emitter D.C. voltage is required here. Fig. 7c is a conventional measurement of emitter-base leakage current in the reverse bias condition (I_{eo}). It is convenient to incorporate all three measurements in one test set-up; they are separated here for the sake of illustration.

Noise in Low-Level Stages

Transistor Noise

This paper will be confined to a discussion of transistor noise from five viewpoints. It is the purpose of this paper to discuss only junction transistors as used in audio frequency amplifiers and similar circuits. It may not be known or accepted in many areas that junction transistors are, in some cases, equal to or better than "low noise" tubes for audio frequency applications. This is more significant when the generator to be used in the device is a low-impedance transducer. High-impedance generators impose a limitation on noise levels obtainable with even the best junction transistors.

Fig. 8 is a plot of noise figure vs. frequency for junction transistors derived from the literature. The area marked F_1 contains the so-called semi-conductor, flicker, or 1/f noise. This noise varies inversely with frequency and may start at frequencies as high as several thousand cycles in high-noise transistors or as low as 100 or 200 cycles in low-noise transistors. The F_2 area indicates a sort of transition zone between 1/f noise and broadband or "white" noise. In the F_3 area, the noise factor is again increasing. The start of this area probably varies from 100 kc to 500 kc. The F_3 area is the same as the white noise area in characteristic, but the noise factor is degraded due to a loss in stage gain. Fundamentally, the noise factor NF, db equals:

$$10 \log \left(\frac{S_1 / N_1}{S_o / N_o} \right)$$

where S_1/N_1 is the input signal-to-noise ratio and S_o/N_o the output signal-to-noise ratio.

If S_o decreases due to loss in stage gain, the overall stage signal-to-noise ratio will be degraded. The noise above the 1/f region (white noise) is primarily thermal noise from the base resistance and shot noise from the base-emitter junction and base-collector junction. The noise in the F_1 area depends on the transistor material, surface conditions, re-combinations, applied voltage, and 1/f.

To explore a little further the three types of noise with which we are concerned, we can add the following comments:

1. White noise is directly proportional to the D.C. emitter current and inversely proportional to the D.C. beta and base resistance.

2. 1/f surface noise is proportional to the squared value of D.C. emitter current and inversely proportional to frequency. 1/f leakage noise is proportional to the D.C. collector-base voltage (V_{cb}) and squared value of leakage current (I_{co}).

3. The noise in the F_1 area is apparently composed of three components - 1/f surface noise, 1/f leakage noise, and thermal noise. It is interesting to note that thermal noise and leakage noise both show an increase with temperature; while surface noise is insensitive to temperature. We might also point out that some transistors show a marked increase in noise level after about one hour or more operation; while others show almost no change in noise level with operating time. As would be expected, transistors that show a very minimum of 1/f are the most stable.

Transistor Operating Conditions. Since white noise, 1/f surface noise, and 1/f leakage noise are all affected by the D.C. operating conditions that the transistor "sees", we cannot emphasize too strongly the necessity for selecting the best operating conditions if the highest signal-to-noise ratio is to be obtained.

Let us consider the transistor as two separate diodes operating in opposite modes. We have the forward biased emitter-base junction with its current-squared noise. Then there is the reverse biased collector-base diode with its voltage-dependent noise (leakage noise). Thus, the quiescent emitter current should normally be below 1 milliampere, preferably .2 to .5 milliampere. This step alone tends to minimize white noise and 1/f surface noise. 1/f leakage noise acts as a noise generator that is dependent upon the transistor D.C. voltage of the reverse biased collector-base diode. Experience has shown that the D.C. collector-to-base voltage should not exceed approximately 6 volts with present-day transistors. At voltages below this range, the 1/f leakage noise is independent of collector voltage. Therefore, the low-level transistor stage should be operated at D.C. emitter currents below 1 milliampere and D.C. collector voltages not exceeding about 6 volts.

Note that a two-to-one increase in emitter current could result in a 3 to 6 db increase in noise output, depending upon whether white noise or surface noise predominates in the particular transistor.

Transistor Circuit Conditions. Having selected the D.C. operating conditions for the transistor to obtain the best signal-to-noise ratio, we should look to the external circuit conditions to determine to what extent they contribute to the noise picture. Probably the most important point is the relationship of the resistive component of the generator impedance to the signal-to-noise ratio. There is an optimum range of generator impedances to be used if the highest signal-to-noise ratio is to be achieved. This generator impedance is in the order of a few hundred to perhaps 5,000 ohms. Note that any high resistance between the generator and the base input of the transistor acts to degrade the signal-to-noise ratio to some extent. Also, it would be well to point out that the generator should be used with as little attenuation as possible between the generator and the base input.

In connection with this, we would like to indicate that work in progress in our laboratory shows that a fairly wide range of generator impedances (100 to 1) can be tolerated if signal-to-noise ratios in excess of about 50 db are not required. Some papers have pointed out that an approximate 12 db improvement can be realized by keeping the resistance R_s between generator and transistor input small compared to R_g. We have not been able to realize a noise reduction of such magnitude. Our tests to date with both low and high-noise transistors show about 3 to 8 db difference with resistance in the base circuit ranging from 1,000 ohms to 1 megohm. We find that, if the base blocking capacitor is a D.C. electrolytic type, the signal-to-noise ratio may drop, as compared to that of a low-leakage paper capacitor or equivalent. This may be due to the effect of a small D.C. leakage current on composition resistors, the capacitor itself, or both. Any unbypassed resistors in the base-emitter or collector circuits should be scrutinized in attempting to achieve high signal-to-noise ratios.

Our main object here is to point up the fact that noise problems should not always be blamed on the transistors themselves.

Negative shunt feedback does not appreciably change the signal-to-noise ratio but is helpful in reducing distortion. Negative series feedback is not recommended. Feedback in general offers very little improvement in the signal-to-noise ratio of the stage; however, feedback can be used to good advantage indirectly in many cases to obtain improved signal-to-noise ratio by making it possible to reduce or eliminate R_g.

Noise Control and Parameter Control. Early in our investigation of transistor noise, it seemed reasonable to assume that transistors showing low collector-to-base and emitter-to-base leakage current, low I_c', and normal to high beta values should be the lowest-noise transistors. To our dismay, we found that this does not necessarily follow. In fact, some of the transistors we have selected for low noise levels show high values of leakage currents. Our purpose in investigating this approach was a simple one. If noisy transistors always showed correlation with such items as I_{co}, I_{eo}, low cut-off frequency, or I_c', then we might eliminate noisy transistors by the control of these parameters and avoid the problem of measuring noise as a part of receiving-inspection of transistors. We do not have sufficient information to indicate why this seemingly contradictory condition exists. We believe that it is necessary to measure the noise level of transistors whenever they are going to be used in low-level stages of equipment requiring signal-to-noise ratios in excess of about 40 db. An additional reason for bringing up the noise vs. parameter control is to solicit information from groups who may have done more extensive work in this area.

Transistor Noise Measurements. It may be well to point out again that this paper is written with the problems of the circuit designers and equipment manufacturers in mind, rather than those of the transistor manufacturer. How we use the device - not how it is made - is our primary concern.

Our objective here is to provide a simple, reliable means of measuring and specifying transistor noise in the audio frequency spectrum. A study of Fig. 9 illustrates the wide variation in the character of transistor noise as we have determined it by our laboratory measurements. What is not apparent from a study of this figure is the high peak factor of the noise in the 1/f regions. The peak/RMS ratio is highest in transistors with the highest values of 1/f noise. Transistors with predominantly white noise show the lowest and nearly constant peak/RMS ratio. An oscillographic observation of the noise patterns of both high and low-noise transistors demonstrates this condition.

It is the writer's opinion that transistor noise figures (NF), based on narrow band measurements at approximately 1 kc, are not adequate as a means of selecting transistors for wide-band audio frequency applications. We refer here to the so-called one cycle band noise figures at 1 kc, so widely publicized in connection with transistors. Our preference is for a noise-measuring device of the general type shown in Fig. 10, using a well defined bandpass measurement in the 1/f region and a peak-to-peak responding meter. For routine inspection, the set-up shown in Fig. 10 seems to be fast and reliable.

To describe briefly the transistor noise test set: The noise measurement is made in the common emitter configuration, using typical values of collector supply voltages V_{cc}, I_c, collector load resistance, and base series resistance. Typically, we would measure transistor noise with R_1 set at 1,000 ohms and .5 milliampere of collector current I_c. The noise output of the transistor is coupled to amplifier (A), then to a bandpass filter (F), and then to a peak-to-peak voltmeter. The voltage preamplifier (A) should be a wide-band, high quality, low noise preamplifier, providing uniform response from approximately 50 cycles to 20,000 cycles. The Gray AM-3 preamplifier is entirely satisfactory for this purpose. Filter (F) should be either an octave band or a 1/3 octave band spectrum analyzer. The three base resistors (R_1, R_2, and R_3) and the collector load resistor (R_7) should all be selected for stable, low noise characteristics even in the presence of small D.C. currents. Obviously, this test set can be used to check NPN transistors by reversing the battery supply and collector current meter. In order to illustrate the peak-to-peak vs. RMS ratios, meter M_1 is a typical vacuum tube voltmeter, essentially an average responding device calibrated to read RMS of a sine wave. Meter M_2 is a true peak-to-peak meter, and M_3 is an oscilloscope. Our recommendations for most laboratory testing would be to use a 1/3 octave band spectrum analyzer and a peak-to-peak vacuum tube voltmeter. M_1 is not necessary unless crest factors in the transistor noise are of interest.

For inspection use, the test set-up shown in Fig. 10 can be simplified as follows: The base resistor switching is not necessary, and a 1,000 ohm base resistor is recommended. Standard collector current of .5 milliampere is used. Amplifier (A) is already described. Filter (F) should preferably be either the 75-150 or the 150-300 cycle octave band of a standard audio frequency spectrum analyzer.

As shown by Fig. 9, low-noise transistors yield readings that are nearly constant in each octave band over the spectrum. However, high-noise units generally exhibit the rapidly increasing 1/f noise in the low frequency region. It is in this region that the crest-to-RMS ratio generally departs quite radically from the approximately 2.8/1 ratio encountered with broadband noise. Therefore, the three important requirements in our transistor noise set-up are peak-to-peak noise measurements, bandwidth of 1/3 octave or more, and the center frequency of this bandpass filter in the 75-300 cycle region.

An important reason for specifying peak-to-peak, or at least quasi-peak, meter is that this yields results which more closely correlate with listening tests.

We hope that it may be appropriate to suggest that the principles involved in this transistor noise test be considered as a standard method for routine inspection. Correlation between the user and the transistor manufacturer should be relatively simple to achieve and maintain. We realize that there are many possible variations in the biasing set-up and the possibility of introducing single frequencies to measure signal-to-noise. The collector supply voltage V_{cc}, the collector load resistance, and the collector current can be varied over a considerable range without changing the principles involved.

One important factor should be noted in this test set-up; i.e., that 1/f noise due to leakage currents is minimized by virtue of the fact that the collector-to-emitter voltage is well below 10 volts. This has been done deliberately for the simple reason that few circuit applications where noise is important use collector-to-emitter voltages higher than those used in this test set-up. In short, it judges the transistor fairly for the application intended.

Stabilization and Feedback

Biasing and Thermal Stabilization

I_{co}. If the collector-base junction were an ideal diode, the current I_{co} under reverse bias or cut-off conditions would be zero. Junction diodes to date have finite amounts of cut-off current I_{co}. In common emitter stages (to which our discussion is confined), the effect of I_{co} can be exaggerated by a factor approaching the current gain of the stage.

I_{co} is composed of two components - a thermally-generated current that is an exponential function of temperature and a voltage-dependent leakage current. The leakage current may also show some degree of temperature dependence.

Fig. 11a illustrates the need for a low resistance path between the base and emitter junction for the common emitter stage. If R_b is much greater than R_{be}, the thermal current of generator I_t flows through the base-emitter junction, acting as a forward biasing current. Because of the current gain (H_{cb}), $I_{c'}$ is approximately equal to $I_{co}H_{cb}$ or beta x I_{co}. To graphically show this effect, a plot of $I_{c'}$ as a function of R_b is suggested.

Fig. 11b, c, and d shows several biasing methods. It is beyond the scope of this paper to analyze them. The writer will try to emphasize a few guides toward satisfactory biasing and stabilization.

Base Current Bias or Fixed Bias. See Fig. 11b. This is the kind of circuit you would recommend only to your competition. Cheapest of all circuits, it requires only a 1/2 watt resistor. Drastic shifts in operating point with transistors as well as with temperature occur.

Feedback Self-Bias. Fig. 11c includes voltage feedback from V_c. This circuit is extremely stable if R_2 is made as small as possible without undue loading on the driving generator and if the collector D.C. load resistance is at least a few hundred ohms.

This circuit actually adjusts the base-emitter bias voltage. Any increase in I_c increases the I_cR_L drop; the voltage V_{cc} - I_cR_L is available to the divider R_2, R_3 to bias the transistor. The important factor is the inverse variation of base-emitter bias with collector current.

Self-Bias with Current Feedback. Fig. 11d is an excellent means of providing a well stabilized operating point and freedom from thermal run-away. With R_2 as small as possible and R_e as large as possible, either connection of R_3 provides good stability. (R_3 connected at the collector introduces signal feedback.) The D.C. base voltage can be kept relatively constant by the R_2, R_3 divider.

The I_eR_e voltage will change with any increase or decrease in D.C. emitter current, thereby providing an adjustable base-emitter voltage that serves to maintain the average emitter current substantially constant in the face of large temperature variations.

To summarize:

1. Avoid biasing arrangements that maintain or attempt to maintain constant base current.

2. The biasing objective should be that of maintaining constant emitter current.

3. Provide bias circuits that adjust the bias voltage V_{be} by (a) feedback methods, either voltage or current or both; (b) use of non-linear elements such as thermistors or diodes.

4. The biasing circuit and/or elements should adjust the base-emitter voltage at the rate of 2.0 to 2.5 millivolts per degrees C. for germanium transistors. This is an inverse relationship between bias and temperature. Therefore, the temperature sensing device should be germanium or possess a temperature coefficient of resistance substantially equal to that of germanium.

5. Thermistors or temperature-sensitive diodes should be closely coupled thermally to the transistor. This is necessary for accurate proportional temperature sensing of the transistor junction by the bias-regulating circuit element.

Class A and Class B Power Stages

It is the purpose of this section to highlight some of the important factors for stable and linear operation of Class A and push-pull Class B power stages. It is not the purpose of this paper to discuss in any detail the design data for power stages, as this area is well treated in current texts and papers.

Class A Power Stages

Dissipation and Efficiency. Probably the first considerations in a power amplifier stage are the maximum power rating of the transistor, maximum ambient temperature, maximum collector voltage rating, and the thermal resistance of the transistor.

Maximum collector dissipation P_c occurs when the driving signal equals zero. This means that, for Class A operation of a transistor, thermal stability problems will be most severe during those periods of zero signal operation. The maximum collector dissipation can be conveniently related to the junction temperature rating of the transistor, the ambient temperature, and the thermal resistance. The maximum dissipation, P_{max}, is

$$P_m = \frac{T_j - T_a}{R_t}$$

T_j = junction temperature, degrees C.
T_a = ambient temperature, degrees C.
R_t = thermal resistance in degrees C. per watt

A statement to the effect that a certain transistor will dissipate "X" number of watts has little significance by itself.

The maximum theoretical efficiency for an ideal shunt-fed or transformer-coupled stage equals 50%.

The collector supply voltage V_{cc} should not exceed 1/2 the peak collector voltage rating.

The load line should be tangent to the hyperbola representing maximum power dissipation. See Figure 12a, 12c. The practical limits on the slope of R_L can vary considerably provided that (a) the product of I_{co} and V_c max. does not exceed the power dissipation, (b) the collector current rating is not exceeded, and (c) the linearity of current gain is satisfactory.

In Fig. 12c, the requirements that determine the slope of R_L are power output and linearity. Therefore, $E_{c2} - E_{ca} = E_{ca} - E_{c'}$ and $I_7 - I_a = I_a - I_o$; or the voltage and current swings above and below the operating point should be as nearly equal as possible for the lowest distortion. (I_a is the operating current at the intercept of E_{ca} and the load line.) It is not to be inferred that the load line has to be tangent to the curve of the maximum power. The hyperbola of maximum power simply sets a limit on the operating points.

The power gain of a Class A stage is approximately equal to the squared value of large-signal current gain times the ratio of load resistance to input resistance, $beta^2 \times R_L/R_{in}$.

$$P_g = 10 \log \left(beta^2 \frac{R_L}{R_{in}}\right)$$

Efficiency. Two of the outstanding advantages of transistor power stages are the high power gain and the high efficiency available. This is particularly true of the common emitter circuit. Stage efficiencies of 40% to 47% can be realized; in fact, it would seem that 40% could be considered a design target low limit.

Biasing and Heat Dissipation. The circuit of Fig. 12a is well suited for many Class A power applications. The use of a germanium diode or thermistor-resistor network to stabilize emitter current is urged. Safety factors in operating voltages, current, power, and a low thermal resistance heat transfer circuit are the backbone of stable and reliable transistor power stages.

Distortion. Non-linear base-emitter input impedance and beta fall-off are two important factors contributing to distortion. The fall-off in current gain at higher collector currents is a device design problem and implies that transistors with the flattest current gain vs. collector current characteristics are to be preferred. Whenever possible, the transistor should be driven from a low-impedance source to reduce the stage distortion.

Care in the design of the output transformer also pays large dividends. As noted in Fig. 12a, an auto type transformer arrangement is used. Our experience indicates that the auto transformer connection gives very good results and is to be desired over a conventional transformer when circuit conditions permit its use. If a conventional transformer connection is used, it should be carefully designed to provide the highest practical incremental inductance with the D.C. magnetizing current and signal power level expected. After these conditions have been met, negative feedback can be used to provide an additional reduction in distortion. We would like to recommend to circuit designers that every effort be made to achieve the best possible performance, particularly in the area of distortion and bandwidth, before resorting to negative feedback.

Frequency Response. The low frequency response is almost entirely determined by the transformer characteristics and loading. The high frequency response is controlled by the transformer and the cut-off frequency of the transistor. The transistor cut-off frequency can readily be determined by means of a test circuit, as shown in Fig. 7c on power transistor measurements.

Class B Power Stages

Maximum Dissipation. Refer to notes on dissipation under "Class A Power Stages". In a Class B push-pull stage, maximum collector power dissipation occurs near maximum power output.

Power Supply Voltage. The power supply voltage V_{cc} should not exceed 1/2 the maximum collector voltage rating.

Signal Efficiency. Theoretical maximum signal efficiency equals 78%. Practical efficiencies of 60% to 65% should be realized in commercial designs.

Operating Point and Load Lines. See Fig. 12c. E_{ca} at I_a might be a typical zero signal operating point for a Class A power stage. E_{c1}, E_{c2} and I_7, I_0 are practical limits of collector voltage and current swings for each transistor in push-pull Class B. Note that the zero signal operating point shifts from E_{ca} to E_{cb} and that the transistors are operated with a small forward, base-emitter bias voltage to eliminate cross-over distortion. The highest value of collector load impedance possible, consistent with the power output required and voltage rating, should be used in order to reduce beta fall-off and attain high efficiency.

Biasing, Stabilization, and Heat Transfer. The base-emitter bias voltage should be a temperature-dependent constant voltage source. The base-emitter bias voltage should look like a low-impedance voltage source whose voltage changes inversely with temperature at the rate of approximately 2.0 to 2.5 millivolts per degree C. If this is not provided, rather severe cross-over distortion can occur at lower operating temperatures and/or low signal levels.

Diodes or thermistors are strongly recommended when any appreciable variation in temperature is encountered. Tight thermal coupling between transistors and the temperature-compensating diode is extremely important. Good practice would indicate that the temperature-compensating element should be mounted on the transistor heat-sink. This should provide a low thermal resistance between the transistor heat-sink and the temperature-compensating element. Such a mounting technique provides a temperature at the diode that is at least proportional to the junction temperature of the transistors.

The use of a resistor in the emitter leg of each transistor offers considerable improvement in preventing thermal run-away. We suggest that high values be used. The loss in signal gain is traded for a much improved stability factor.

Distortion. The loss in large-signal current gain at high collector currents and non-linear input resistance are also important causes of distortion in Class B power amplifiers.

With Class B, the input impedance is high at low signal levels, due primarily to the fact that the transistor is operated nearly at cut-off. Cross-over distortion is corrected by biasing each Class B transistor slightly above collector current cut-off. Fig. 12c shows that the operating point for true Class B operation would be at the E_{cb} point. While on the subject of bias and operating points, it should be mentioned that the base biasing resistor, thermistor, or diode should not be by-passed due to the rectification and charging of this capacitor, which could result in D.C. blocking.

Transistor matching is a cause of distortion. Ideally, the transistors should be matched for power gain, beta fall-off, and input characteristics.

Output and driver transformers can cause severe distortion. Our experience indicates that this is probably the most severe distortion problem. As an example, under sine wave signal conditions, we have what should resemble a 1/2 sinusoid current waveform in each collector circuit. These two 1/2 sine wave currents must be added in the transformer primary so that the resultant current looks like the original sine wave driving signal. This is no small problem in transformer design. However, we would like to refer to what we consider the very excellent results cited in Fig. 1 for a Class B output stage when the distortion, as measured at a power output beyond the stated capabilities of the transistor pair used, was actually below 3%. This performance was achieved without the necessity of resorting to bi-filar windings. However, grain-oriented lamination steel was used. Care in placement of primary and secondary windings, adequate inductance, and correct value of terminating impedance can be relied upon to give good bandwidth and low distortion. Then, the addition of negative feedback will yield a further improvement in the amplifier distortion figure. The R-C network shown as Z_C on Fig. 12a and 12b is helpful in reducing distortion, especially with resistive-inductive terminating impedances, such as loudspeaker loads.

Power supply regulation should be studied as a cause of distortion and thermal instability. A glance at Fig. 4 will show what can be done to improve the regulation of a conventional power supply. In a compact unit such as the Gray "Key-Noter", space is not available for such artifices as high-inductance "swinging" chokes for a choke-input filter, or for feedback regulated supplies. The dissipation problem at outputs below rated power can be surprisingly high with the regulation of curve (a) of Fig. 4.

Frequency Response. The limitations in frequency response with Class B push-pull stages are much the same as encountered with Class A power stages. For a given load impedance, the low frequency response is almost entirely controlled by the transformer characteristics. The high frequency response is controlled by both the transistors and the transformers and for primarily the same reasons as cited under Class A operation. The driver transformer, if used, should come in for a similar close scrutiny if broadband, high efficiency, and low distortion amplifier performance is to be achieved.

GENERAL DATA

I AMPLIFIERS

	4 STAGE, CLASS A	3 STAGE, CLASS B	4 STAGE, CLASS B
BAND WIDTH	100 – 6 KC MIN.	100 – 6 KC MIN.	100 – 6 KC MIN.
POWER OUTPUT	1.5 W	1.0 W	1.0 W
POWER GAIN	93 DB	86 DB	93 DB
DISTORTION, THD @ 1 W @ 400 CPS	5 %	4.5 %	3 %
SIGNAL / NOISE	40 DB MIN.	40 DB MIN.	40 DB MIN.
OUTPUT IMPEDANCE	4 Ω & 200 Ω	4 Ω & 200 Ω	4 Ω & 200 Ω
INPUT IMPEDANCE	2.4 KΩ	3 KΩ	2.4 KΩ
INPUT, VOLTAGE FOR RATED OUTPUT	300 u VOLTS	600 u VOLTS	220 u VOLTS

II TRANSDUCERS

MICROPHONE — DYNAMIC, 200 Ω
SENSITIVITY -- 1 M VOLT / u BAR

REPRODUCER — CERAMIC, 400 uuf

RECORDER — MAGNETIC, 4 Ω

RECEIVERS — A - 200 Ω @ P = 300 MW
B - 15 Ω @ P = 25 MW
C - 4 Ω @ P = 1 W

Fig. 1

TRANSISTOR AMPLIFIER
4 STAGE, CLASS A

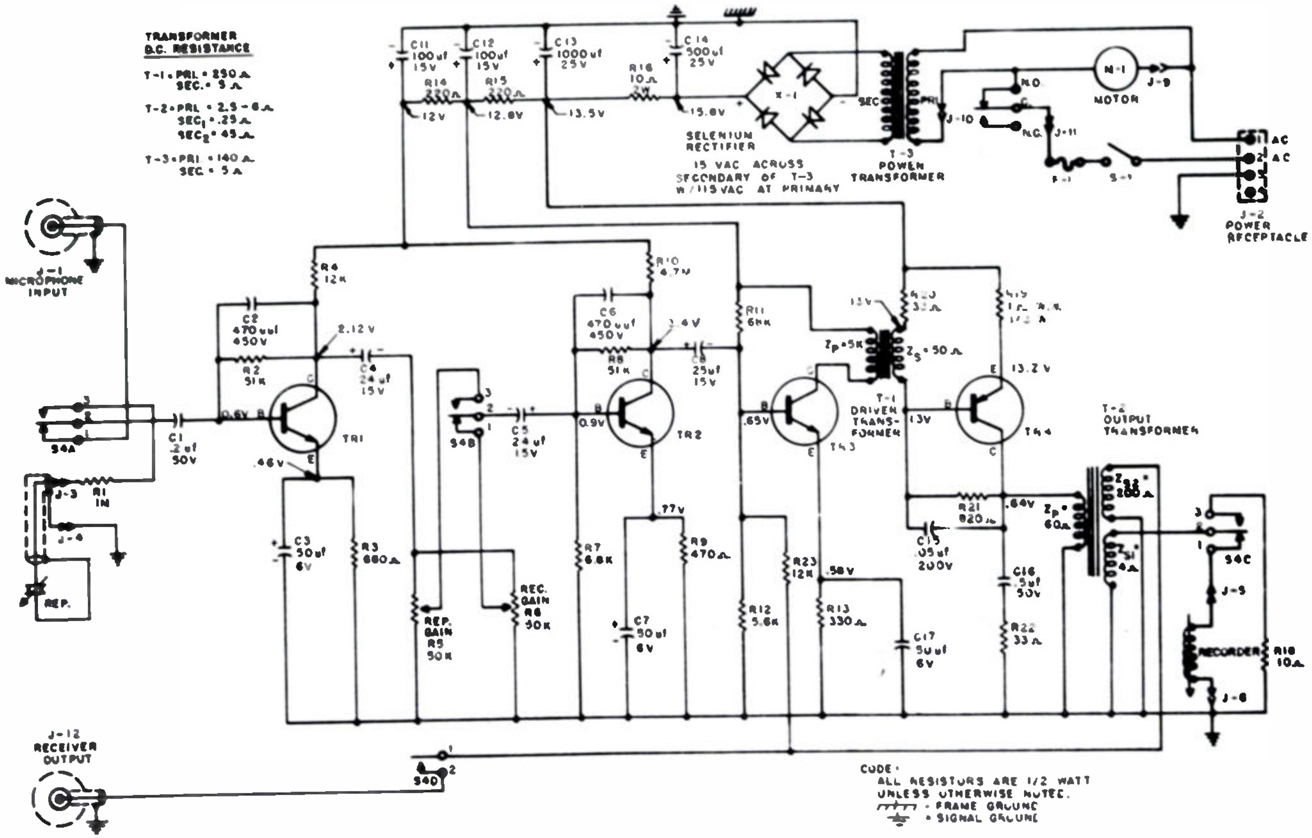


Fig. 2

TRANSISTOR AMPLIFIER
3 STAGE, CLASS B

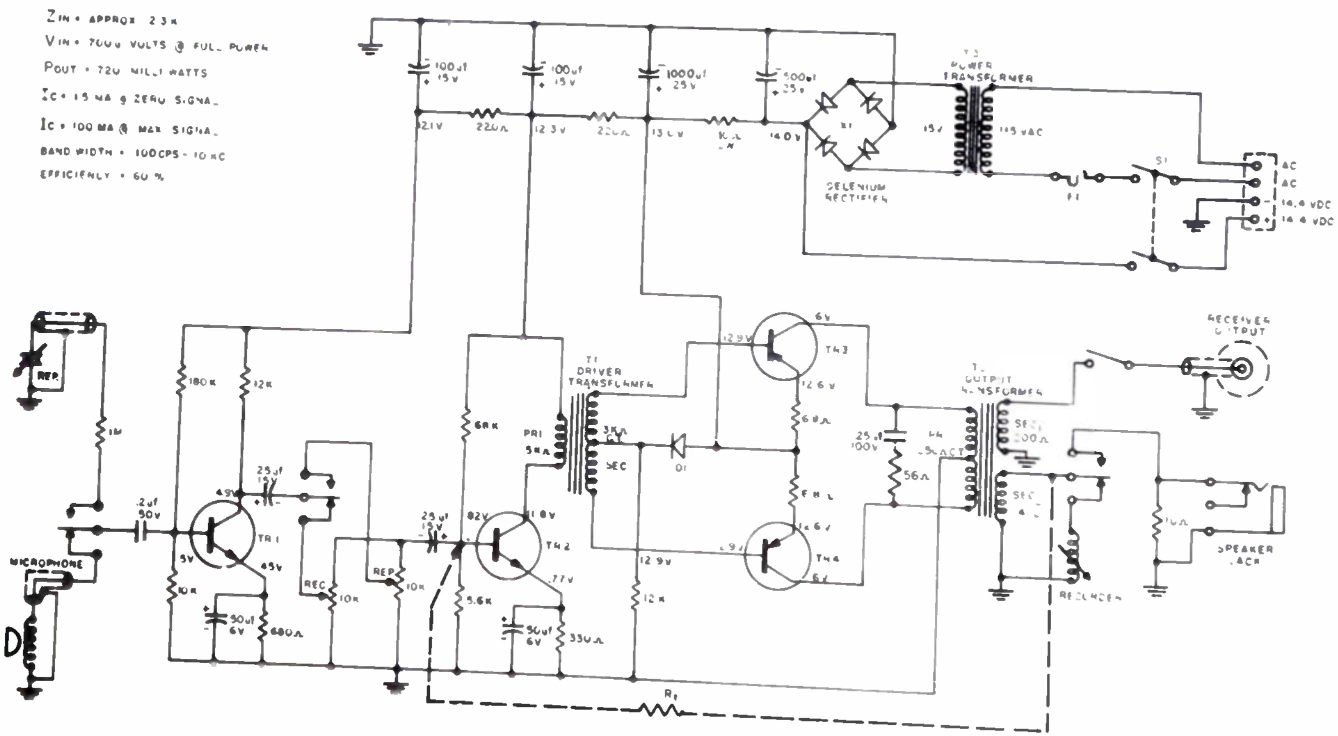


Fig. 3

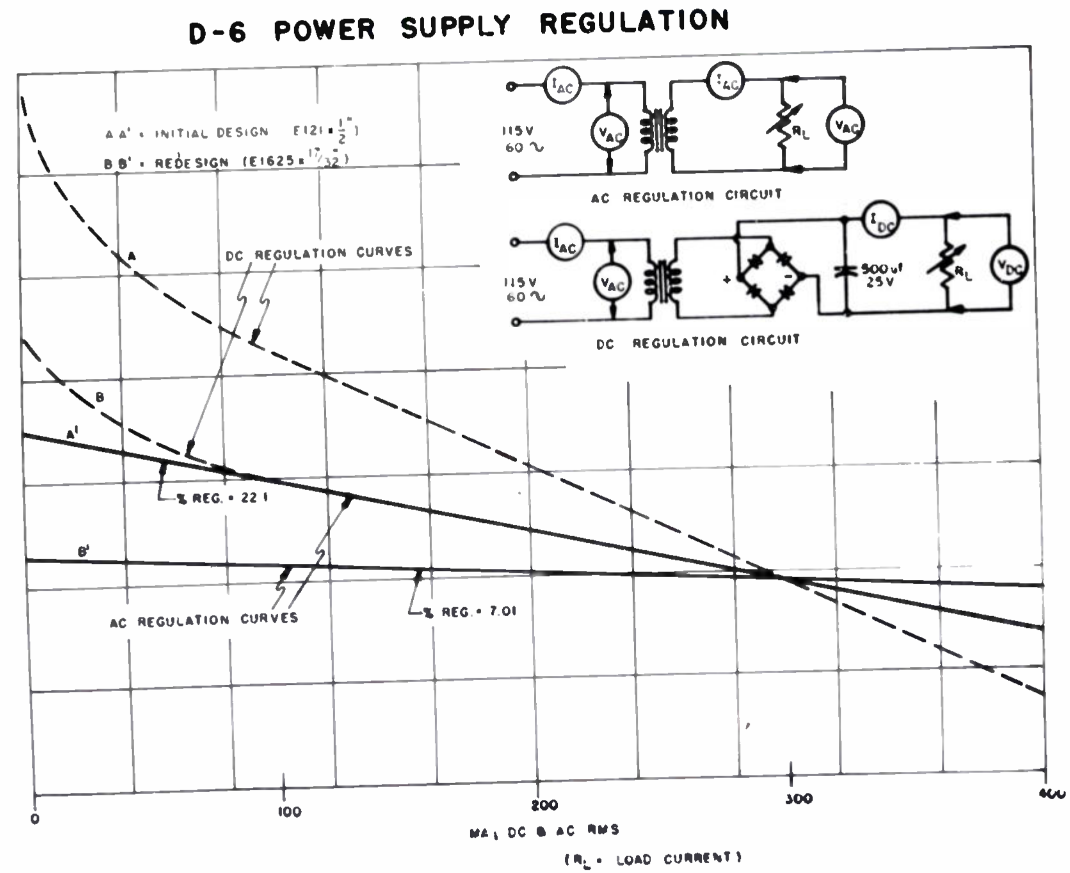

D-6 POWER SUPPLY REGULATION
A A' = INITIAL DESIGN (E121 = 1/2")
B B' = REDESIGN (E1625 = 17/32")
DC REGULATION CURVES
AC REGULATION CURVES
% REG. = 22.1
% REG. = 7.01
115 V
60 ~
AC REGULATION CIRCUIT
DC REGULATION CIRCUIT
500 uf
25 V
0
100
200
300
400
MA, DC & AC RMS
(R_L = LOAD CURRENT)

Fig. 4

TRANSISTOR TEST SET

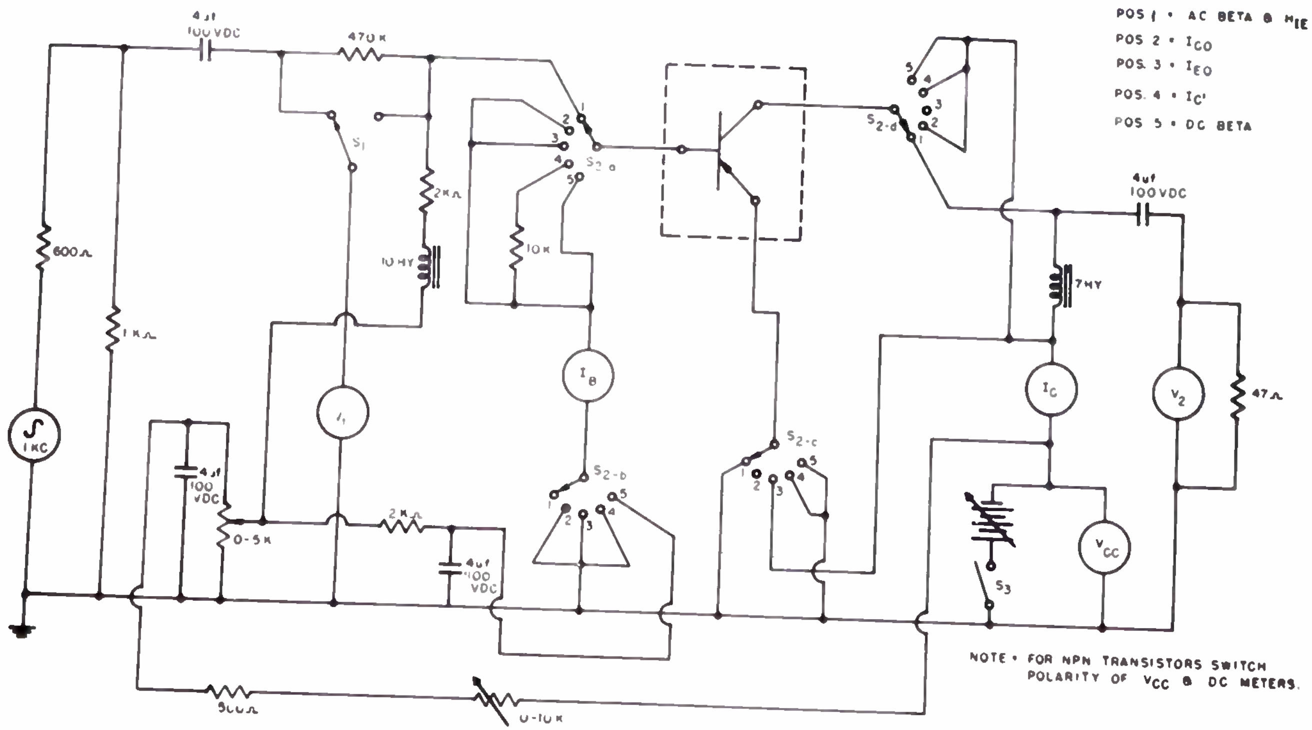


Fig. 5

AC BETA & H_{IE}

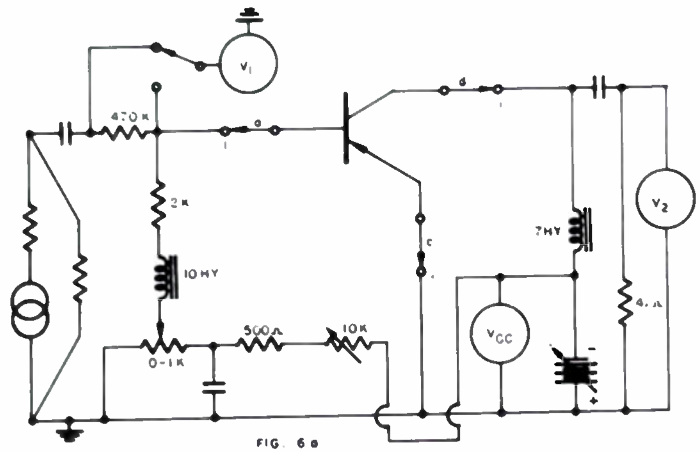


FIG. 6a

I_{CO} & I_{EO}

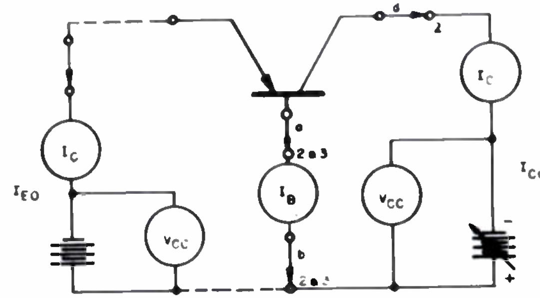


FIG. 6b

$I_{C'}$

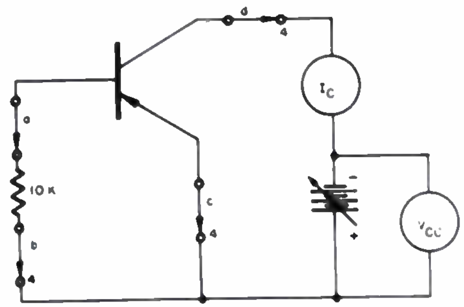


FIG 6c

DC BETA

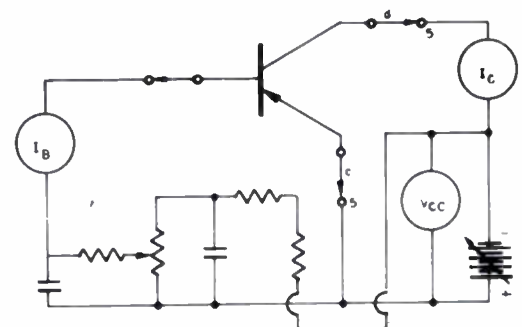
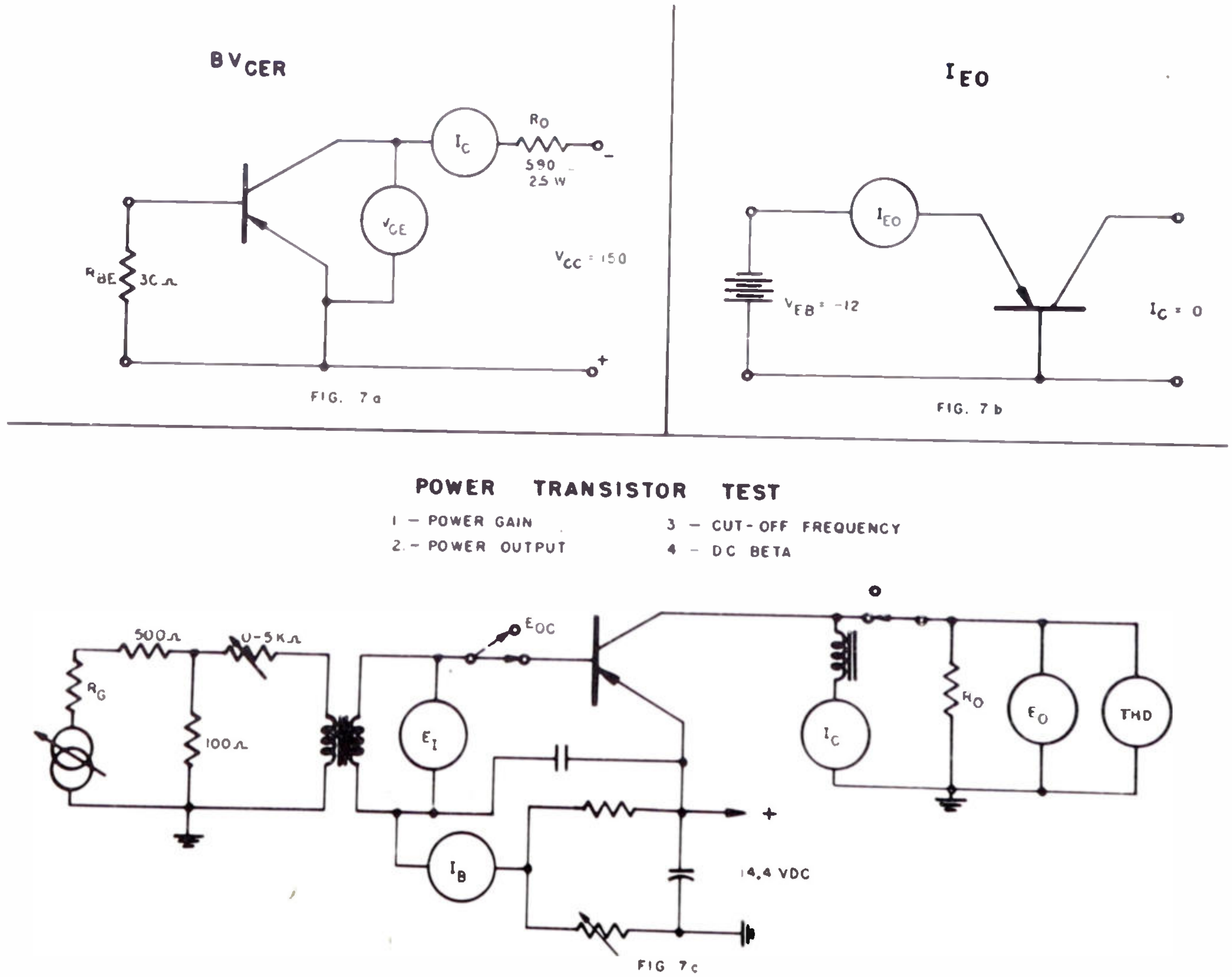


FIG 6d

Fig. 6



BVCER
I_C
R_O
590
25 W
V_{CE}
V_{CC} = 150
R_{BE} 30Ω
FIG. 7a
I_{EO}
I_{EO}
V_{EB} = -12
I_C = 0
FIG. 7b
POWER TRANSISTOR TEST
1 – POWER GAIN
2. – POWER OUTPUT
3 – CUT-OFF FREQUENCY
4 – DC BETA
500Ω
0-5KΩ
E_{OC}
R_G
100Ω
E_I
I_B
I_C
R_O
E_O
THD
+
14.4 VDC
FIG 7c

Fig. 7

TRANSISTOR NOISE VERSUS FREQUENCY

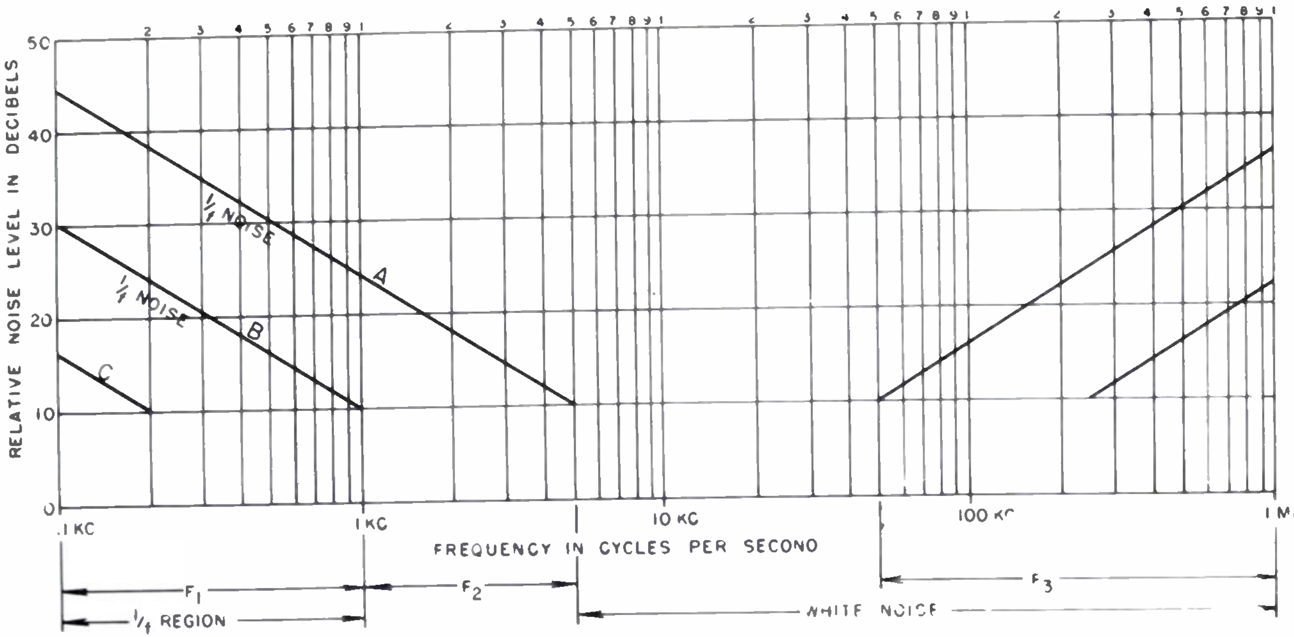


Fig. 8

TRANSISTOR NOISE
OCTAVE BAND ANALYSIS

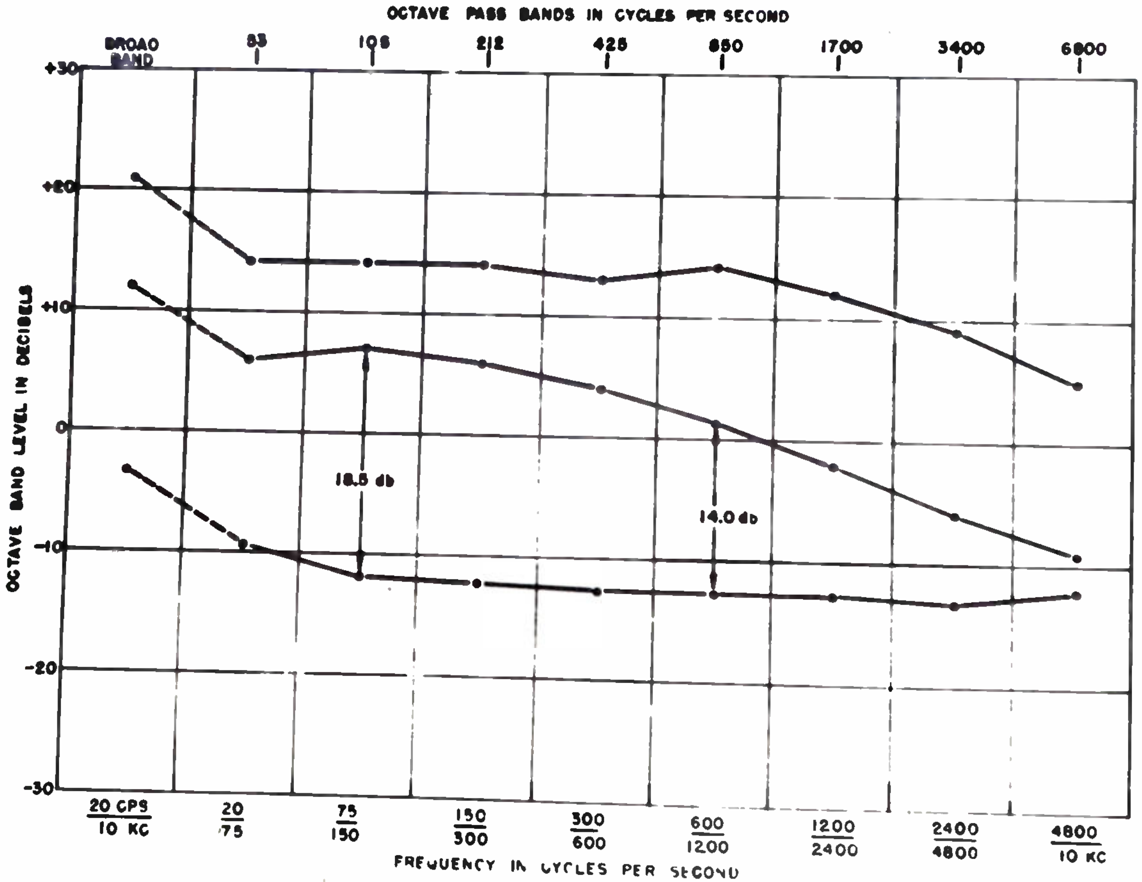


Fig. 9

TRANSISTOR NOISE TEST SET

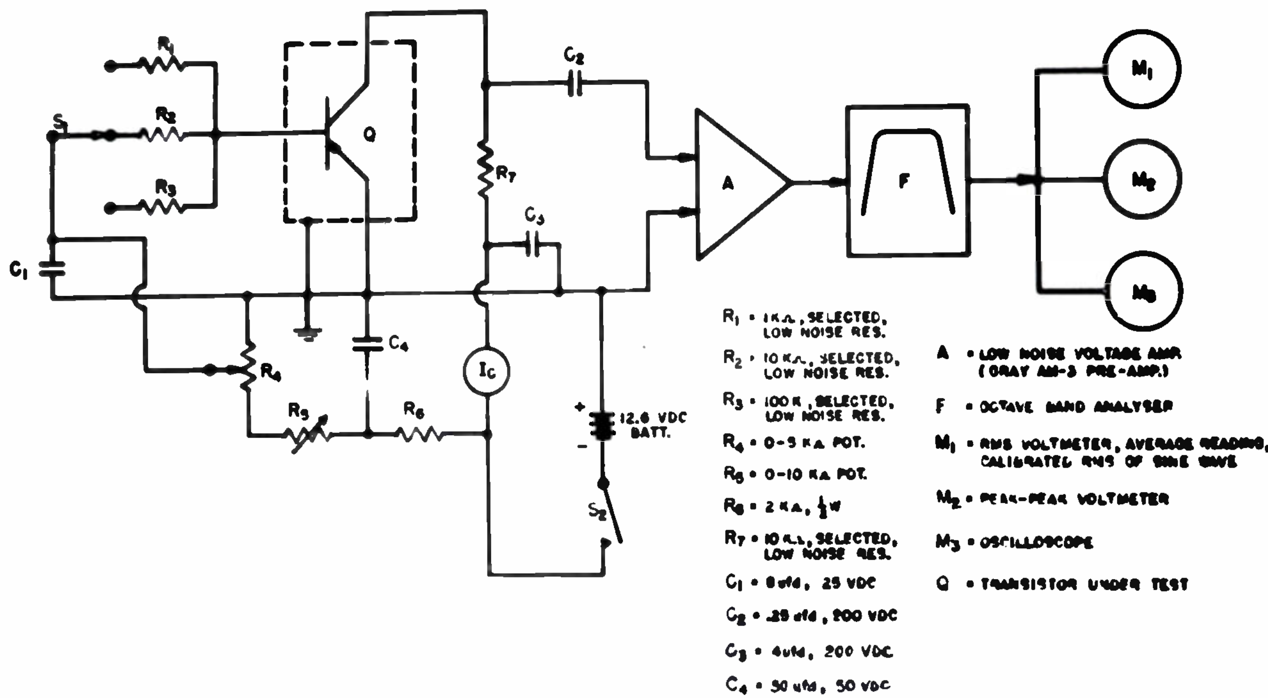


Fig. 10

BIAS & STABILIZATION
COMMON EMITTER

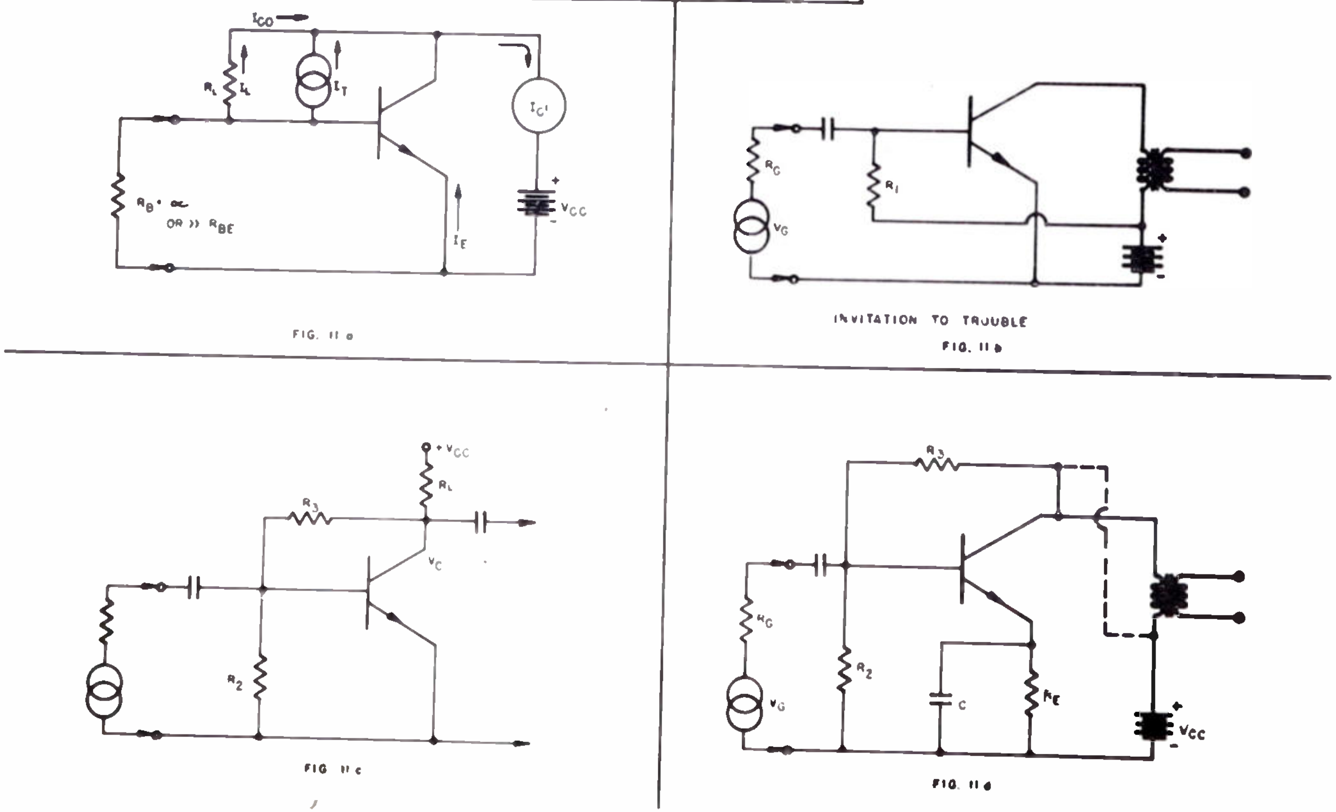
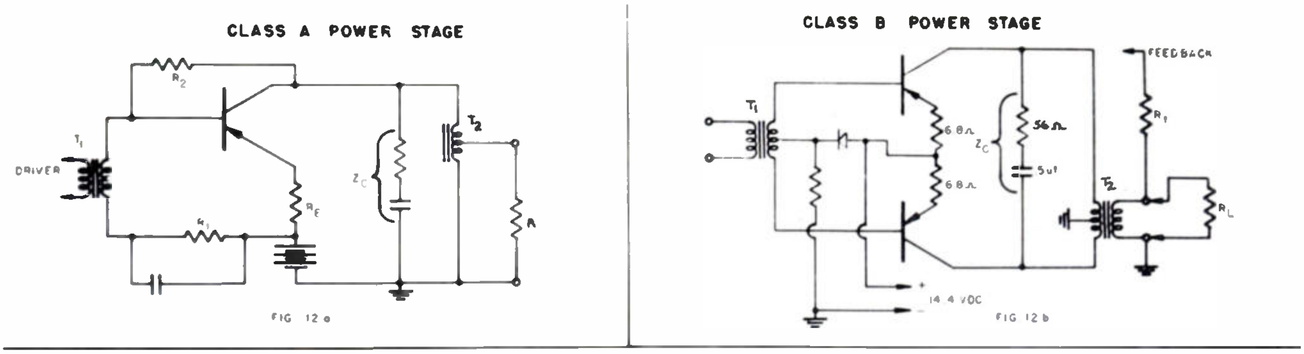
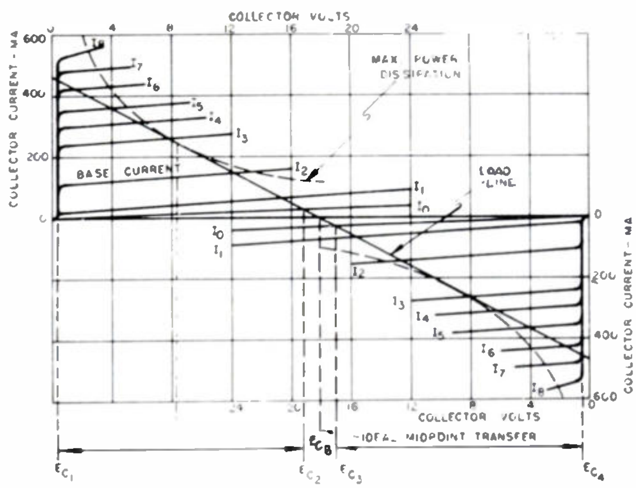


Fig. 11





COMPOSITE COLLECTOR CHARACTERISTICS FOR TWO JUNCTION TRANSISTORS IN CLASS "B" PUSHPULL

FIG 12 c

Fig. 12

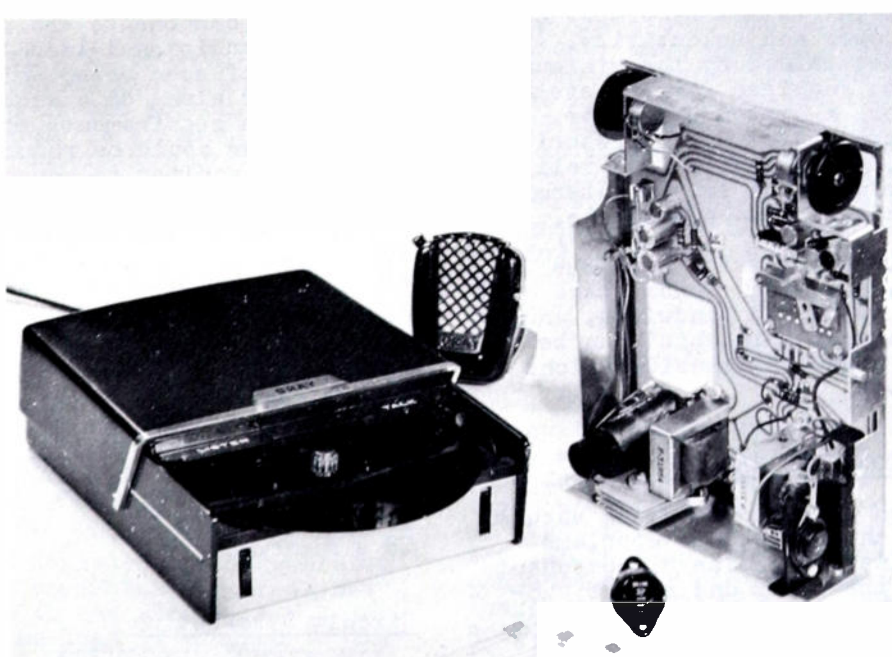


Fig. 13

SINGLE TUNED TRANSFORMERS FOR TRANSISTOR AMPLIFIERS

by

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SUMMARY

Single tuned circuits have required reexamination since advent of transistor applications, because the concept of power transfer, rather than voltage transfer, must be used. Match, under the constraint of a specified bandwidth, is defined. The effects of component losses upon the efficiency of the interstage are discussed.

INTRODUCTION

One basic role of electronics is the control by a small power of a large power for the operation of various types of transducers such as loudspeakers, cathode-ray tubes, and rocket fuel valves. The ratio of these powers is, of course, power gain. Unwanted signals and noise make the output power less faithfully controllable by the input. They are reduced to practical values by proper choice of the input power and selectivity. The bandwidth is determined by the minimum transition time required for the system and is therefore a system parameter. It is, then, the system which prescribes the bandwidth. The function and reliability required of the system determine the size and cost of the components.

Turning the discussion from the general to the specific, each stage must have first, a specified bandwidth, and, second, the largest gain which can be obtained with practical considerations of size and cost.

TRANSISTORS AND VACUUM TUBES

The design of circuits using vacuum tubes, in the main, presents problems where the engineer's choice of components is limited by the input and output capacitances present. Therefore, in order to produce a prescribed bandwidth it is necessary to add dissipative losses. Consider how few thousands of ohms are used to obtain the bandwidth in video amplifiers. Often the losses are part of other components. Remember how many coils have measured Q's practically identical with their operating Q's in the completed circuit.

Transistors, however, are resistance limited for most applications. Another viewpoint is that the intrinsic bandwidth of the input and output circuits is generally wider than is needed for the use intended. This is not usual when tubes are used. With transistors, it becomes necessary to add shunt capacitance to obtain the specified bandwidth. Often, too, components which were satisfactory for use with vacuum tube amplifiers are not acceptable with transistor amplifiers. Dissipation is needed with tubes to reach the desired bandwidth. With transistors, however, a bandwidth, which may already be excessive, is almost always increased if equivalent components are used. Furthermore, the additional loss involves a sacrifice of some power which could have been passed along to a succeeding stage or else the requirements for input signal power could be reduced. This discussion applies to any circuitry between resistance limited devices.

THE SINGLE TUNED CIRCUIT

The single tuned circuit serves as an excellent example of the effect of component dissipation. There is the source conductance of the transistor driving the interstage as well as the input conductance of the following transistor. There will, unfortunately, be some dissipation in any other component which is added. These losses are best kept to a minimum.

Circuit Description

Examine Slide 1 which shows at (A) a typical transformer coupled interstage. The next drawing (B) shows the equivalent circuit elements, while drawing (C) is the equivalent circuit which applies at resonance. G_s is the source conductance, G_L is the input conductance of the driven transistor, while G_u is the equivalent shunt conductance due to the losses in the transformer, all referred to the input terminals.

Efficiency

The gain or efficiency of the interstage is the ratio of the power delivered to the load to the power available from the source. The power available from the generator is P_A as shown in Slide 2, while the power in the load is P_L. The next line expresses this in terms of the generator current, i. Therefore, the power gain is A, as detailed at the bottom of the slide. We expect A to be a number less than one, therefore the word "loss" is applied here.

Bandwidth

The bandwidth of this stage, as a contributor to the bandwidth of the complete system, determines the operating Q of the interstage. This is called loaded Q and is the bandwidth condition as expressed in Slide 3.

Size and economics will generally prescribe a practical upper limit to the unloaded Q of the resonant tank, Q_u.

Analysis

By eliminating X_c between these equations and solving for G_U we obtain the relationship at the bottom of Slide 3. We thus find G_U is a function of four quantities determined by the system, its bandwidth (Q_L), its economics, (Q_u), and the choice of transistors, which determine the impedance level at the input (G_s, G_L). This can be substituted into the previously shown expression for gain and by simple algebraic processes we obtain the relationship shown in Slide 4.

Note that there are two independent factors controlling the efficiency of a single tuned interstage when bandwidth is fixed. These are A_L, the effect of loaded to unloaded Q, and A_m, which will be recognized as the efficiency of matching a load, G_L, connected to a generator having a source conductance G_s. A_m has a maximum equal to one when $G_L = G_s$; that is, when the load is matched to the source. Slide 5 shows how the loss increases as the match deteriorates.

Need for Matching

Input, as well as output impedances of transistors, often vary a great deal from unit to unit in one type number. To obtain least variation in match the geometric mean of the distributions should be chosen as the design value.

An amplifier for signals of uncertain strength, such as radio signals, will generally include Automatic Gain Control. The matched gain of transistors vary little with operating point but their input and output impedances change greatly. This property has been used to advantage by designing for matched conditions when the greatest gain is required and using the losses due to mismatch for Automatic Gain Control. Consequently, it is desirable to design the single-tuned interstage for nearly optimum match for impedance levels occurring without AGC.

Loss and Q Ratio

Slide 6 is a plot of A_L, the loss due to the loading ratio $\frac{Q_L}{Q_u}$ and covers the range most often used. A loading ratio of .1 produces a loss of approximately 1 db. This means that for an IF interstage having a bandwidth of 5 KC at 500 KC we have a Q_L of 100. If we desired this transformer to have only 1 db loss, we would have to specify an unloaded Q of 1000. While such a coil can be made, some exotic and expensive techniques must be used. Fortunately, in broadcast receivers, the bandwidth of the individual stages are much wider than 5 KC, and more loss has generally been acceptable as a design compromise. The example illustrates that a specific loss must be allowed by the engineer in designing an amplifier and we must keep in mind that this uniquely determines the unloaded Q.

REDUCTION TO PRACTICE

Our analysis complete, let us now reduce theory to practice. Measurements which enable the separation of the two loss factors are necessary. Making these measurements with a signal source of high, but known, resistance provides the information required when the added conductance is plotted against bandwidth. The circuit, Slide 7, includes capacitances to complete the substitution for the two transistors and produce the correct adjustment of a variable coil.

Rather than use a switch with its accompanying capacity, we have tack-soldered the G_s and g_L connections. The data obtained is plotted in Slide 8, where:

Point A. shows the bandwidth to the half-power points with the coil and generator alone, plotted with the known generator conductance, G_r.

Point B. shows the bandwidth and the additional source loading making a total of G_b.

Point C. is measured on a transformer, improperly tapped. The bandwidth is plotted with the g_L, but without G_s, connected.

The first two measurements are sufficient to define the straight line of this graph. Since the load is applied to the tank through the mutual coupling, the exact value of G_L, the shunt conductance (referred to the input terminals), is not directly known.

Interpretation of Data

Loss

The intercept on the G axis gives the shunt conductance of the tank, G_u, while the other intercept gives the unloaded bandwidth of the coil, B_u. Knowing that bandwidth is inversely proportional to Q, we can compare the required loaded and unloaded bandwidths and determine the loading efficiency from the relationship $A_L = (1 - \frac{B_u}{B_L})^2$. The error inherent to this approximation is usually smaller than the variation of G_u over the band.

Match

You have observed that the bandwidths added by the load and source are different in this illustration. This indicates an unmatched condition. Proper loading exists when half of the bandwidth deficiency from B_u to the specified bandwidth is contributed by the source and half by the load. That is $B_B - B_A$, and $B_C - B_A$, are each made equal to the quantity as shown in Slide 9. This can be accomplished by proper choice of the tapping point for the input and the turns ratio for the output.

Tank Impedance

Any impedance level of resonant tank can be used provided the tank inductance, which is measured across the terminals of the resonant tank, is high enough. At any terminal, bandwidth and conductance are directly proportional, with zero intercept, provided that we include the coil losses. This relationship is shown in Slide 10. Combining with the condition at match we solve the relationship for G_U. Substituting the expression for G_U in terms of Q_u and Reactance, we find the value of X_L to connect G_s at the top of the tank. Consequently, X_L must be larger than this value to have G_s tapped down on the tank. X_L is here expressed in an inequality, where G is the smaller of G_s and G_L to be certain that both can be tapped down on the tank.

CHOICE OF TERMINALS

Some properties of conjugate match have been used in the previous discussion: they are important enough to warrant discussion. Let us first consider a source and load, of different impedance, but connected by lossless elements, as in Slide 11. If we have adjusted the circuit so that the maximum power available from the source is dissipated in the load, conjugate match must exist. Consequently, conjugate match must also exist at any section through the network such as A, B, or C even though different values of R + jX and R - jX appear at each of these breaks. In addition, it follows that the deterioration of match as frequency is varied is independent of the terminals used. Therefore, the bandwidth measurements can be made from any terminal to ground. The inclusion of component losses adds some fixed conductance G_U as in the equivalent circuit analysed. It has been tacitly assumed that no other resonance occurs anywhere near the passband of interest. The output terminals have been used for measurements because they provide the least reaction by the vacuum tube voltmeter on the circuit being measured. A millivoltmeter is desirable to preclude measurements of magnetic materials at unreasonable flux densities.

EFFICIENCY OF SECONDARY

The discussion has completely neglected until now the effect of winding resistance in the secondary of this single-tuned interstage. For a properly designed transformer it is negligible, but the design criteria must be known.

Let us consider that we have an interstage designed as previously described. We have therefore matched (as shown in Slide 12) all of the secondary dissipation, r_1, to the source as though g_L absorbed all the secondary power, P_2. This means the power which should have gone into the load is actually divided between the load and the AC series resistance of the winding, r_2. A series resistor will absorb power when secondary current flows. This is the effect with which we are now concerned. As shown, A_2 is the efficiency of power transfer to the load and is equal to $\frac{1}{1 + r_2 g_L}$. The expression for efficiency of the whole interstage then becomes $A = A_L A_m A_2$.

A practical consideration is that r_2 is to be kept low. This means that a larger conductor may be desirable. It is even more important to keep the coupling high between the resonant tank and the output winding so that fewer turns are needed in the latter. Measurement of the value of r_2 is not straightforward. A slightly optimistic value to use is the dc resistance. When the inductance has sufficient Q to be able to measure it on the Q-meter, (a Q greater than 10) the secondary loss is negligible for most loads. The winding, however, need not be good enough to be measurable on a Q-meter.

Measurements and Efficiency

One typical sample had a secondary unloaded Q of one, which meant only .1 db loss. This Q value was determined as follows. A slightly pessimistic determination of r_2 for the secondary winding of a 262 KC IF transformer was obtained by plotting its measured values (obtained from conventional bridges) at 60 cycles and at 1 KC, and above 500 KC using an RF bridge and computing the equivalent series resistor. From the curve joining these points the value of r_2 was taken, and the Q of one computed.

MEASUREMENT METHODS COMPARED

Before using this design and measurement method on transformers we checked this new method by comparing it with data yielded by our old method. Slide 13 shows this method of measurement. Its results confirmed the new approach; the discrepancy was less than 1/2 db.

A brief comparison is relevant here. The method shown in Slide 13, has the disadvantage of using several different ranges of the voltmeter. This results in cumulative errors. The only calibrations required for the suggested new method are the full level and the .707 points, a calibrated constant resistor, the specified source, and the terminating resistance. The various power points for the bandwidth measurements can be obtained with one meter reading by removing a 3 db attenuator section. These resistive elements can easily be measured and maintained as accurately as desired.

Another method was considered. The loadings referred to the tank terminals can be computed when the self and mutual-inductive elements are known, together with the Q's of each of the windings. This involves quite a few measurements on several ranges of several different pieces of test equipment. The errors could be cumulative, and consequently, the method was not used.

SUMMARY

The main points which have been covered in this paper are that a source and its load should be matched as though losses did not occur in the windings of a single tuned interstage. The loss of the transformer is then a minimum and is dependent only on the ratio of loaded to unloaded Q. Measurement methods were described which separate the various loss factors.

We wish to express our appreciation to Robert C. Moore, Harold B. Collins, Jr. and John J. O'Grady for their cooperation at various phases of the development and confirmation of this method.

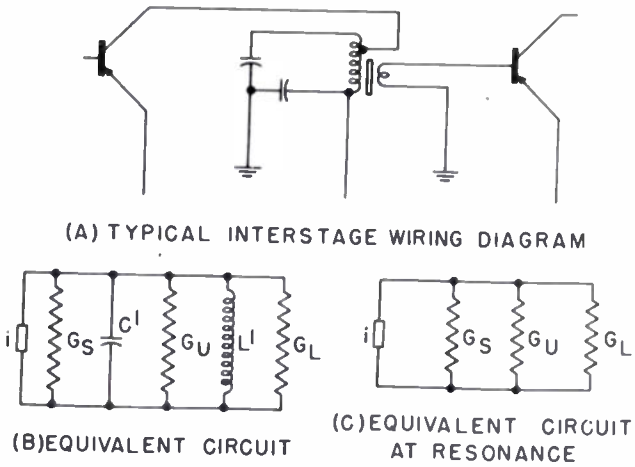


Fig. 1

AVAILABLE POWER $\quad P_A = \dfrac{i^2}{4 G_S}$

POWER IN LOAD $\quad P_L = \dfrac{i_L^2}{G_L}$

$$= \left(i \, \frac{G_L}{G_S + G_U + G_L} \right)^2 \frac{1}{G_L}$$

GAIN $\quad A = \dfrac{P_L}{P_A} = \dfrac{4 G_L G_S}{(G_S + G_U + G_L)^2}$

Fig. 2

BANDWIDTH CONDITION

$$Q_L = \frac{1}{(G_S + G_U + G_L) X_C}$$

DESIGN CONSTRAINT

$$Q_U = \frac{1}{G_U X_C}$$

$$G_U = \frac{(G_S + G_L)}{\left[\dfrac{Q_U}{Q_L} - 1 \right]}$$

Fig. 3

$$A = \left(1 - \frac{Q_L}{Q_U} \right)^2 \frac{4 G_S G_L}{\left[G_S + G_L \right]^2}$$

$$= A_L \cdot A_M$$

WHERE $\quad A_L = \left(1 - \dfrac{Q_L}{Q_U} \right)^2 \cong \left(1 - \dfrac{B_U}{B_L} \right)^2$

AND $\quad A_M = \dfrac{4 G_S G_L}{\left[G_S + G_L \right]^2}$

Fig. 4

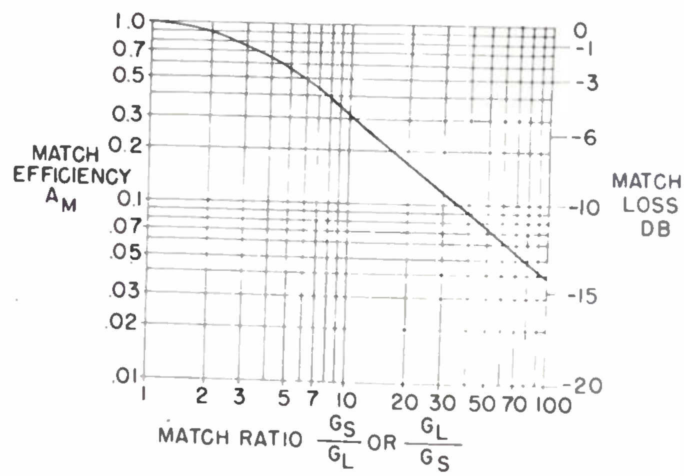


Fig. 5

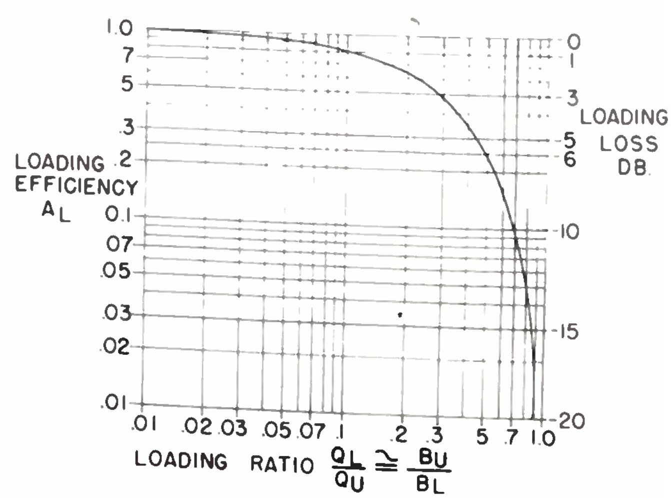


Fig. 6

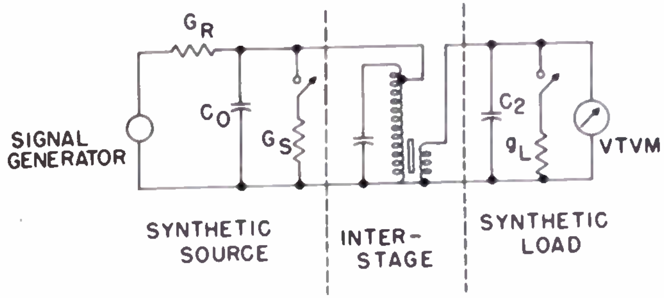


Fig. 7

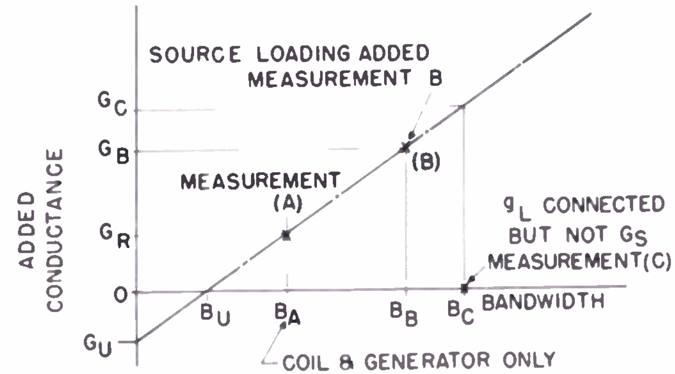


Fig. 8

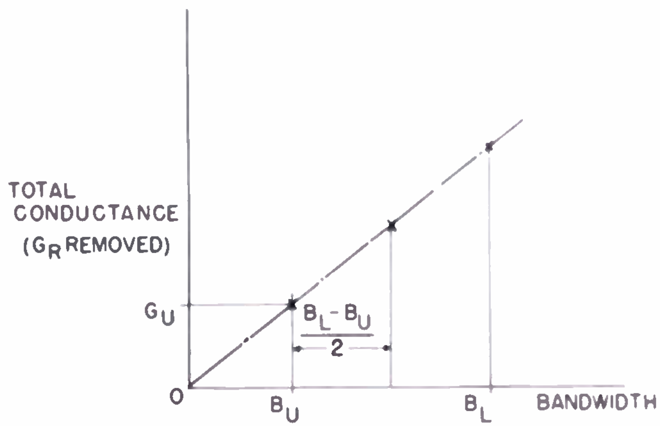
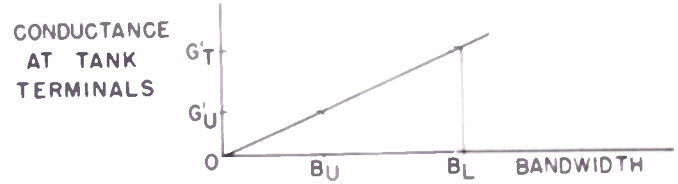


Fig. 9



$$\frac{G'_U}{B_U} = \frac{G'_T}{B_T}$$

AT MATCH $G'_T = G'_U + 2G'_S$

$$G'_U = \frac{B_U}{B_T - B_U} 2G'_S$$

BUT $G'_U = \frac{1}{Q_U X_L}$

$$\therefore X_L = \frac{B_T - B_U}{B_U 2G'_S Q_U}$$

AND $X_L > \frac{Q_U - Q_L}{2GQ_UQ_L}$

WHERE $G < G_S, G_L$

Fig. 10

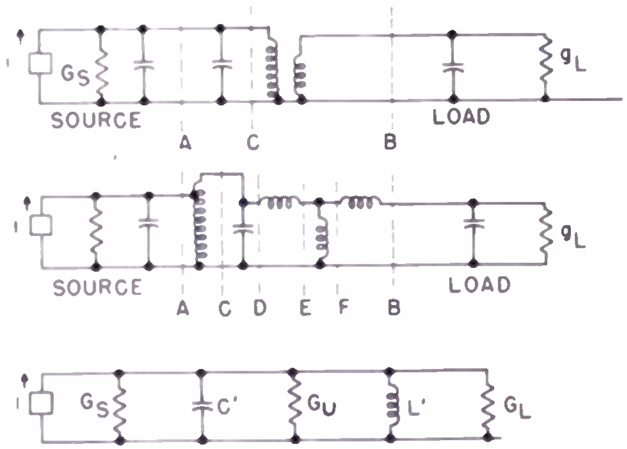


Fig. 11

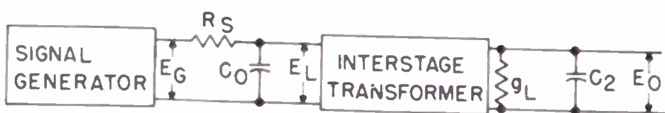
$$P_2 = i_2^2 r' \qquad \text{OR} \qquad i_2^2 = \frac{P_2}{r'}$$

$$\text{BUT} \quad P_L = i_2^2 r_2 \quad = \quad P_2 \frac{r_L}{r'}$$

$$\text{EFFICIENCY, } A_2 = \frac{P_L}{P_2} = \frac{r_L}{r'} = \frac{r_L}{r_2 + r_L}$$

$$= \frac{1}{1 + r_2 g_L}$$

Fig. 12



SIGNAL GENERATOR IMPEDANCE MUCH LESS THAN R_S

POWER GAIN

$$P_i = \frac{E_i^2}{R_S}\left(\frac{E_G}{E_i} - 1\right)$$

$$P_O = E_O^2 g_L$$

$$A = \frac{P_O}{P_i} = \left(\frac{E_O}{E_i}\right) R_S g_L \left(\frac{1}{\frac{E_O}{E_i} - 1}\right)$$

BANDWIDTH

TO 1/2 POWER POINTS OF E_O (E_G CONSTANT)

Fig. 13

DESIGN CONSIDERATIONS FOR TRANSISTORIZED AUTOMOBILE RECEIVERS

By

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The relatively lower maximum available gain of conventional alloy transistors in the rf and if stages of a transistorized automobile receiver requires the use of more stages than commonly considered economical by the automobile industry. The use of the drift transistor, however, which has a high maximum available gain and a low feedback capacitance permits a design having the same number of stages used in present automobile receivers of the "hybrid" variety. Because drift transistors have an inherently low feedback capacitance, they can be used in unneutralized rf or if circuits with only a small sacrifice in gain compared to that of neutralized circuits. This paper analyzes the operation of drift transistors in the rf and if stages of a typical automobile receiver. In addition, audio transistor considerations are discussed in order to present some of the basic over-all considerations involved in the design of a complete auto receiver. Economic as well as technical considerations will be included. The transistors described are RCA developmental types designed specifically for the application.

Internal Feedback In Drift Transistors

Fig. 1 shows the hybrid-pi equivalent circuit for a drift transistor in the common emitter configuration. Two of the most important parameters in this circuit are the depletion-layer capacitance ($C_{b'c}$) and resistance ($r_{b'c}$) of the reverse-biased collector-to-base diode. These parameters provide the internal feedback path from the output to the input circuit. If the output circuit is terminated in a purely resistive load, the output voltage is approximately 180 degrees out of phase with the input voltage. Consequently, the feedback is degenerative. When the output circuit is terminated in a resonant circuit, as in most rf and if receiver circuits, a small change in frequency causes a considerable phase shift in the output. As a result, at or below the resonant frequency, the feedback is degenerative while above resonance, the feedback is regenerative. If the maximum available gain is considered equal to $g_m^2 R_{in} R_{out}/4$, the magnitude of the feedback capacitance is a primary consideration in determining the maximum stable gain that can be obtained from an unneutralized amplifier and, to some degree, the maximum stable gain that can be obtained under neutralized conditions. The value of $C_{b'c}$ in drift transistors varies between 1 and 2 micromicrofarads for collector-to-base voltages between -32 and -5 volts and is practically independent of emitter current. Theoretically, the ideal value of $C_{b'c}$ is zero since for a given device, as $C_{b'c}$ approaches zero the useful gain approaches the maximum available gain. Practically, feedback will always exist because of stray capacitance. Because of the magnitude of $r_{b'c}$ the energy fed back through this resistance can be neglected. Fig. 2 shows the maximum available and maximum usable gain characteristics for the drift transistor over the broadcast band. The maximum usable gain is calculated to permit maximum stability[1] and interchangeability and includes such considerations as the variation in internal feedback capacitance ($C_{b'c}$) from unit to unit, the tolerance of the neutralizing feedback capacitor (in neutralized stages), and a safety factor to insure that the selectivity curve will indicate no tendency toward regeneration. For maximum available gain calculations and in unilateralized amplifiers, the parameters of interest are the unilateralized parallel input and output resistances as shown in Fig. 3. The variation in input resistance is approximately 1.7 to 1.0 over the band, decreasing with increasing frequency. The variation in output resistance is approximately 4 to 1 over the band and is the primary reason for the decrease in gain with frequency as shown on Fig. 2.

In an unneutralized amplifier, the resistance curves are modified by the internal feedback. Fig. 4 shows the unneutralized input and output resistance over the broadcast band for a given set of conditions. The input resistance is taken with the output terminated in 2000 ohms and the output resistance is taken with the input terminated in 600 ohms. The variation of these resistances with frequency is essentially the same as for the neutralized case.

RF Amplifier Stage

Fig. 5 shows the circuit diagram for an inductance-tuned rf amplifier utilizing tuned input and output. The input circuit is tuned with a capacitance of 120 micromicrofarads. This consists of a 60 to 80 micromicrofarad capacitance whip antenna (23 micromicrofarad antenna capacitance and lead in capacitance) plus a trimmer capacitor. The unloaded Q of the input circuit varies from 50 to 60 over the broadcast band and the loaded Q has been selected at 0.7 this value. This loaded Q permits a slight increase in image and if rejection with a loss of only 0.8 db in power transfer as compared to the matched condition. This is only an apparent loss since the losses required for the stability of the stage (see Fig. 2) may be made up at either the input or the output of the circuit (or at both). The maximum available gain of the stage under the conditions shown is 46 db and the maximum usable gain is 26 db (unneutralized). An increase of approximately 4 db may be obtained if the stage is operated neutralized. The unloaded Q of the rf transformer varies from 50 to 60 over the band and the loaded Q is almost half this value.

In the circuit shown in Fig. 5, an image-rejection ratio greater than 60 db can easily be obtained at low and mid-band frequencies with an if of either 262.5 or 455 kilocycles by a compromise between image-rejection and insertion loss[2]. If higher image-rejection ratios are required, as is usually the case, the pi image-rejection filter shown in the insert may be used in place of the rf tuned circuit; or in addition to the antenna tuned circuit and the output operated broadband. Another possibility is the use of the pi image-rejection filter at the input and the rf tuned circuit at the output. The use of one of these techniques in preference to the other is dictated by the intermodulation distortion characteristics and to some degree the signal-to-noise characteristics required by the receiver.

AGC voltage is applied to the base of the transistor in Fig. 5 to control the emitter current, and as a result, the gain. The stage is operated at an emitter current of 0.5 milliamperes to reduce the AGC power required to cut off the stage. In addition, the change in input and output resistance of the device with emitter current is greater when the initial operating point is 0.5 milliampere rather than, for example, 1.0 milliampere. (These characteristics will be shown in a later figure.) The result is a more effective AGC system. Although this type of operation is advantageous when an AGC amplifier is not being used, it causes a 3 db loss in gain as compared to the 1.0 milliampere operating condition. This loss results from the rise in input and output resistance with decreasing emitter current for approximately the same maximum available gain. For the same feedback capacity, this requires additional losses for stability and as a result, a lower usable gain. Increases in the input and output resistance of the transistor caused by the application of AGC produces no instability because the stage is kept stable by the low terminating impedances of the antenna and rf circuit.

Converter And Mixer-Oscillator Stages

Fig. 6A shows the circuit diagram for an inductance-tuned converter. The oscillator section of the converter is connected in the common-base configuration and the mixer section in the common-emitter configuration.[3] Feedback from the collector to the emitter for the oscillator section, is obtained by the tickler winding and the voltage divider C_1 and C_2. The conversion gain of the stage into a 262.5 kilocycle if is 38 db at 1.0 megacycle for a collector-to-emitter voltage of -12 volts and an emitter current of 0.6 milliampere. (The input and output to the converter in Fig. 6A is matched as closely as possible consistent with practical transformer design.)

Presently, the use of a converter circuit has one disadvantage. Under strong signal conditions, the large signal at the base of the converter blocks the oscillator. When the set is operating, the AGC network helps to control this problem. However, when the set is turned on under strong signal conditions the oscillator is blocked before the AGC network has time to react. Methods of attenuating strong signals (not totally dependent on AGC) before they reach the converter are being investigated and the results look promising.

Oscillator-mixer circuits, such as the one shown in Fig. 6B, have the advantage that the oscillator is relatively independent of the mixer and as a result, the blocking problem discussed previously is not encountered. In addition, the mixer portion may be subjected to AGC. These advantages are obtained at the cost of an additional transistor. The mixer circuit shown provides 40 db of conversion gain into a 262.5 kilocycle if. Dc operating conditions are the same as those in the converter stage. Optimum conversion gain in the converter and the mixer is obtained with between 100 and 150 millivolts of injection voltage.

IF Amplifier Stage

Fig. 7 shows the circuit diagram for a two stage 262.5 kilocycle if amplifier. The first stage incorporates AGC and is operated at an emitter current of 0.5 milliampere for the same reasons discussed previously in connection with the rf amplifier. This stage provides an unneutralized stable gain of 33 db. The second stage is operated at an emitter current of 1 milliampere and provides 36 db of gain. A 5 db increase in gain per stage can be obtained under neutralized conditions. At 455 kilocycles, the usable gain under the above conditions is 3 db less than that at 262.5 kilocycles, primarily because of the decreasing output resistance of the device at higher frequencies.

For good adjacent channel attenuation, the circuit employs two double-tuned transformers and a single-tuned transformer. The first double-tuned transformer (T_1) and the single-tuned transformer (T_2) are employed in the conventional manner. The primary of the output double-tuned transformer (T_3) is used for AGC and the tapped secondary for audio output. As a result, the AGC bandwidth is wider than the audio bandwidth as shown in Fig. 3. The reason for this is as follows: Under strong signal conditions, the increased signal at the converter or first if base causes an effective increase in modulation which results in an increased audio output. This is particularly important when tuning onto or off a strong signal. If the AGC bandwidth is made wider than the audio bandwidth, the audio will follow the AGC as the receiver is tuned onto the station until the station is peaked.

The load presented to the second if transistor of the amplifier shown in Fig. 7 should be chosen with regard for large signal considerations so that the second stage can supply sufficient AGC power to cut off the rf and first if amplifier and at the same time, be capable of over driving the audio system. The load impedance chosen for this stage was approximately ten thousand ohms.

In order to provide the required losses for stability in these two stages, the load impedances must be low. As a result, if transformers can be designed which have relatively low unloaded Qs and still obtain the desired bandpass characteristics. The use of single-strand wire instead of Litz wire, and powdered iron instead of ferrite, results in a considerable saving.

It is desirable to split the stability requirement of the first if stage so that both the input and output of the transistor are heavily loaded. The bandpass characteristic, as indicated by the unloaded and loaded Qs shown at the top of Fig. 7, depends almost entirely on the transformer characteristic. Under AGC conditions, the first if stage produces little sharpening of the bandpass characteristic because of increases in the input and output impedances of the transistor with decreasing emitter current. Fig. 9 shows the variation of these impedances with emitter current.

Driver Stage

Fig. 10 shows the circuit diagram of an audio driver stage.[4] When the wiper arm of the 0.5 megohm volume control potentiometer is in the extreme left position, maximum gain is obtained. As the wiper arm is moved to the right, the gain is reduced as a result of the series resistance. At a point on the potentiometer, the resistance from the wiper arm to the collector starts to enter as feedback. One advantage of the system is a dynamic range of control of 120 db. In addition, the system provides a dc to ac load ratio for the detector which approaches unity as the wiper arm is moved to the right. This results in a good distortion versus per cent modulation characteristic for strong and average signal conditions.

The addition of a tone control to a single stage driver with feedback provides a serious problem since the tone control should not be included in the feedback loop. One possible layout is shown in Fig. 10. This arrangement has the disadvantage, however, that it is ineffective at the maximum volume position. The maximum gain from the volume-control wiper to the primary of the driver transformer is 41 db (minimum feedback).

Output Stages

Fig. 11 shows the circuit diagram for two possible output circuits: a single-ended circuit capable of delivering 4 watts at less than 10 per cent distortion, and a push-pull circuit capable of delivering 8 watts at less than 10 per cent distortion. The power gain of both stages is 32 db from the driver-transformer secondary to the output transformer primary.

Receiver Circuit

Fig. 12 shows the circuit diagram for a developmental six-transistor automobile receiver incorporating the transistors and circuits described in the preceding sections. The sensitivity, image-rejection ratio and if rejection ratio as a function of frequency are shown on Fig. 13. Using the dummy antenna shown in the insert, the sensitivity varies from 0.9 microvolts at the low end of the band to 1.7 microvolts at the high end of the band for one watt out. Utilizing tuned input and tuned output in the rf stage, the receiver provides an image-rejection ratio which varies from 63 db at the low end of the band to 47 db at the high end of the band and an if rejection ratio which varies from 55 db at the low end of the band to 85 db at the high end of the band.

Fig. 14 shows the AGC and noise characteristic of the receiver. The set has a 66 db AGC Figure of Merit using a 5000 microvolt reference and a 20 db signal-to-noise ratio for approximately 7 microvolts at the antenna. With the AGC disconnected oscillator blocking occurred at 10,000 microvolts at the antenna. With the AGC connected and if the receiver is tuned onto a station, oscillator blocking occurs at 150,000 microvolts. With the AGC connected and the receiver turned on in the presence of a strong signal, oscillator blocking occurred at approximately 30,000 microvolts.

Fig. 15 shows the distortion characteristic of the receiver as a function of per cent modulation. For 100 milliwatts out, the distortion varies from one per cent at 30 per cent modulation to 2.1 per cent at 80 per cent distortion. For one watt out, the distortion varied from 3.2 per cent at 30 per cent to 5.5 per cent at 80 per cent modulation.

At the time of this writing, a pi image-rejection filter has been incorporated into the rf amplifier of the receiver and methods for eliminating the oscillator blocking are being investigated.

Conclusion

This paper has described some of the design considerations involved when drift transistors are used in the rf and if circuits of an automobile receiver. These considerations, coupled with a compatible audio system, result in a highly commercially acceptable automobile radio receiver. The use of these transistors should enable such receivers to be built in large volume with adequate economic justification.

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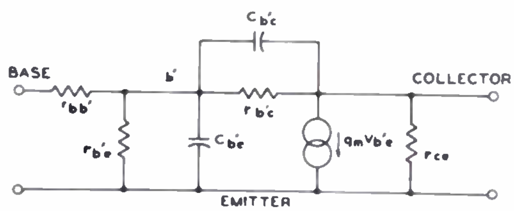


TYPICAL CHARACTERISTICS

V_{ce} = 12 volts I_E = 0.5 ma

$r_{bb'}$ = 35 ohms $C_{b'c}$ = 1.55 $\mu\mu f$

$r_{b'e}$ = 3000 ohms $C_{b'e}$ = 155 $\mu\mu f$

$r_{c'e}$ = ∞* g_m = 18900 μmhos

$r_{b'c}$ = ∞*

* For all practical considerations.

Fig. 1

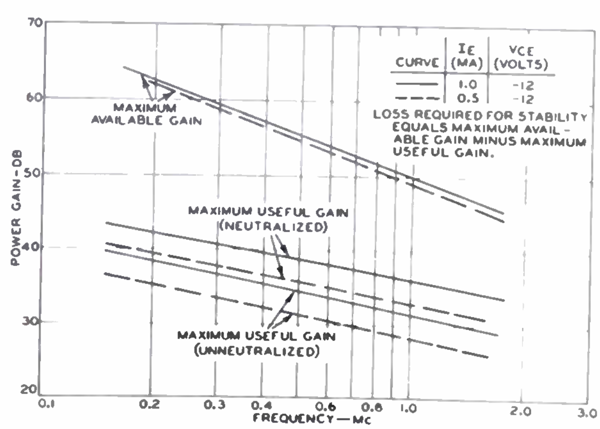


Fig. 2

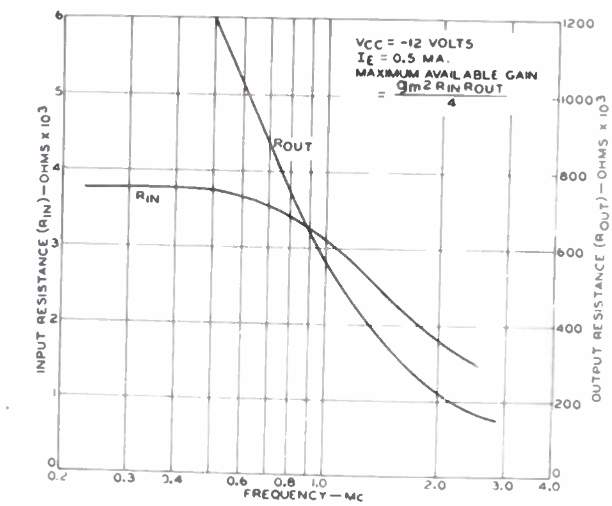


Fig. 3

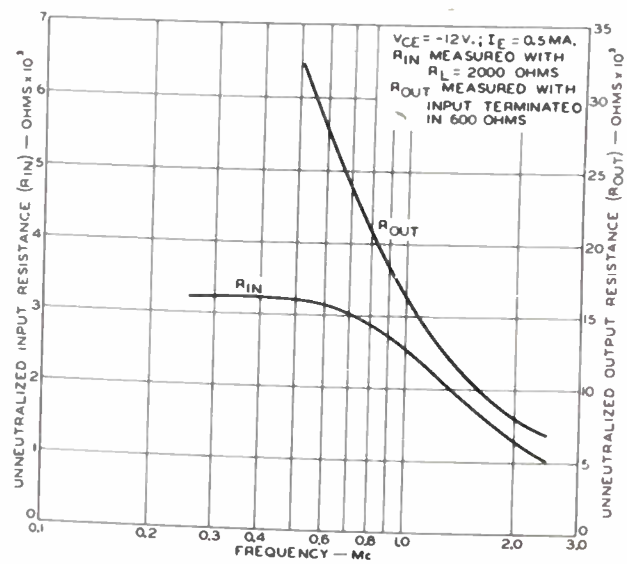


Fig. 4

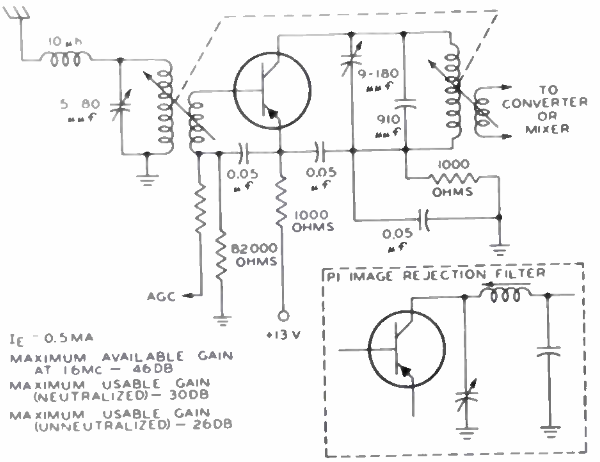


Fig. 5

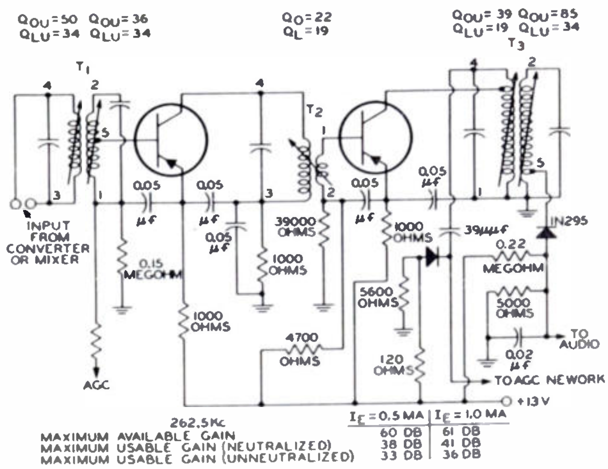
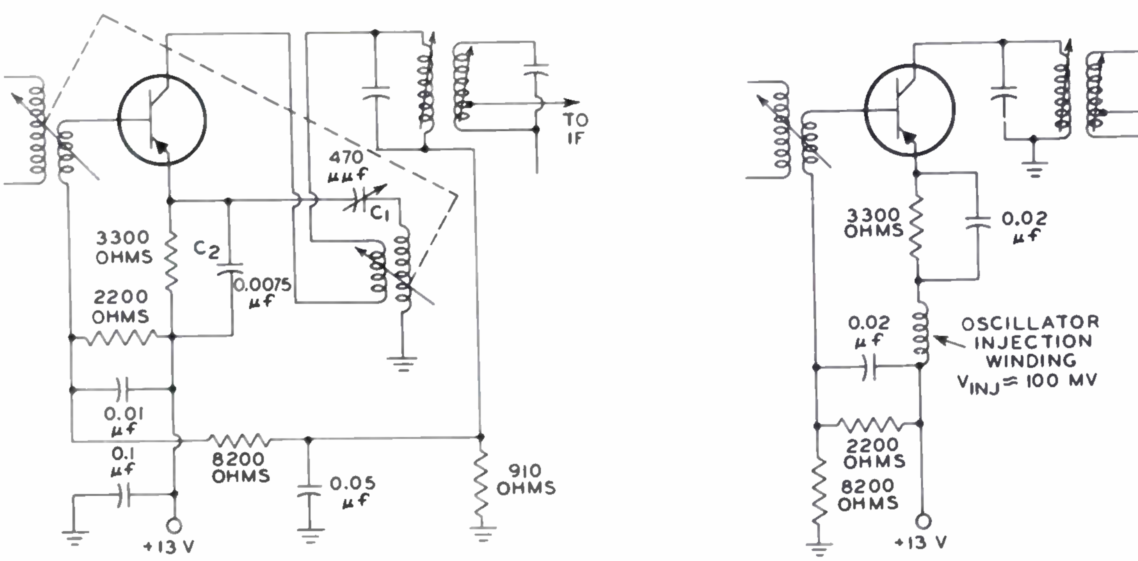


Fig. 7



1-Mc CONVERSION GAIN = 38DB
(a)

1-Mc MIXER GAIN = 40 DB
(b)

Fig. 6

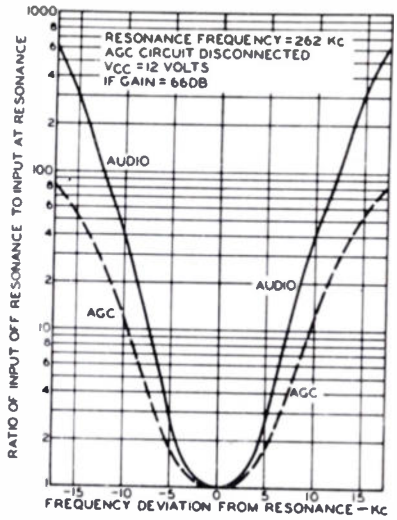


Fig. 8

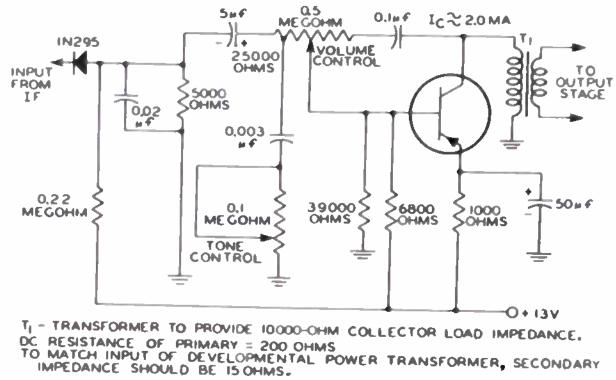


Fig. 10

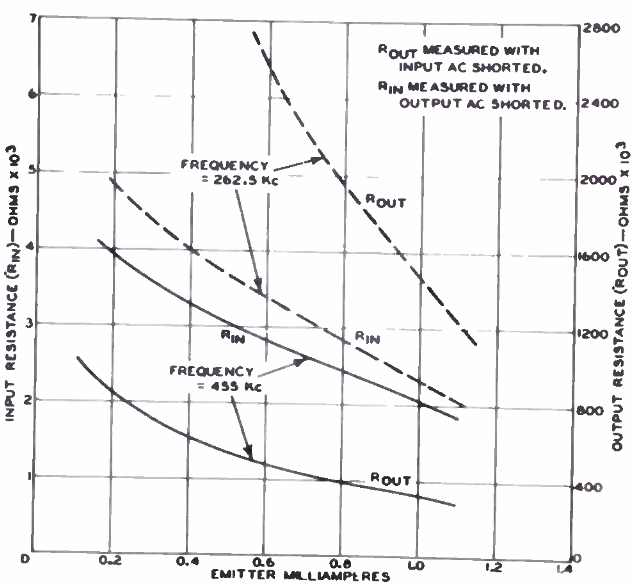


Fig. 9

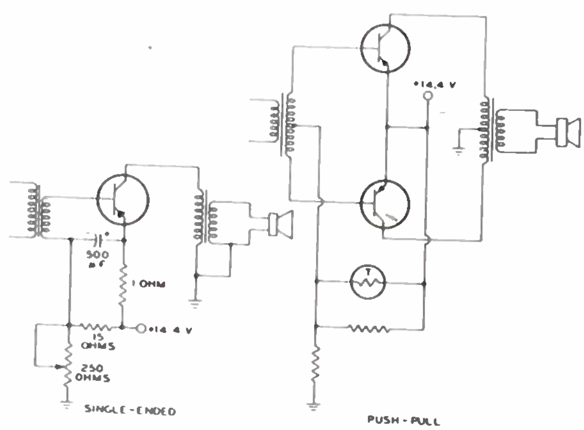


Fig. 11

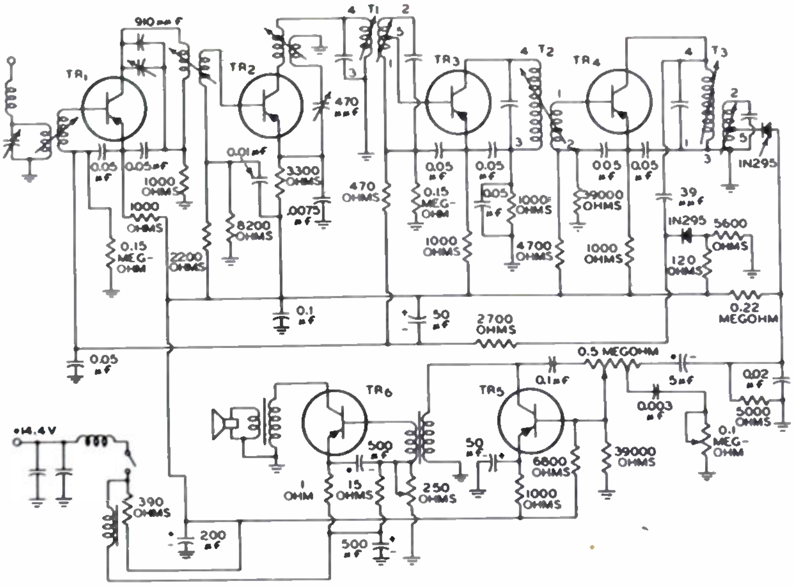


Fig. 12

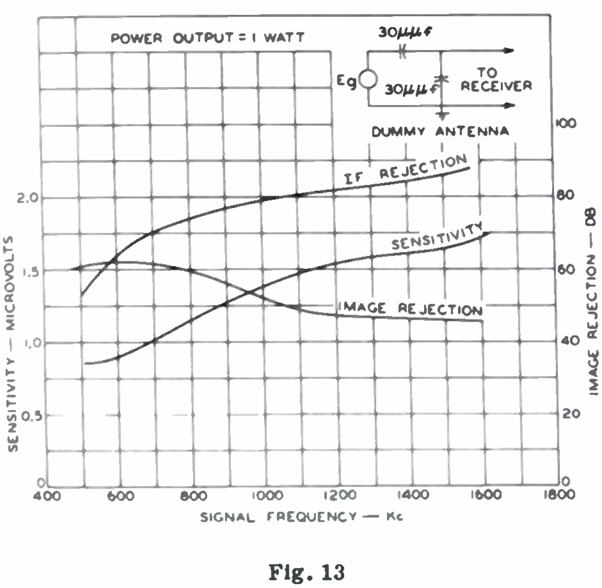


Fig. 13

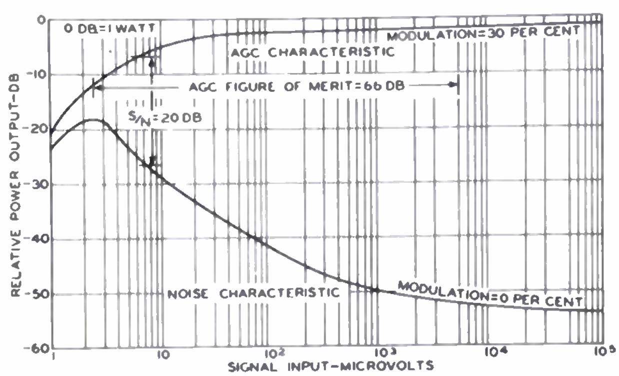


Fig. 14

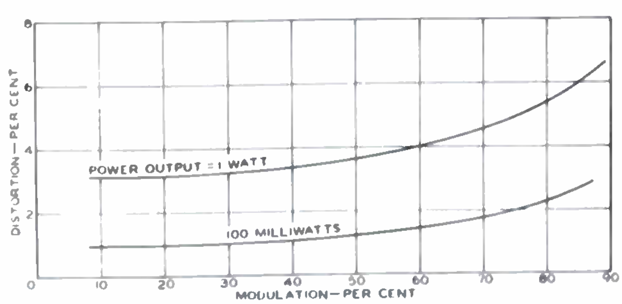


Fig. 15

CHARACTERISTICS AND APPLICATIONS OF LOW IMPEDANCE DIODES USED AS VOLTAGE VARIABLE CAPACITORS

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(The text of this paper was not available at publication time.)

DESIGN CONSIDERATIONS IN TRANSFORMERLESS SINGLE RECTIFIER TELEVISION RECEIVERS

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SUMMARY

A television receiver design using a single rectifier half-wave rectification power supply has many advantages and difficulties. The design problems which center mainly in the horizontal scanning system and video amplifier are discussed in some detail. Practical solutions to these problems are proposed and the performance data obtained is reviewed.

INTRODUCTION

It has been customary to design commercial television receivers to operate from two or more substantially different levels of rectified DC potential. This has been necessary because the scanning and video amplifier circuits have required much greater operating potentials than the other circuits. The goal of achieving complete receiver operation from one DC potential source and that source the voltage normally obtained from half-wave rectification of the standard AC line voltage is a most desirable one. The power supply cost of a television receiver averages between five and ten percent of the total cost of the entire chassis and the cost reduction made possible by the use of simple power supplies can represent a significant reduction in the complete chassis cost. In addition to the obvious economic advantage, the reduction of total input power and the simplicity so obtained directly results in greater reliability.

The dual power supply requirement of 250 to 300 volts for the deflection and video circuits and 125 to 150 volts for the other signal circuits leads to designs that are either wasteful of power or generally complex. In one such design the lower voltage source is obtained simply by dropping the full power supply output voltage to the required amount by means of a series dropping resistor. This results in substantial power loss and heat generation as well as poor regulation, directly affected by tube variations and AGC control characteristics. In another somewhat more sophisticated design the audio amplifier and intermediate frequency amplifier circuits are operated in series for direct currents and in parallel for alternating currents across the full power supply output voltage. This method is relatively complex and expensive and has the disadvantage, among others, that the sound output is made a function of the I.F. amplifier tube tolerances and of the signal level through interaction of the AGC system on the audio output tube direct current.

On the other hand, operation from a single 125-150 volt DC potential generally results in low video drive capabilities, inadequate picture tube second anode potential and undue reliance on specialized tube characteristics to maintain even this level of degraded performance.

This paper will describe some methods of providing high level reliable performance with a low voltage single rectifier power supply.

HORIZONTAL SCAN AND HIGH VOLTAGE SUPPLY SYSTEM

Figure 1-A is the equivalent circuit of a horizontal sawtooth current generating scan system. A source of potential E_{BB} is connected in series with a switch S and an inductance L_Y which has a distributed capacitance C_D associated with it. When the switch is closed, the current will build up at the rate E_{BB}/L_Y. At t_1 the switch is opened, the energy stored in the inductance flows into the capacitor C_D. The current flowing in L_Y at t_1 drops to zero and reverses direction following a cosine waveform. The time of this reversal T_R is very closely equal to one-half of the period of one cycle at the resonant frequency determined by L_Y and C_D. At t_2 with the current in L_Y equal and opposite in direction to its value at t_1, the switch is again closed and the cycle repeats. The peak-to-peak amplitude of the sawtooth current thus generated is equal to $E_{BB}T_S/L_Y$.

In the simplest practical embodiment of this circuit shown in Figure 1-C, the switch function is performed by two vacuum tubes, a tetrode and diode so connected as to permit the flow of bidirectional switch current essential to recovering the energy stored in the inductance at time t_1. Neglecting diode voltage drop the peak-to-peak current is $E_{BB}T_S/L_Y$ as in the equivalent circuit of Figure 1-A and E_B is the voltage necessary to operate the switching tubes.

In the ideal lossless case all the energy stored in the inductance at the end of the trace period is recovered and no power is withdrawn from voltage source E_{BB}. In this case the switch currents will look as shown in Figure 2-A. All the current above the sawtooth current axis is supplied through the tetrode driver tube while the current below the axis is supplied through the diode tube. Since the average tetrode and diode currents are equal and opposite in direction, it can be readily seen that no power is taken from source E_{BB}.

In the practical case energy is dissipated during retrace and not all the current flowing in the inductance at the end of the trace period is available after retrace. The forward and reverse current carrying sections of the switch now carry different currents as shown in Figure 2-B and the average tetrode current is greater than the average diode current.

Referring still to Figure 2-B, we see that the peak current flowing in the inductance at the end of the trace period, T_S, is equal to I_P, and if k is the fraction of current recovered at the start of the trace, then the peak-to-peak inductance current can be expressed.

$$I_Y = I_P + kI_P = I_P(1 + k) \tag{1}$$

By simple trigonometry the average driver and diode currents are related.

$$\frac{I_{\text{Diode Average}}}{I_{\text{P Average}}} = k^2 \tag{2}$$

and since the average current from source E_{BB} is the difference between the average tetrode and average diode current, we can state that

$$I_{BB} = I_{\text{P Average}} - k^2\, I_{\text{P Average}} \tag{3}$$

or

$$I_{BB} = I_{\text{P Average}}\,(1-k^2) \tag{4}$$

Figure 3 is the form of this circuit most often used in current practice. An auto-transformer is used to provide coupling from the driver tube to the deflection yoke, L_Y, to match the deflection yoke to the driver tube characteristics. The diode portion of the switch is also tapped down on the transformer at a turns ratio such that the average diode current is equal to the average driver tube current. From equation (2), the driver tube to diode transformer turns ratio to achieve this is k^2. Expressed algebraically and using the symbols of Figure 3,

$$\frac{N_D}{N_P + N_D} = k^2 \tag{5}$$

With the average diode and driver currents equal now as in the lossless case the current flow from voltage source E_{BB} is zero and therefore the battery of Figure 1-C can be replaced by the capacitor C_{BB} of Figure 3. The tetrode plate voltage during the trace period T_S can now be written, referring to Figure 3,

$$E_P = E_B - E_D - N_P/N_Y\ (I_Y L_Y/T_S) \tag{6}$$

where,

E_P = tetrode driver plate voltage
E_B = the supply voltage
E_D = the diode voltage drop
N_P = the number of transformer turns between the diode tap and the tetrode plate tap
N_Y = the number of transformer turns across the deflection yoke section
I_Y = the peak-to-peak required deflection yoke current
L_Y = the deflection yoke inductance

The deflection yoke current of equation (1) can be expressed in terms of the peak driver tube current and the transformer turns ratio,

$$I_Y = \frac{N_P + N_D}{N_Y} I_P + (k)\frac{N_P + N_D}{N_Y} I_P$$

$$= (1+k)\frac{N_P + N_D}{N_Y} I_P \tag{7}$$

where, N_D = the number of transformer turns across the diode section and the other symbols are as previously defined. Solving equation (7) for N_Y,

$$N_Y = (1+k)\frac{N_P + N_D}{I_Y} I_P \tag{8}$$

and substituting in equation (6)

$$E_P = E_B - E_D - \frac{N_P}{(1+k)\dfrac{N_P + N_D}{I_Y} I_P} \cdot \frac{I_Y L_Y}{T_S} \tag{9}$$

$$= E_B - E_D - \frac{N_P}{N_P + N_D} \cdot \frac{1}{(1+k)\ I_P} \cdot \frac{I_Y^2 L_Y}{T_S} \tag{10}$$

from equation (5) we saw that the driver tube to diode turns ratio was,

$$\frac{N_D}{N_P + N_D} = k^2 \tag{11}$$

then, by simple algebra,

$$\frac{N_P}{N_P + N_D} = (1 - k^2) \tag{12}$$

and substituting equation (12) into equation (10) we get,

$$E_P = E_B - E_D - \frac{1-k^2}{1+k} \cdot \frac{I_Y^2 L_Y}{I_P} \cdot \frac{1}{T_S} \tag{13}$$

and by rearrangement we find the minimum required supply voltage to be,

$$E_{BMin} = E_{BMin} + E_D + (1-k) \cdot \frac{I_Y^2 L_Y}{I_P} \cdot \frac{1}{T_S} \quad (14)$$

From equation (14) we see that the required B supply voltage is equal to the minimum required driver operating plate voltage and the diode voltage drop added together plus a voltage which is directly related to the deflection yoke energy requirements, inversely proportional to the peak driver plate current and all multiplied by a factor proportional to the system losses. For a lossless system, (k = 1), the last term in equation (14) drops to zero and the B supply voltage differs from the plate voltage only by the diode drop. For a minimum B supply voltage requirement the following rules can be stated:

1. Minimize the diode voltage drop.
2. Minimize the transformer core and copper losses.
3. Use a low impedance driver tube switch element, that is tubes of high plate current at low plate voltage.
4. Reduce the deflection yoke energy requirements.

The first three of these principles have been explored by those in the field with results that are well known. Improved vacuum tubes for scanning switching purposes such as the 12D4 and 6AU4 diodes, and the 12DQ6 high perveance driver tubes are now available. The introduction of improved core materials for scanning transformers and deflection yokes has permitted system efficiencies which would have seemed fantastic in 1946.

Let us now examine what can be done to reduce the deflection yoke energy requirements by the use of a novel deflection yoke design.

Deflection yoke performance may be defined by the measurement of several individual characteristics, including,

1. The required energy storage in the horizontal deflection coils and the power dissipated in the vertical deflection coils to produce a desired electron beam deflection.

2. The neck shadow clearance, that is the ability of the yoke to deflect the beam through the required deflection angle without the beam striking the neck of the picture tube at the angular extremes of deflection.

3. The optical characteristics, particularly with regard to geometrical distortion of the scanned raster and scanning field effects on the electron beam focus.

Table I presents a table of data of some of these characteristics showing the performance of a new so-called Extended Field Yoke designed for application in a low voltage scanning system shown together for comparative purposes with data for a conventional 90 degree deflection yoke.

TABLE I

CHARACTERISTIC	STANDARD 90° YOKE	EXTENDED FIELD YOKE
HORIZONTAL WINDING		
Energy - $\frac{1}{2}LI^2$ (Full Scan at 15 Kv)	0.0050	0.0042
Inductance	20.6 mh	20.5 mh
Resistance	30 ohms	38 ohms
Q at 60 Kc	50	60
VERTICAL WINDING		
Power I^2R (Full Scan at 15 Kv)	6.8	7.0
Inductance	42 mh	42 mh
Resistance	42 ohms	39 ohms

A reduction of between 15 and 20 per cent has been made in the horizontal energy requirements without suffering any performance loss in either the vertical power requirements or in the neck shadow clearance. As shown earlier the horizontal energy reduction may be directly used to reduce the DC voltage requirements of the horizontal system.

Good focus is maintained over the entire scanned raster and geometrical raster distortion is negligible without the use of corrective magnets.

Figure (4) shows the schematic diagram of a 135 volt horizontal scanning system using the extended field yoke. A conventional cathode coupled multivibrator is used to generate the driving waveforms that produce the proper switching currents in the output driver tube. The right hand section of this multivibrator is fed through resistors R_1 and R_2 from the 135 volt bus and from the "boosted B" bus respectively. If this section were fed only from 135 volts there would be insufficient output to cut off the driver tube during the high voltage retrace plate pulse. On the other hand if it is fed solely from the "boosted B" voltage there would be in effect DC feedback from the scanning system to the horizontal multivibrator and the interdependence of horizontal drive and output circuit performance would be accentuated. This results in added system sensitivity to component, tube, and voltage tolerances. To further insure the rapid and complete opening of the vacuum tube switch at the start of scan retrace, (a prime consideration in the maintenance of system efficiency), the cathode of the horizontal driver tube is returned to ground through an auxiliary winding on the output transformer. The small positive pulse of about

50 volts thus fed back into the cathode circuit at the initiation of retrace results in no measureable loading effects on the transformer and by reducing the voltage output requirements of the multivibrator circuit improves the stability of that circuit.

The performance data of the 135 volt horizontal scanning and high voltage system together with the data on a conventional system operated from a voltage doubler power supply are given in Table II. The zero beam high voltage output of 16 Kv is only 6 per cent less than the conventional voltage doubler system. The high voltage source resistance, however, is increased about 50 per cent. This, of course, is inherent in any reduction of deflection yoke energy requirements, since high voltage regulation is proportional to system stored energy.

TABLE II

	DOUBLER CHASSIS	HALF-WAVE CHASSIS
Deflection Yoke	Standard 90°	Extended Field
High Voltage at Zero Beam	17.0 Kv	16.0 Kv
High Voltage Source Resistance	12.3 Megohms	17.4 Megohms
Supply Voltage	230 Volts	135 Volts
Boost Voltage	503 Volts	407 Volts
Driver Tube	12CU6	12DQ6
Driver Tube Cathode Current	113 Ma	122 Ma
Retrace Time	9.5 Usecs.	9.5 Usecs.
Line Voltage for Full Scan	95 Volts AC	96 Volts AC

VIDEO AMPLIFIER

The video voltage output requirement of a television receiver is subject to many variables including such factors as picture tube size, first and second anode potentials, and picture tube electron gun characteristics. For 21 inch picture tubes of modern design a minimum of 100 volts of video drive from sync peaks to white on a fully modulated television signal is required for commercial acceptance. Other considerations of importance include, sufficient linearity to provide good grey scale rendition, to avoid compression of the synchronizing signal portions of the composite video signals permitting use of the video amplifier as a synchronizing signal amplifier, good 4.5 megacycle amplification with negligible picture information cross modulation, and limiting of impulse noise bursts to assist in noise free synchronizing signal separation.

The transfer characteristics of a typical video output tube, the pentode section of a 6AU8, using a plate load of 5,000 ohms and a supply voltage of 250 and 135 volts respectively are shown in Figure 5. When the plate circuit is operated from a supply voltage of 135 volts the load line falls below the knee of the pentode plate characteristics below about -1 volt of grid bias with the resultant flattening of the transfer characteristic. A maximum output of about 70 volts can be obtained if good linearity is to be maintained. In order to make allowance for tube and component tolerances a maximum linear output of 60 volts is all that can be expected consistently. It is obvious that the requirements of output and linearity cannot be met by this amplifier operating at a plate supply potential of 135 volts. Huskier video output tubes or different plate loads offer no solution because of the fundamental pentode knee voltage limitation in practically realizable tube construction. These difficulties may be overcome by modulating both the cathode and signal grid of the picture tube with opposite polarity video signals in "push-pull" fashion. Figure 6 is a schematic representation of this video amplifier circuit.

The video detector output is fed to the grid circuit of the pentode section of a type 6BH8 tube. Suitable cathode self-biasing is used so that linear amplification of the video signal is obtained for all normal video detector levels. The voltage gain of the pentode stage is between 20 and 25 with a maximum output capability of approximately 60 volts. The output of the pentode section is used to drive the cathode of the picture tube in conventional fashion. A portion of the output voltage of the pentode section is also used to drive the grid circuit of the triode. The exact amount of drive to the triode section is determined by a potentiometer contrast control in the grid circuit of the triode section. This tube is so biased that some crushing of the synchronizing signal components of the composite video waveform occurs at the maximum contrast setting in order to realize the maximum drive capabilities of this section. Since sound and synchronizing signals are taken from the output of the pentode section before this distortion occurs and since the distortion occurs in the blacker-than-black region of the video signal there are no undesirable effects from this method of operation. A maximum output of approximately 50 to 55 volts of video are obtained from the triode section and since the video polarity has been reversed in the triode this signal is suitable to drive the signal grid of the picture tube. A vertical blanking signal is applied to the cathode of the triode section to provide vertical retrace blanking. The polarity of the video signal is such that plate current cutoff limiting of noise impulses occurs in the pentode section where it is desirable because of its use as a sync amplifier.

Figure 7 is a dual beam oscillographic display of the video drive to picture tube grid and cathode. The sum of the two signals is such that between 100 and 120 volts of effective video drive are made available between picture tube grid and cathode.

POWER SUPPLY

The entire DC power supply for a complete single rectifier television receiver is shown in Figure 8. The four electrolytic capacitor sections, all the capacitor filtering required by the receiver, are contained in one single can. A single silicon rectifier is used. These rectifiers, with only 1.0 to 1.5 volts forward drop at the full receiver load current of 315 Ma make possible a filtered DC power supply output of 135 volts. Figure 9 compares the DC output voltage of silicon, selenium, and vacuum tube rectifiers in the same half-wave test circuit. The silicon rectifier provides about 11 per cent more rectified output voltage than the selenium rectifier at 300 Ma load current.

COMPLETE RECEIVER

The principles and circuitry discussed in this paper have been combined in the Westinghouse V-2361 21 inch television chassis. A photograph of the complete chassis is shown in Figure 10. The total power drain of the receiver is 120 watts as compared with 170 watts for an equivalent receiver using a 235 volt voltage doubler power supply system. As a result of the reduced power drain, component temperatures average 10 degrees centigrade less in the single rectifier receiver than the equivalent components in a voltage doubler receiver of similar mechanical construction.

CONCLUSION

Some of the advantages of television receivers using single rectifier half-wave rectification of the AC line voltage to obtain DC supply voltage have been discussed. Methods of obtaining high scanning and video performance standards at the supply voltages thus obtained have been reviewed and a receiver incorporating these methods has been described. In conclusion it may be stated that in manufacturing ease, field performance and service reliability the chassis has demonstrated the benefits of this design approach.

The author wishes to acknowledge his indebtedness to Messrs. C. A. Moore, H. W. Claypool, C. Ondrejik, R. R. Tesno and D. Knoebel of the Television-Radio Division of the Westinghouse Electric Corporation, and to Mr. F. T. John who at the time of this work was associated with the Penn-Tran Corporation.

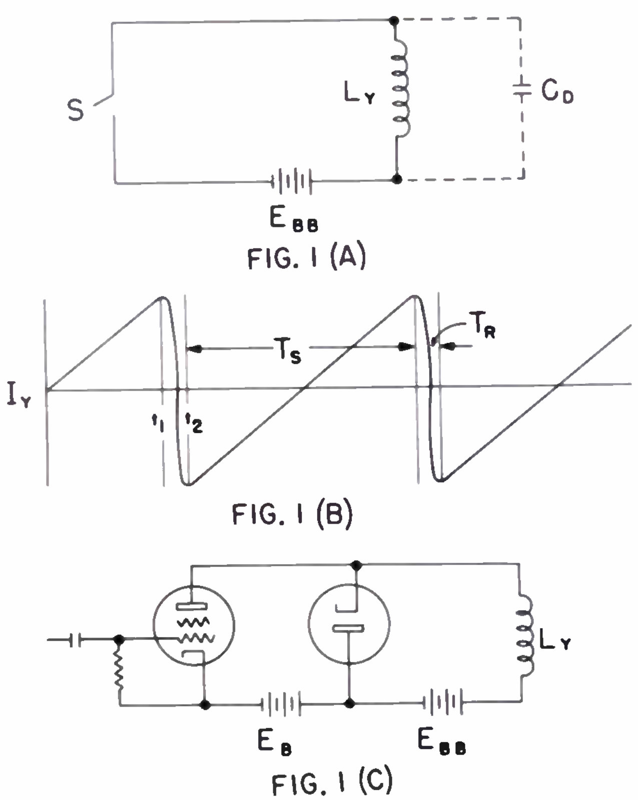


Fig. 1 Equivalent horizontal scan circuit and waveforms.

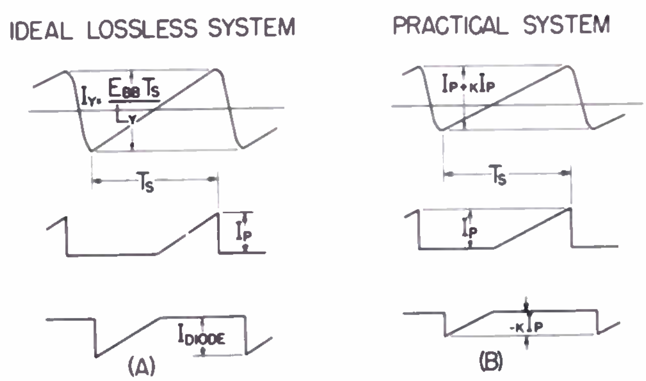
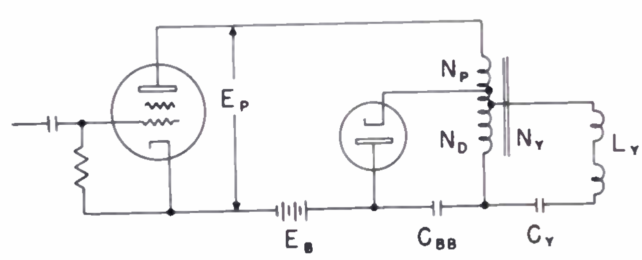


Fig. 2 Deflection yoke and switching-tube waveforms.



$$E_P = E_B - E_D - K \left[L_Y I_Y^2 \right] \left[\frac{1}{I_P} \right] \left[\frac{1}{T_S} \right]$$

Fig. 3 Transformer coupled horizontal scan system.

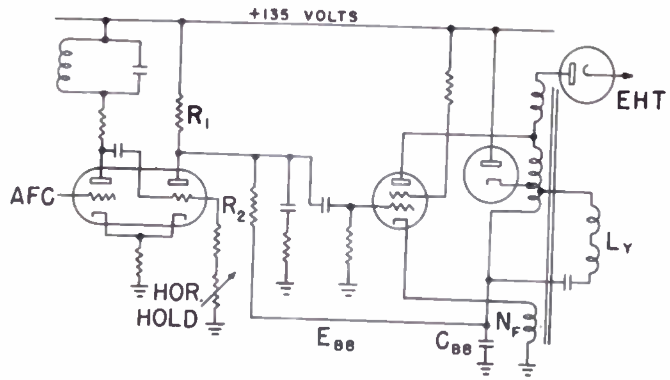


Fig. 4 Complete horizontal scan system.

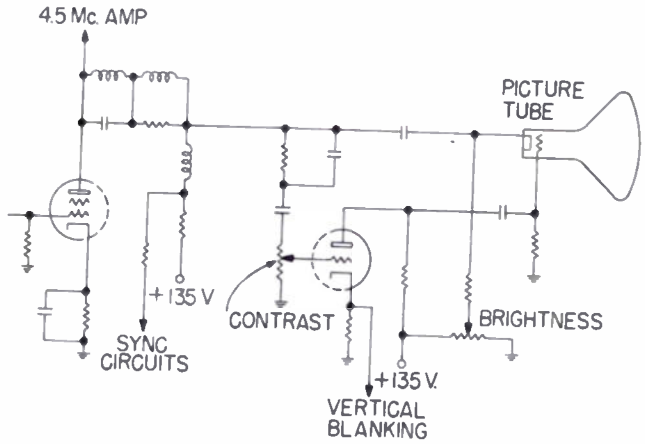


Fig. 6 "Push-pull" video amplifier.

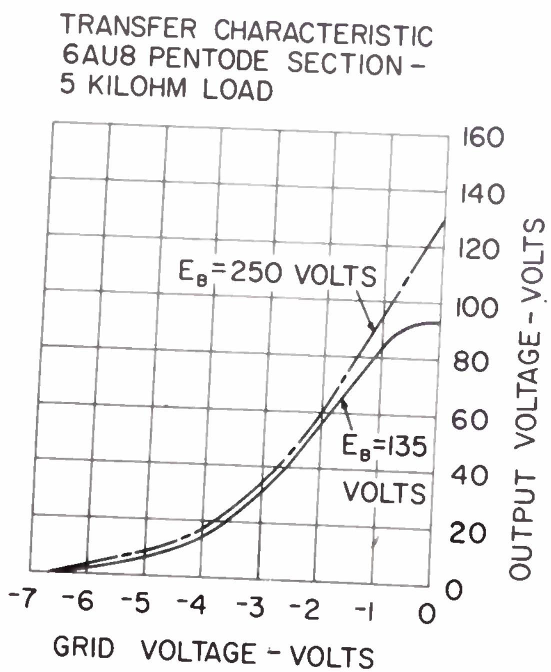


Fig. 5 Video amplifier transfer characteristics.

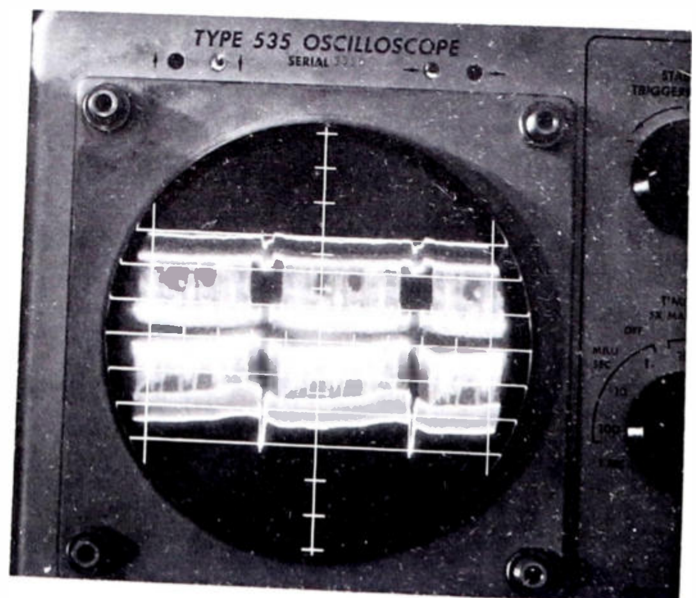


Fig. 7 Oscilloscope display of video output.

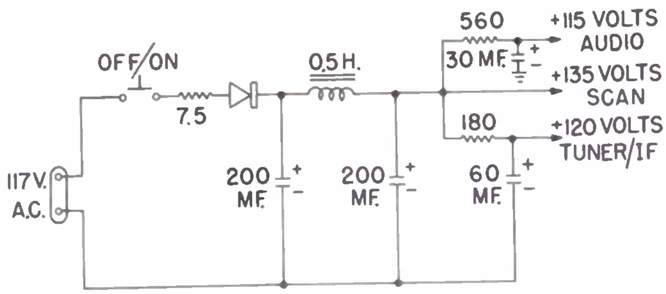


Fig. 8 Complete power supply.

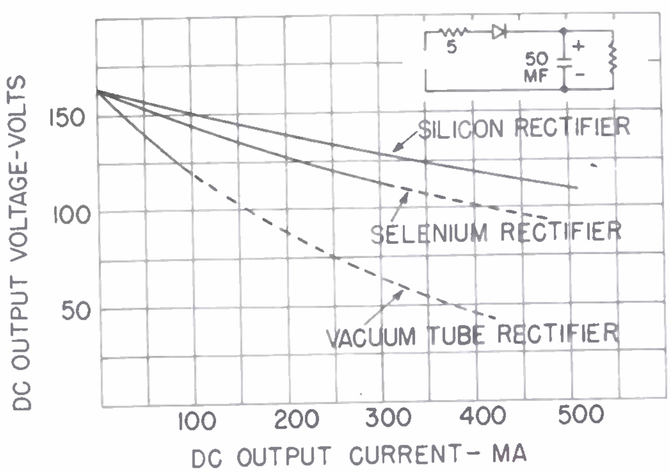


Fig. 9 Comparison of silicon, selenium and vacuum tube rectifiers.

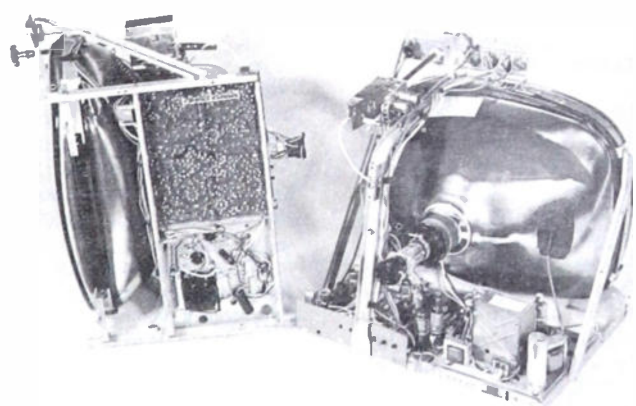


Fig. 10 Complete half-wave rectifier television receiver.

PROBLEMS IN TWO-DIMENSIONAL TELEVISION SYSTEMS

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Since the announcement of the "Sylvatron" last July 2nd, there has been some public speculation about the early commercial realization of a flat television display using electroluminescence. The purpose of this presentation is to describe several approaches to this objective, to point out some problems involved and to talk of progress toward solutions.

Many companies are in this race, as the stakes are high. Several have given demonstrations of image converters, while the literature reveals a number of schemes for scanning a field and for displaying information. A scheme for the electroluminescent display of color television has even been suggested.

An essential part of this form of television display is the phenomenon of electroluminescence in which certain kinds of phosphor powders, when subjected to a varying electric field, extract energy from that field and transform it into light. Let us look at Fig. 1. In the simplest form the lamp comprises a glass plate having a transparent, conductive inside surface coated with a thin layer of the electroluminescent phosphor embedded in a suitable dielectric and backed by another conductor which may be a metal foil. An alternating potential of several hundred volts is applied between the metal and the conducting coating. The phosphor is generally ZnS with a copper activator and a coactivator such as chloride, bromide, iodide or aluminum. There are other electroluminescent phosphors, but I shall not take the time to mention them here. I shall say more about the properties of the phosphors later, as they account for some of those problems that beset the path to commercial realization of a television display.

In the lamp of Fig. 1, the glass plate is rendered conducting on one side by the well-known technique of spraying it with tin chloride solution while it is at a temperature near its softening point. Lamps currently on the market employ chiefly a fused glass frit for the dielectric, partly because of the frit's ability to produce a lamp with long life. Lamps operated at 110 v, 60 cycles, exhibit a brightness that is 65% of the initial value after 40,000 hours of operation. Actually brightness often increases to 150% of the initial value during the first 100 hours of operation. The life expectancy depends upon operating conditions, being reduced by increasing frequency and operating voltage. Some organic plastics produce lamps which are initially brighter but which, to date, have tended to deteriorate rapidly.

When a lamp is made using the frit dielectric, the finished product looks much like a sheet of steel coated with enamel like that on the kitchen sink. You will see one later.

I should like to describe to you briefly how the frit-embedded lamp is made, as it embodies a number of convenient techniques for making the information-display devices of the type in which we are interested today. Let us refer to Fig. 2.

On the steel sheet a layer of plain, unloaded frit is first sprayed. The frit is a finely ground-up glass of special composition to have the same coefficient of expansion as steel and is suspended, often in water. After firing this on, frit mixed with the phosphor is sprayed on and fired; sometimes in several coats. For best results, the frit must be carefully selected to maximize the electroluminescence. A suitable tin compound is then put on to produce a transparent, conducting layer in a manner similar to that described earlier for making the glass plate conducting. On top is then placed an insulating transparent glaze by the process of spraying and firing.

Now let me say something about the brightness characteristics of electroluminescent powders and lamps. When a lamp of the type described above and designed for 600 v operation is excited by ac, one obtains the brightness curves such as those in Fig. 3. These curves fit the following empirical equation:

$$B = A \exp^{-(b/v)^{\frac{1}{2}}}$$

where B is brightness in foot-lamberts, v is in rms volts and the parameters A and b are independent of voltage over many orders of magnitude. In the frequency range from about 50 to 10,000 cycles per second, the parameter A varies roughly as $f^{0.6}$ where f is frequency in cycles per second.

This equation is of interest because it shows two things. First, the brightness increases with frequency within the range covered. Actually there is a limit set to the increase in brightness with frequency by the capacitance of the lamp and the resistance of the transparent conducting film. This limit is determined largely by geometrical configuration and in very small, almost minuscule lamp sizes can reach several megacycles. Second, the equation indicates an exponential-like increase in brightness with voltage, which voltage is limited, of course, by electrical breakdown. The exponential form of this relationship turns out to have its virtues, as will be brought out later in connection with scanning. The practical brightness obtained with such a lamp is currently limited by dielectric breakdown to a value considerably below the instantaneous brightness now attainable by the scanning spot on a picture tube.

As one might expect, the lamp acts as a lossy capacitor. The "loss" is only partially attributable to the light-producing process, since some of the power consumed is transformed into heat by competing processes. For this reason the efficiency currently obtainable is low, being normally about 2 lumens/watt, although 10 lumens/watt has been obtained. For comparison, a 100 w incandescent lamp has an efficiency of 16 lumens/watt.

The efficiency and the capacitance present two other problems that must be surmounted in the eventual design for commercial television display. The efficiency effects the driving power required and the resulting heating of the display panel, while capacitance tends to load the driving circuits.

Let us now turn our attention to television. To display the intended picture one must scan, synchronize and modulate. To these three fundamental problems must be added the obtaining of high brightness levels without excessive persistence of image and the all-important task of making the device cheap enough to compete. On this last point we are all painfully aware that the public has demonstrated its unwillingness to pay very much more for an improvement even if the improvement is color. It appears doubtful that the public will pay much extra to have its television as a mural decoration!

I should like to show these problem areas on the next slide, Fig. 4, as they serve as a check-list.

One of the first schemes which come to mind is the use of an image converter or intensifier and a small projection tube, as depicted in Fig. 5. I expect to show you such an image converter today using, in place of the projection tube, an 8 mm motion-picture projector fitted with a deep red filter. The image converter is an electroluminescent lamp combined with a photoconductor, as shown in Fig. 6. It comprises, first, a sheet of glass rendered conductive on the side next to the electroluminescent layer. An almost opaque, non-conducting layer between the electroluminescent layer and the photoconductor prevents optical feedback. Finally, there is a very fine mesh serving as the other electrode. I should like to say something more about the photoconductor and its function as we shall encounter these in other more complicated devices.

The photoconductor is usually cadmium sulfide, although others have been used. In certain devices its resistivity can be changed by three or four orders of magnitude upon the application of light. A typical curve is shown in Fig. 7. The knee in the curve is due to the effect of dark current or leakage. The dark resistance and the range of resistances over which the photoconductor works can be changed by the level of activation and the method of preparation. This makes it possible, to some degree, to design the photoconductor for the application.

For functional analysis, the image converter may be considered as broken up into elemental areas, though actually the surface is unbroken. Each such elemental area can be pictured as a small electroluminescent area in series with a photoconductive area, as shown in Fig. 8. It is obvious that if the photoresistance were to go to infinity in the dark, there would still be voltage on the lamp due to the capacitive reactance of the photo-condenser. If the photoconductive layer and the electroluminescent layer were of about the same thickness and dielectric constant, then in the dark approximately half the voltage would be across the electroluminescent layer as the high equivalent shunt resistances of the photoconductive and electroluminescent layers would have little effect. Here the exponential nature of the electroluminescent lamp response can be a distinct boon as it may be used to reduce the brightness of the phosphor under conditions of no input illumination. Under no input illumination, the voltage across the electroluminescent layer would only drop to about half of maximum, but the light would decrease by a factor of 5 or so. Under full input illumination, of course, the equivalent shunt impedance of the photoconductor drops almost to zero and nearly full supply voltage appears across the electroluminescent layer.

In image converters, however, it is quite customary to increase the impedance of the photoconductor under dark conditions relative to that of the electroluminescent layer by making the photoconductor appreciably thicker or by other means. Thus the output brightness for zero input illumination can be reduced by appreciably more than a factor of 5. I shall demonstrate this later in connection with the cross-effect.

I have omitted from the discussion thus far the response and decay times of the photoconductor. Many demonstrations of the past have been limited to still pictures because the decay time constant has been of the order of seconds. By proper compounding of the photoconductor and proper choice of input illumination level, the time constant can be reduced below a tenth of a second, permitting the showing of motion-pictures.

But let us return to the case of the projection television tube and the image converter. This combination fulfills the first three criteria, namely scanning, synchronization and modulation, but it does not satisfy the commercial requirements for television on brightness and cost. Unlike the motion-picture projector, only one picture element is fully illuminated at a time. Since there are some 180,000 picture elements, each one must be very much brighter than the apparent picture brightness. Were there no phosphorescent decay of the projection tube screen and no lag in the photoconductor response, a picture element in the highlights would have to achieve 180,000 times the highlight brightness of the picture! Actually the decay and the lag reduce this number appreciably but probably not enough to come within the range of electroluminescent materials which are presently available commercially. One remedy appears to be to employ storage, that is to have each element of the image converter remember its instruction from the projection tube and to emit the appropriate light level for most of a field of time, then go off automatically. There are ways of getting at this, and I shall mention one of them shortly.

Suppose now, in Fig. 6, we permit some optical feedback from the electroluminescent layer to the photoconductor by reducing of opacity of the middle layer, then we have a system with positive feedback. If the feedback is strong, once the lamp is turned on feedback will keep it turned on. I'll show you a plaque in which strong, positive feedback is employed. In such a plaque one must prevent the lighted area from spreading as the feedback light also goes out laterally to adjacent areas. To accomplish this objective, we break the lamp into discreet, small areas. A schematic cross-section of such a plaque is shown in Fig. 9. As can be seen, the photoconductive layer has been deposited on the sides of discreet glass pillars positioned on top of the electroluminescent layer. When light falling on a pillar initiates electroluminescence in the area beneath, the feedback light is restricted almost entirely to going up the same pillar. Hence only that pillar remains bright and spreading is prevented. Remember that in this plaque each element goes on fully when triggered by outside light and remains on until the whole plaque is turned off. Each element is, therefore, bistable. It will remain either off or on until transferred to the other state.

There are, however, other levels of optical feedback that can be used in Fig. 6. If feedback is not quite adequate for full bistability, the time required for return to the dark condition after triggering is increased, often to a matter of seconds. I'll show you this with a simple experiment.

Remember, we got into this because we were seeking a way of "storing" the picture between scans. How then do we achieve this? It can be shown experimentally and mathematically that both the length of dwell or the duration of the light and the brightness depend upon the amount of the exciting flash of light. This does sound easy, but the condition is a critical function of feedback level and temperature, while the photoconductor exhibits a dismaying memory of the amount of light it saw during the previous several scans as well.

Having now dealt with the matter of brightness, let us turn to the matter of cost. I think that I shall refrain from talking extensively about this subject as prospective costs are a matter of conjecture for the various laboratories engaged in this research. However one does seem to have all of the elements of a projection television system plus an additional image intensifier with its power supply. Further such a system of mural television just doesn't fit nicely on a wall unless one is prepared to cut a hole into the diningroom. Perhaps, then, the scanning, synchronizing and modulating functions could be performed using electroluminescence on a flat sheet. Let us consider the following scheme.

If one starts with an electroluminescent lamp and figuratively slices the back conductor into strips running one way and the transparent conductor into strips running the other, he would have a cross-grid or x-y panel, so called because of its similarity to Cartesian coordinates. Let us look at Fig. 10. In this case the x-y panel is fed by two commutators which, for standard television, would probably be electronic. After the line-rate commutator has designated a line, the dot-rate commutator successively selects the dots along the line. As each is selected a surge of voltage is applied and then removed, producing two superimposed flashes at the intersection of the selected x and y strips -- but this isn't all! One gets a noticeable cross with a bright spot at the intersection. I will show you this effect. The reason for this is easy to see. The strips unused at any instant -- both x and y -- tend to remain at ground potential, while the selected x strip is pulsed positively and the selected y strip is pulsed negatively. This is because they are coupled capacitively about equally to the selected x strip and to the selected y strip. Hence half-potential appears at all unwanted intersections along the selected x strip and also those along the selected y strip. Just as I mentioned earlier, the exponential relationship between brightness and voltage and the effective threshold of visibility under ambient viewing conditions, can be made to aid us. Half-voltage produces appreciably less than half-brightness. The demonstration will show this, I believe.

We have thus achieved scan and presumably synchronization. To achieve modulation one can modulate the voltage made available to the commutators taking into account, of course, that highly non-linear brightness characteristic of the phosphor already mentioned.

There are other ways of suppressing the cross-effect. I will show you one of these. It employs external circuitry without a change in the panel. By incorporating rectifiers in the panel, the cross-effect may also be eliminated.

At this stage I have now described another way of scanning, synchronizing and modulating which could be combined with the storage-type of image converter to produce mural television. My chemist friends now rise to remind me that these functions are all critical and require a degree of uniformity among elemental bits that may well tax commercial capability. Then, too, there are the effects of temperature, memory and aging to be considered. Further, we have not dealt with costs. It would appear, offhand, that 490 conductors leading to the vertical commutator and perhaps 550 to the horizontal would not be inexpensive. If this seems unattractive to you, there are approaches which time will not permit us to explore. For example, the use of a loaded transmission or a lumped constant line for horizontal scanning with a lumped constant line for vertical scan has been reported.

I have spoken here chiefly of schemes employing essentially optical scanning and optical feedback. Though many of these ideas have been demonstrated, there remains a great deal to be done in extending these and other schemes to meet commercial television requirements. All in all, it appears that the commercial advent of mural television is not imminent. Before the day of mural television arrives, however, there will be ample use for the technique of the Sylvatron and similar devices in other forms of data display such as battlefield surveillance, missile tracking, stockmarket listings and many others where the high speed and low costs of mural television will not apply.

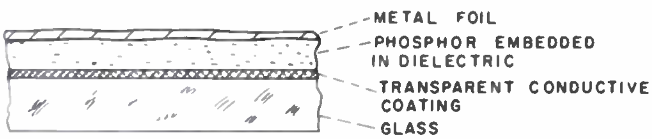

Fig. 1

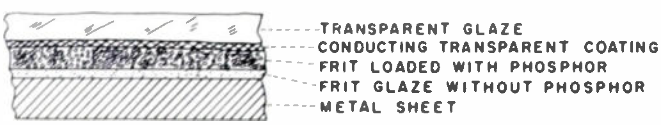

Fig. 2

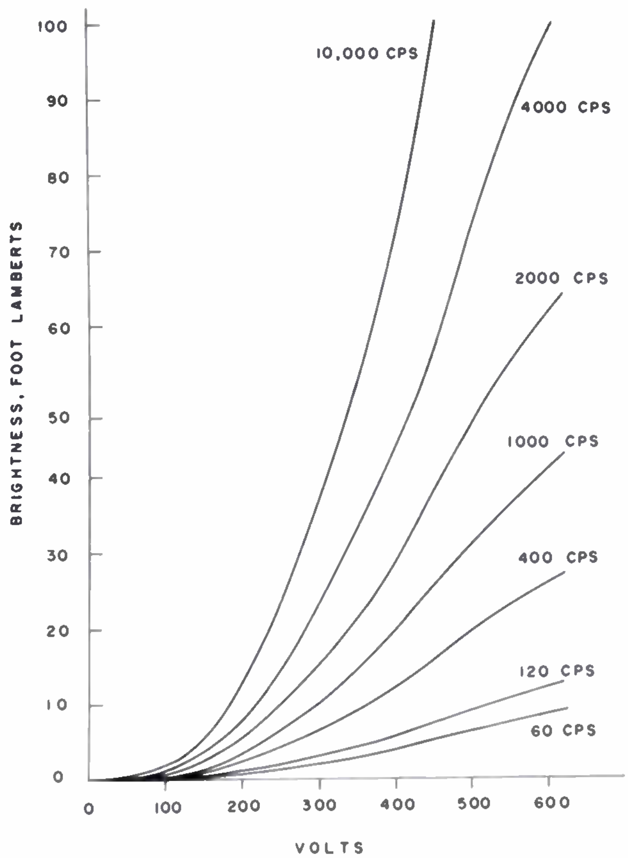

Fig. 3

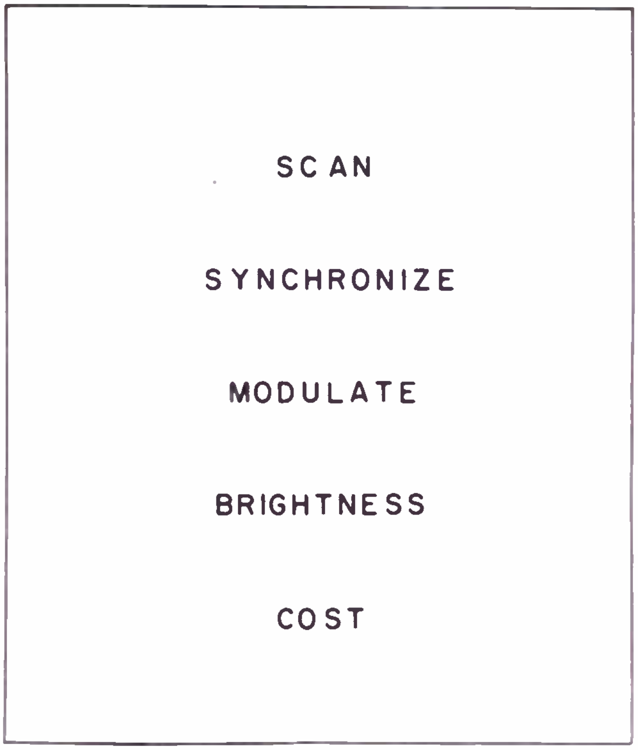

Fig. 4

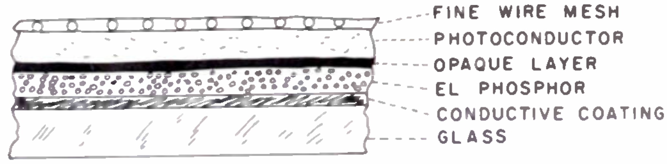

Fig. 6

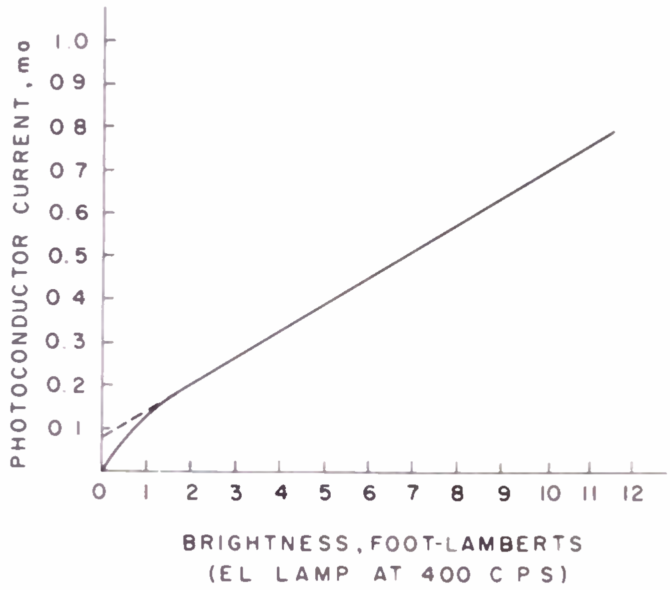

Fig. 7

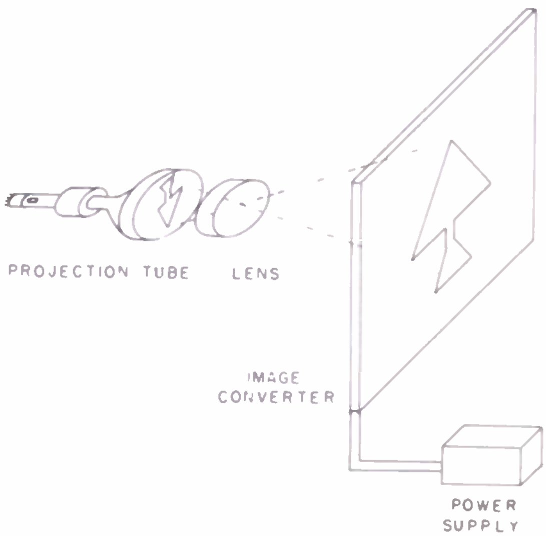

Fig. 5

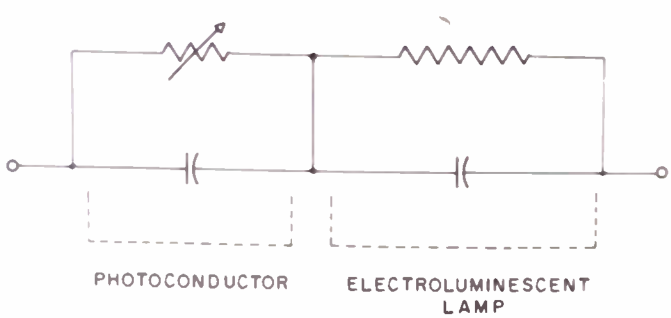

Fig. 8

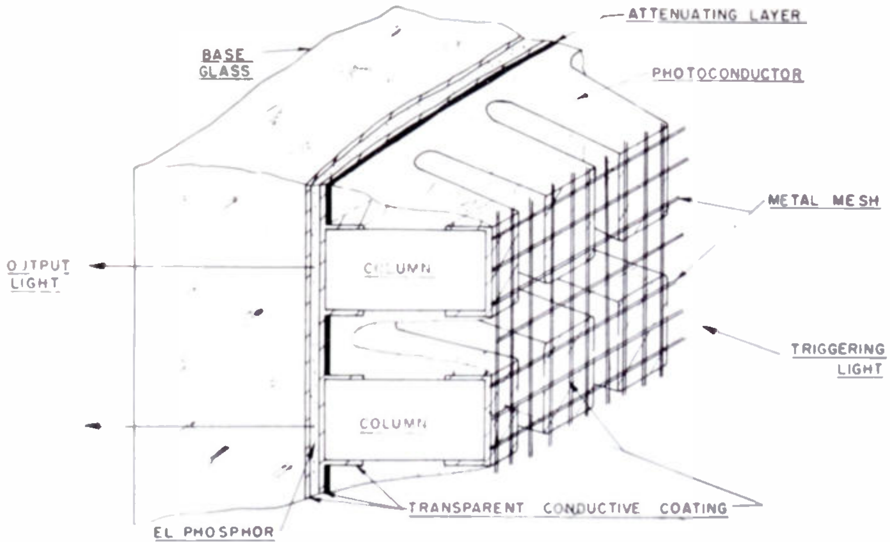
ATTENUATING LAYER
BASE GLASS
PHOTOCONDUCTOR
METAL MESH
OUTPUT LIGHT
COLUMN
TRIGGERING LIGHT
COLUMN
TRANSPARENT CONDUCTIVE COATING
EL PHOSPHOR

Fig. 9

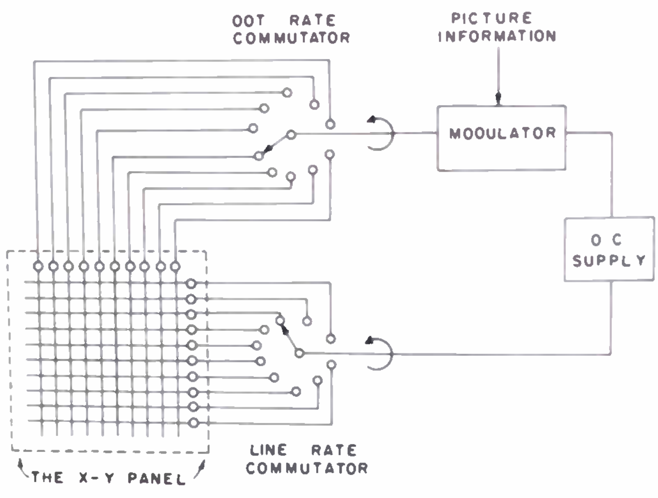
DOT RATE COMMUTATOR
PICTURE INFORMATION
MODULATOR
D C SUPPLY
LINE RATE COMMUTATOR
THE X-Y PANEL

Fig. 10

ON THE QUALITY OF COLOR TELEVISION IMAGES AND
THE PERCEPTION OF COLOR DETAIL

Otto H. Schade
Radio Corporation of America
Harrison, New Jersey

(The text of this paper was not available at publication time.)

IMPROVEMENTS IN DEFLECTION AMPLIFIER DESIGN

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Introduction

The horizontal-deflection amplifier has long been recognized as one of the most critical tube applications in a television receiver. The human eye continually monitors the operation of this tube in terms of its effect on picture scan and brightness. Thus, any loss in performance is immediately evident to the viewer.

The requirements imposed on the horizontal-deflection amplifier are quite severe. The energy it delivers to the horizontal-deflection system must perform many functions, such as providing horizontal scan, high-voltage to the picture tube, focusing voltage, filament voltage for the high voltage rectifiers, keying pulses for the AGC system, and a feedback timing pulse for some types of horizontal-oscillator control systems.

It can be seen that the horizontal-deflection system is being monitored not only by its ability to provide sufficient scan and high voltage, but also by other circuits whose operation depend upon the signal derived from the horizontal system.

The relatively high failure rate of the horizontal-deflection amplifier was pointed out in the paper "Progress in TV Receiver Reliability" which was presented at the IRE Fall Meeting in Toronto, by Mr. E. H. Boden of Sylvania.

Fig. 1, borrowed from Mr. Boden's paper, illustrates the comparative failure rates of tubes used in the four most critical applications in TV receivers. The data is based on recorded observations of several-hundred commercial receivers of ten different manufactures over a three-year period. Tests were conducted at a line voltage of 130 volts for 1500 hours. This accelerated the tube failure rate by roughly 2.4 times the rate encountered at normal 117 volt operation.

In recognition of the problems encountered with horizontal-deflection types, Sylvania undertook an extensive program to develop a fundamentally improved tube for horizontal-deflection service.

The result of this investigation, the Sylvania "Framelok" Type 6FH6, is an unusual tube that represents a radical departure from conventional tube structures. Even though it embodies a new concept in tube design and requires new manufacturing techniques, it ultimately will be produced at a high-volume level and at a relatively low cost. A sectional view of the 6FH6 is presented in Fig. 2. It will be noted that the conventional, round, grid-support rods with the familiar wound grid lateral wires are not present in this structure; instead these are replaced with a solid-frame structure that supports the grid lateral wires at an exact ninety degrees in relation to the sides of the frame.

I hope to make it evident in subsequent sections of this presentation, that the unusual characteristics and structure of the 6FH6 offer advantages not to be found in conventional deflection types.

Reasons For Tube Failure on Life

Life data collected on deflection amplifier types discloses that short circuits, intermittent arcing between the elements, and screen emission account for the greater part of the high failure rate of these types. Experience has shown that these failures are aggravated by excessive screen dissipation and that a failure can occur in either of two ways--first, the screen may simply burn up or become bowed and short to the other tube elements. Second, the screen may become a primary emitter, and the uncontrolled flow of current from the screen to the plate will cause reduced scan and high voltage.

The usual expedients, such as heat-radiating tabs attached to the grid siderods, two grid connectors and heavier leads in the stem, are only moderately effective in minimizing the adverse effects of heat, because they are too far removed from that area most susceptable to heat, the grid lateral wires.

"Framelok" Grid Structure

The view shown of the "Framelok" grid, in Fig. 3, clearly illustrates how this structure can contend more ably with those factors that influence tube life. The mass and large-surface cooling area of the frame makes it inherently a more effective heat sink than the siderods of the conventional screen grid. Cooler operation of the lateral wires is further enhanced by their short, straight, heat-conducting path to the supporting frame.

Fig. 3A is a view of that surface of the screen grid closest to the control grid. The reverse side of the screen grid is shown in Fig. 3B. Each of the grid lateral wires is under tension and is firmly anchored to the inner surface of the ridged frame by a nick-and-peen operation. Nicking and peening very aptly describes the process of grooving the inner surface of the frame and peening the grid wire into place.

The tension applied to the wire lateral prevents sagging, and bowing of the grid wire and vertually eliminates the possibility of shorts between the cathode, control grid, and screen grid.

The end-view of a complete tube mount shown in Fig. 4 illustrates that two of these frames are contained in a complete grid. It will be noted that each frame is bent at the outer edges to give added rigidity. This formed edge is also a valuable aid in preventing jamming of the grid wires during the assembly of the tube. This view also demonstrates the exact parallel alignment, symmetry, and spacing that is attained between the elements, because of the absences of grid wire warping and bowing, a defect that occurs with the conventional wound grid.

Each of the two frames, comprising a complete grid, is accurately positioned by four mounting tabs that extend through the upper and lower mica supports. A view of the upper mica contained in Fig. 5 illustrates that both halves of each grid are joined together by a U-shaped strap.

This part of the presentation has dealt primarily with the structural design considerations of the horizontal-deflection amplifier that influence tube life. It was shown in the course of the discussion how, through highly effective cooling and freedom from distortion, the "Framelok" grid contributes materially to improved tube life by minimizing three of the major causes for tube failure: short circuits, intermittent arcing between the elements, and screen emission.

Limiting Factors in Set Performance

Improvements in deflection components, more efficient circuits, and the small-neck picture tube have made the 110° television receiver a reality. However, the circuit-design engineer still must contend with those conflicting requirements represented by marginal scan and high voltage at reduced line voltages, as well as excessive screen dissipation and short tube life of the horizontal deflection amplifier at high line voltages. It has already been indicated in the preceeding comments that some of the causes for tube failure are reduced by the 6FH6 structure. Further relief from these circuit design considerations is offered by the reduced screen dissipation made possible by virtue of the high plate-to-screen-current ratio of the 6FH6. It subsequently will be shown that the screen dissipation of the 6FH6 is approximately 30 percent lower than that of the 6DQ6A in a horizontal scan application. In addition to this, the 3.6-watt screen dissipation rating of the 6FH6 gives further assurance that screen dissipation will not be the limiting factor in set performance.

Plate-Screen Current Ratio

The reduced screen current and excellent plate-to-screen-current ratio of the 6FH6 is evident in the zero-bias plate- and screen-current curves displayed in Fig. 6. A twenty-to-one ratio is realized at a plate current of 300 MA and a screen current of just 15 MA at those plate and screen voltages shown in Fig. 6. Corresponding curves of the 6DQ6A disclose an eleven-to-one ratio at a plate current of 300 MA and a screen current of 27 MA.

A brief description of the tube design considerations that affect the grid alignment of conventional wound grids will help to explain how a high plate-to-screen ratio is achieved in the 6FH6.

If low screen current is to be realized, it is essential that the screen lateral wires be directly in line with, or lie in the shadow of, the control grid wires. It is evident in the wound grids shown in Fig. 7B that even though the turns-per-inch of both the control grid and screen grid are the same, the difference in the major diameter of these grids causes the grid laterals to have a different slope and that the ideal situation just described exists only at the midpoint. As the wires extend away from the midpoint, more of the screen grid becomes exposed to the cathode current. Obviously, this results in a less favorable plate-to-screen-current ratio. In addition, the screen intercepts cathode current that might have been realized as useful output in the plate circuit. It should be pointed out, that the grid wire cross-over has been exaggerated in this slide to make it more obvious to the viewer.

Compare this with the parallel-planar alignment that is attained with the "Framelok" grids shown in Fig. 7A. Cross-over of the grid wires has been completely eliminated and the ideal alignment exists over the entire length of each lateral wire. The plate-to-screen-current ratio can be further improved if the diameter of the screen lateral wire is smaller than that of the control grid. This was done with the 6FH6 "Framelok" grid. These are the two features of the "Framelok" structure that account for the high plate-to-screen-current ratio of the 6FH6 deflection amplifier.

High-Voltage Cutoff

Another aspect of TV set performance, insufficient high voltage and brightness, is often caused by a remote cutoff characteristic in the horizontal deflection amplifier. The spread in the cutoff characteristic that might be expected in tube production did not represent a problem with the TV set designs of several years ago, which invariably included a grid-drive control. With the current trend towards reduced costs and simplification of circuitry, this control has been eliminated and tighter cutoff specifications had to be adopted by the tube industry.

High voltage cutoff in deflection amplifier types is affected by the grid-cathode spacing, alignment of the grids, the care taken to avoid jammed grid turns, and conduction around the end turns of the grid structure. It can readily be perceived from the preceeding discussions how each of the first three items has been improved upon with the "Framelok" grid structure. But the last of these items warrants some further discussion.

Fig. 8 shows a section of the control grid that is adjacent to the mica support. Experience has shown that this part of the grid structure is the most susceptable to uncontrolled conduction because the electron flow in the vicinity of the mica is primarily influenced by only the last turn of the grid winding. Any variation in the position or distortion of this last turn, due to any one of a number of reasons, will have an adverse effect on the tube cutoff characteristics.

The bridge of the "Framelok" grid also shown in Fig. 8 is a part of the solid structure that rests directly against the surface of the mica. It prevents the grids from becoming jammed against the mica in the mounting operation and automatically positions and aligns the grid laterals of the control grid with those of the screen grid.

Electrically, the bridge can be thought of as an extension of the control grid that contains an increased number of turns. The control grid is then effective over the entire length of the grid structure and provides better control over the plate current in the presence of high plate pulse voltages.

It is the more uniform consistent cutoff characteristic and closer tolerances that can be attained in production with the "Framelok" tube that assure good high-voltage operation in horizontal-deflection circuits.

Horizontal Deflection Evaluation

Both the 6FH6 and 6DQ6A were examined in the 110° deflection circuit shown in Fig. 9. This circuit is typical of those presently used in most 110° television receivers. Data recorded at a screen voltage of 150 volts reveals that the low screen current, which is characteristic of the 6FH6, resulted in a screen dissipation of less than 1.0 watt. This compares quite favorably with the 1.45-watt dissipation of the 6DQ6A. The low screen current also became evident as a lower cathode current, and, even though conservatively operated, both the width of scan and high voltage were slightly better than that of the 6DQ6A. The reduced grid drive required by the 6FH6 reflects the good high-voltage cut-off characteristics of the "Framelok" tube.

No doubt the 6FH6 will be also considered in terms of the increased scan and high voltage that might be realized in those deflection circuits presently using the 6DQ6A. At the higher screen voltage and increased grid drive shown in Fig. 9, the 6FH6 gave 1.0 KV more high voltage and 0.4 inches more scan than the 6DQ6A. The screen dissipation at 1.35 watts was quite reasonable and did not exceed that of the 6DQ6A.

It should be noted that the low screen current of the 6FH6 will also become evident as a higher screen voltage in most deflection circuits. And, if optimum circuit performance is to be realized, it may be necessary to increase the grid drive beyond that presently required with the 6DQ6A. A few trial-and-error adjustments should readily determine the correct grid drive for a given screen voltage.

These comparisons illustrate that the low screen dissipation that can be expected with the 6FH6 permits the circuit design engineer a wider latitude in selecting those operating conditions that give the desired set performance.

Conclusion

This paper has dealt with some of the tube design factors that influence the life expectancy and performance of the horizontal deflection amplifier.

It was also shown how the tube characteristics, the unique "Framelok" structure, and the closer tolerances that can be attained in the manufacture of the 6FH6, have resulted in a deflection tube type that can more ably contend with the stringent requirements of horizontal deflection service.

In addition to the horizontal deflection application described in this paper, the many desirable features of the "Framelok" grid structure could be utilized to good advantage in the design of vertical deflection, audio and video amplifier tube types.

CIRCUIT	PERCENT FAILURE BY CIRCUIT		
	1954-'55	1955-'56	1956-'57
HORIZONTAL AMP	25	34	17
VERTICAL AMP	25	29	16
DAMPER	33	17	9
VHF CASCODE AMP	22	18	7

FIG I - PERCENT TUBE FAILURES BY CIRCUIT — RECEIVER LIFE TEST

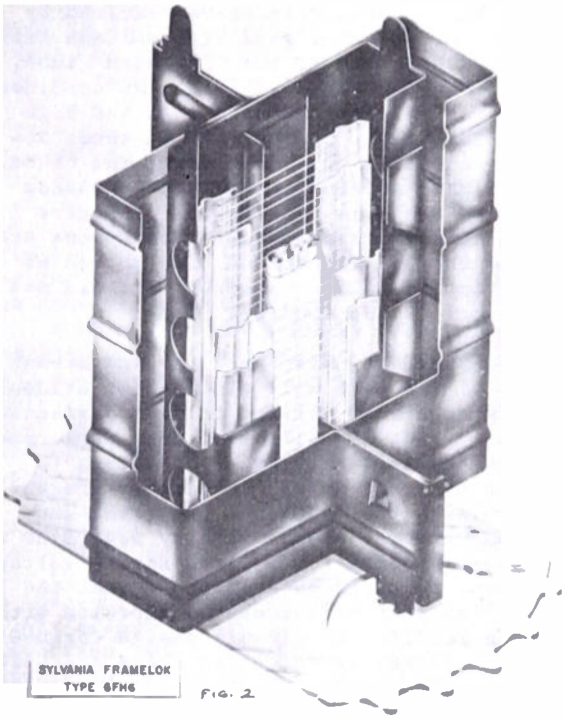


FIG. 2

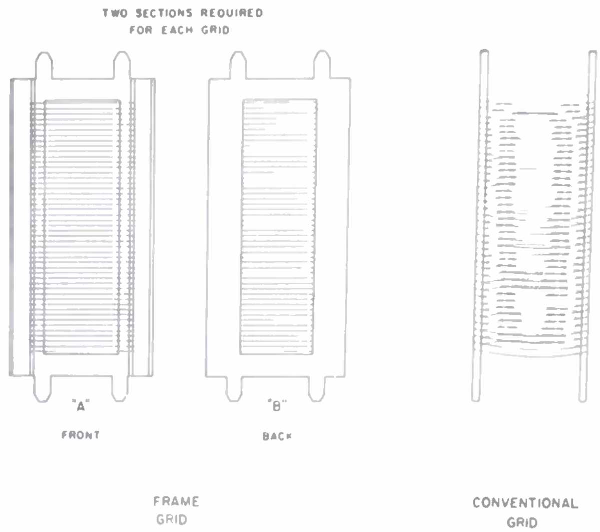


FIG 3– COMPARISON OF SCREEN GRIDS

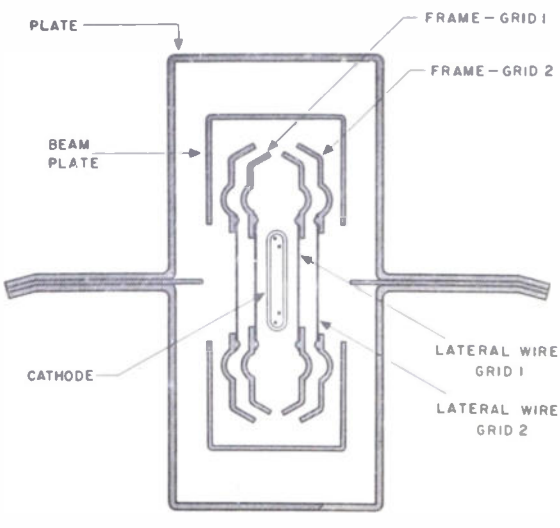


FIG. 4
CROSS SECTION OF MOUNT
EMPLOYING FRAME GRIDS

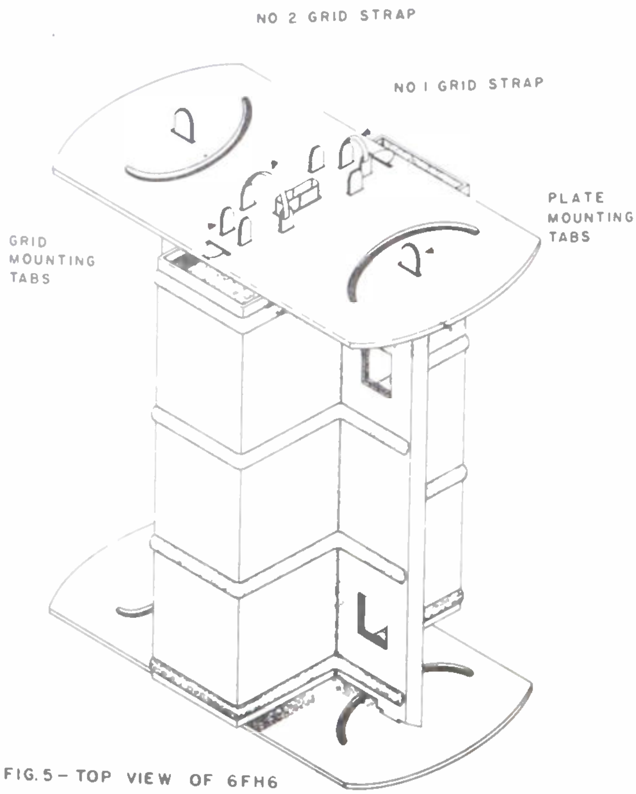
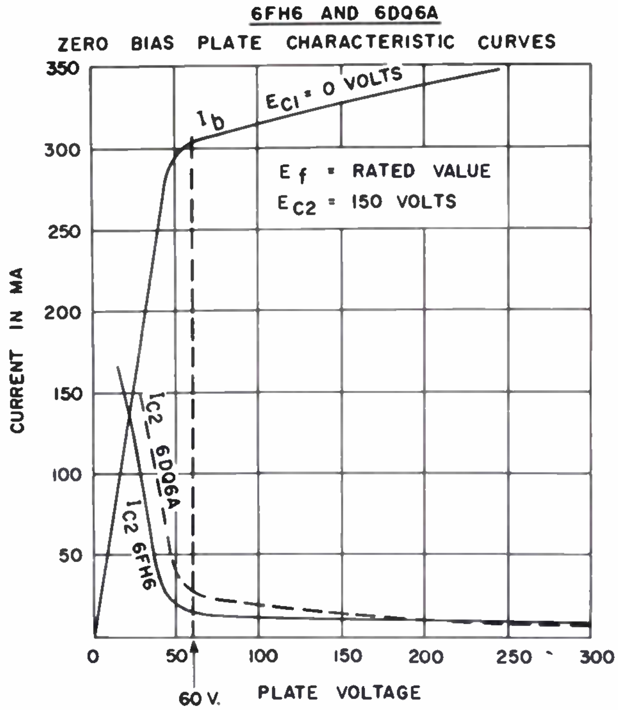


FIG. 5 – TOP VIEW OF 6FH6
SHOWING RELATION OF MOUNTING TABS AND MICA



	I_b	I_{C2}	PLATE TO SCREEN CURRENT RATIO
6DQ6A	300	27	11:1
6FH6	300	15	20:1

FIG. 6

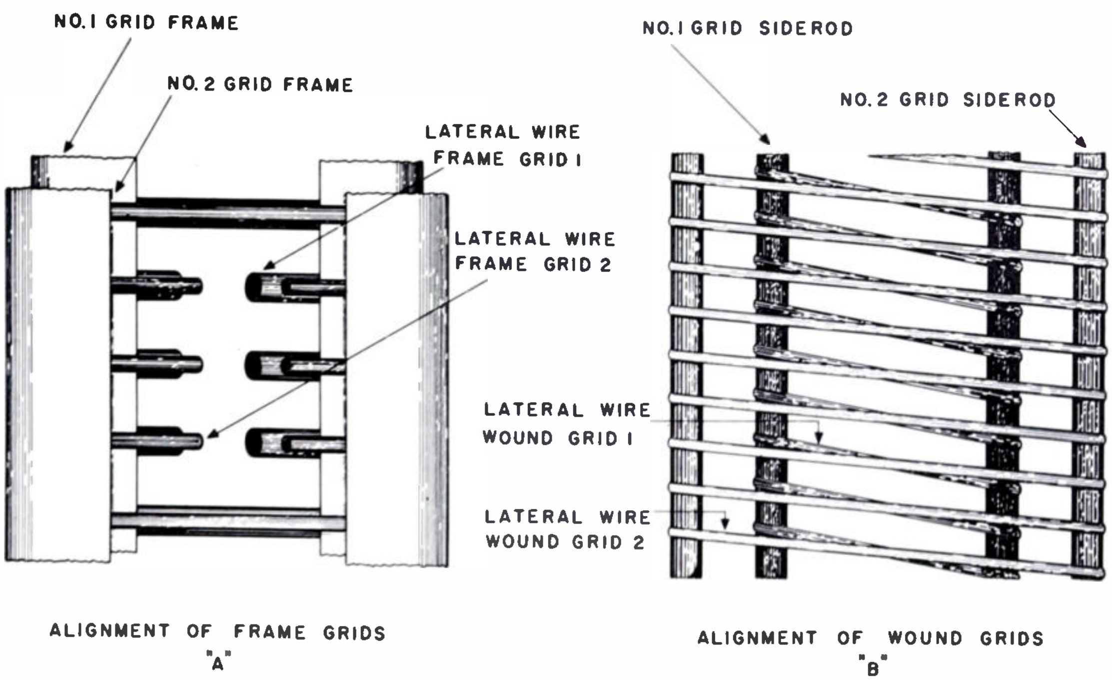


FIG. 7 – COMPARISON OF CONVENTIONAL AND FRAME GRID ALIGNMENT

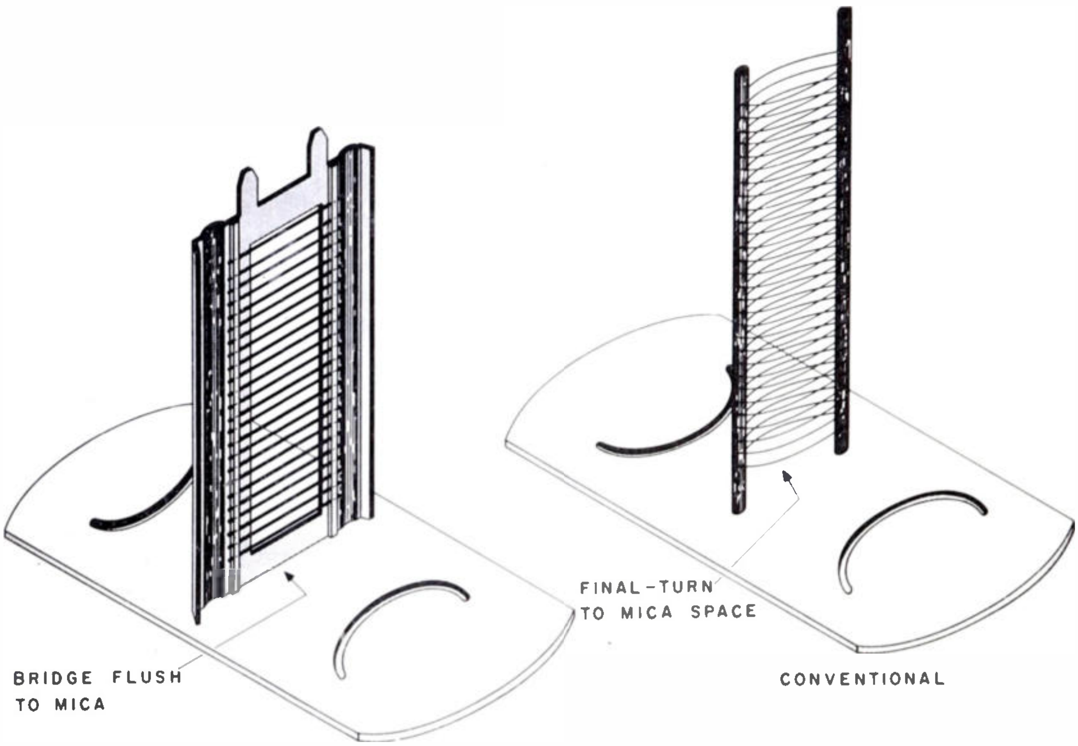
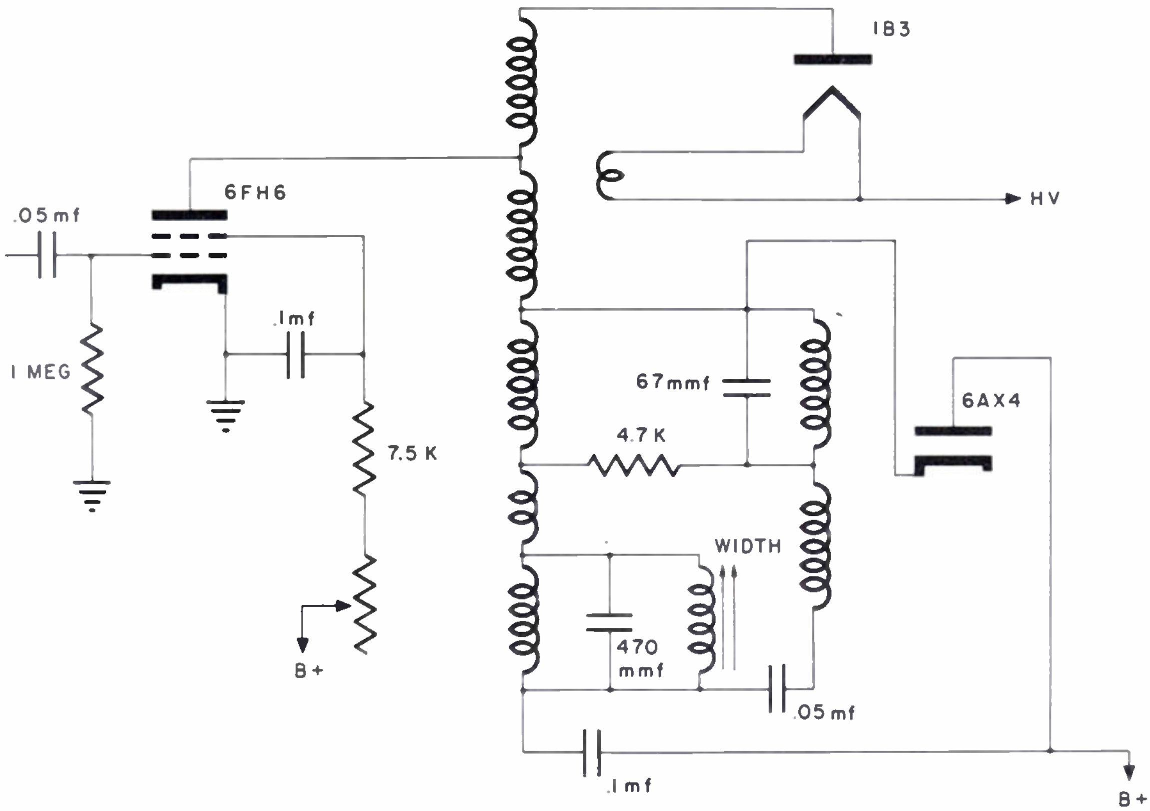


FIG. 8 – GRID TO MICA RELATION

110° DEFLECTION CURCUIT



110° DELECTION DATA

	B+	I_K	I_{C2}	E_{C2}	P_{C2}	DRIVE	HV 120 μa	HV 0 μa	SCAN
6DQ6A	250V	128 ma	9.7 ma	150 V	1.45 W	148 V	15.0 KV	15.9 KV	FULL
6FH6	250V	123 ma	6.5 ma	150 V	0.98 W	134 V	15.2 KV	16.1 KV	+0.2"
6FH6	250V	134 ma	8.3 ma	162 V	1.35 W	142 V	16.0 KV	17.0 KV	+0.4"

FIG. 9

A.G.C. DESIGN CONSIDERATIONS FOR TELEVISION RECEIVERS

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Summary

The type or form of automatic gain control used in a modern television receiver is the result of many factors and requirements. It is the purpose of this paper to examine these factors and relate them to the circuits which have been developed.

The signal to be controlled must certainly be listed first as a factor which determines the form of the automatic gain control system. There are however many other factors and considerations which also determine the form of the final system. Among these are the cost objectives to be met and the environment in the receiver itself. The tuner and the Intermediate Frequency Amplifier, as well as the design of the video amplifier and sync separator all have an influence on the design of the Automatic Gain Control system.

Examination of Design Requirements

For the purposes of this paper, we will define the Automatic Gain Control system of a television receiver to be any element of the television receiver which is part of the closed loop system whose function it is to maintain a constant signal level at some point in the receiver. This definition obviously includes elements, such as the tuner and I.F. amplifier, which have primary functions not connected with gain control. This is deliberate and their inclusion will now be justified. Perhaps the most logical method and the one to be used in this paper is to follow the control loop from the tuner through the receiver back to the tuner.

Tuner

The RF stage is the first element to be considered. If a pentode RF stage is to be used, the voltage supplied by the A.G.C. circuit can be the same magnitude as that supplied to the I.F. amplifier. It can also increase in the same manner since nothing is gained by delaying the application of bias as far as receiver noise figure is concerned. The use of a dual triode in a cascode circuit as a RF amplifier however imposes new requirements. To make best use of the available noise figure of such a stage, the application of control voltage must be postponed until just below the value of input signal at which the following I.F. amplifier starts to overload. This is known as delayed A.G.C. and is accomplished by injecting a positive voltage into the negative control voltage supplied to the RF stage. Since the cascode circuit requires more bias than the pentode circuit and we are now by the use of delay subtracting a fixed positive voltage from our control voltage supply; we must have a source of larger magnitude than needed for the pentode tuner or for the pentodes used in the I.F. amplifier. In order to prevent the grid of the RF amplifier from going positive and drawing grid current, it is necessary to provide a diode to be used as a clamp on the point at which the delay voltage is injected. This is usually done by using a tube complement in the audio section which includes an extra diode in one of the bottles. It is also possible to use the diode fromed by the grid and cathode of the RF stage as the clamp. In this case the clamp point will vary with RF tubes and performance will not be consistent from set to set.

At high signal levels, in the neighborhood of 100,000 micro volts and higher the cross modulation characteristics of the tuner affect the A.G.C. circuit design. If the RF stage is not completely neutralized it is possible to completely cut off the RF stage since feed through around the tube by means of distributed capacity will provide enough signal to supply the requirements of the rest of the receiver. A tuner which contains a well neutralized RF stage however must not be biased to cut off since the signal will disappear. The author well remembers spending several hectic days trying to find the fault in a design when all that had happened was that the tuner design group had supplied a more neutralized tuner than previously used.

I.F. Amplifier

The next section of the receiver to be considered which affects the A.G.C. system design is the Intermediate Frequency Amplifier. The obvious requirement in this section is to control the gain of this amplifier in such a manner that the detected output remains relatively constant over the range of signals involved. It is also important that the signal not be distorted by too high levels in this amplifier. This appears in the output as sync crushing and sound buzz.

Under high noise conditions, the impedance of the IF A.G.C. source becomes important. A zero impedance source would be ideal, because any grid current in the IF tubes caused by large amplitude noise spikes would not add to the A.G.C. voltage and produce white dots or bands in the display. It is of course impossible to have a zero impedance source but it is desirable that the impedance be kept as low as possible.

It was realized sometime ago that the receiver sensitivity could be increased by decreasing the bandwidth under low input signal conditions. In fact not only is the sensitivity increased but the apparent noise figure of the receiver is also improved by the elimination of high frequency noise components. This is accomplished by shifting a pole in the IF amplifier under the carrier giving a form of exalted carrier reception. The sound sensitivity may also be increased in the same manner. This pole shifting is performed by changing the input admittance of the stage concerned as a function of the grid bias voltage, i.e., the A.G.C. voltage. This requires that the A.G.C. voltage not only control the signal level but also it must change in a manner which causes the desired pole shifting to occur.

Video Amplifier

The next element in the A.G.C. loop with which we are concerned is the video amplifier. The matter which first claims our attention is the gain of the video amplifier or the first video amplifier if two or more stages of video amplification are used. Since the video voltage available to the cathode ray tube is always specified, the gain of the video system automatically sets our second detector level which is a level controlled by the A.G.C. system. Concurrently the video gain determines the signal available to drive the device which acts as the source of A.G.C. voltage. The gain distribution in a multiple stage video amplifier is usually such as to require a split load in the plate circuit of the first video stage. This consists of a resistor bypassed with a small capacitor as a sync load, providing typically 30 to 40 Volts of signal to the A.G.C. source, and a video load providing 15 to 20 Volts drive for the second video.

The subject of the contrast control also comes under this discussion of the video amplifier. In the past it has been common practice in single stage video amplifiers to place the contrast control in the plate circuit. A compensated potentiometer was used, placed across the load resistor so that it did not affect the gain to the A.G.C. source. If the contrast control is located in the video cathode circuit, it becomes a control on the A.G.C. gain as well as the contrast. This leads to a situation in which the contrast control lowers the video gain, the A.G.C. source interprets this as a decrease in signal strength and attempts to raise the detector level to restore the original signal level at its input terminals. It is possible to compensate for this effect to a degree but the variation caused by contrast control is limited to approximately three to one.

The choice of whether or not to dc couple the second detector signal into the grid of the video amplifier is directly a function of the type of A.G.C. system used. This, as well as the effect of the grid base of the video amplifier, will be discussed later.

Filter

We will omit for now the actual source of A.G.C. voltage. The final element in the loop is the filter used to remove signal frequency components from the A.G.C. voltage. The requirements on this section are common to any type of A.G.C. system. The filter must remove the signal components and yet not have such a slow response that certain types of signal disturbances become objectionable. Airplane flutter is an example of such a disturbance. It must also be realized that the A.G.C. system is a closed loop feedback system and as such is subject to breaking into oscillations or motorboating, if the conditions governing this oscillatory phase are met. This condition usually occurs when switching on to a station or with abrupt changes in picture content. The resistance values in the filter are determined by the magnitude of A.G.C. voltage desired at each control point (the tuner and the IF stages) and by the current available to develop this voltage. This leaves only the capacitor values in the filter as elements to be manipulated to avoid oscillations.

The factors discussed above are all subject to engineering decisions. There are others; including cost objectives, customer operation, performance levels and the received signal, which are not under the engineers control. They do affect the design of the A.G.C. system however and their effect must be evaluated.

reached. To prevent the video amplifier from cutting off, it must be AC coupled from the detector and this is a requirement for the averaging type of A.G.C.

Gated A.G.C.

The second type of A.G.C. system consists of a gated DC amplifier. The signal at the second detector is amplified by the video stage and applied to the grid of the A.G.C. tube. The DC value of the signal during sync interval is an accurate measurement of the signal amplitude being received. This signal is direct coupled to the video amplifier and the plate of the video amplifier is direct coupled to the grid of the A.G.C. tube. This involves operating the A.G.C. tube above ground and the heater to cathode voltage ratings must not be neglected. The plate voltage of the A.G.C. tube consists of a large amplitude (typically five (5) to six (6) hundred volts) pulse taken from the horizontal output transformer and AC coupled to the plate. If the plate resistor is returned to ground, the voltage developed by the tube current will be negative with respect to ground and is used as the A.G.C. voltage.

This system is more expensive than the averaging system due to the extra tube but has many advantages. It is a high gain system. If a pentode is used as the A.G.C. tube, it is possible to develop seventy (70) to eighty (80) volts of A.G.C. voltage. Since we are using milliamperes of plate current instead of micro amperes of plate current, the same voltage can be developed across a much lower impedance.

The noise performance of the gated system is usually superior to that of the averaging system. It is immune to all noise except that occurring during gating time. The high gain of the system can be used to give a practically constant detector level. This allows the operating point of the direct coupled video amplifier to be chosen such that noise pulses are limited just above sync tips. This minimizes the effect of noise which occurs during sync interval.

Since this system measures the amplitude of sync tips to develop the A.G.C. voltage, it is not affected by picture content. Its performance under high input signal conditions with respect to blocking will not be a function of picture content but only of signal amplitude. The well designed gated A.G.C. system does not require a control or switch of the complexity used with the averaging system. All that is required is a simple switch to insert an attenuator in the input line under conditions of very high signal inputs.

Many forms of these A.G.C. systems exist since many companies and engineers are active in television receiver design. They are, however, only variations of these basic systems and once classified will be found to have the advantages and disadvantages of the systems discussed.

Conclusion

This paper has examined the requirements and considerations which affect the design of an A.G.C. system for a modern television receiver. The principal types of systems which are in current useage have been described and the manner in which they meet the requirements described. Experts in A.G.C. work will recognize the fact that subleties of A.G.C. design have been avoided. It was impossible not to do this in a paper of this type and still include the important things.

It would be improper to conclude without expressing the author's indebtedness to his fellow workers in the Television Design Group at Philco Corporation and particularly to A. Peers Montgomery, whose knowledge of A.G.C. work is encyclopedic.